Founded 1925
Incorporated
by Royal Charter 1961
To promote the advancement of radio, electronics and kindred subjects by the exchange of information in these branches of engineering

Volume 47 No. 10 October 1977

The Radio and Electronic Engineer

The Journal of the Institution of Electronic and Radio Engineers

99 GOWER STREET

THE permanent staff of the Institution of Electronic and Radio Engineers are now installed in new offices just a few hundred yards from the building in Bedford Square which has housed the Institution since 1942. Situated at 99 Gower Street, the new headquarters, which provides a more compact and cost-effective base for the Institution's work, is a fully restored and modernized Georgian house with a new office block at the rear. In addition to suitably appointed offices for the management and administration of the Institution the building accommodates the Library and Information Department and meeting rooms for the Institution's Council and Committees.

The eighteenth-century house on which the Institution's new headquarters is founded is of similar period to the Bedford Square premises, and is part of a terrace on the west side of what was first named Upper Gower Street. The land was originally a field on the Southampton estate, and the immediate area of this part of Bloomsbury was first built on between 1780 and 1790. For much of the time since it was first leased in 1790, No. 99 has been occupied by professional men, especially after the building of University College London and University College Hospital. The College now extends over most of the east side of 'Upper' Gower Street, the latest buildings being immediately opposite No. 99. In 1926 the political and literary weekly *The Spectator* took over the house as offices and remained there for nearly fifty years until the property was taken in hand for restoration and extension to its present form.

The house has four main floors and a basement which provide general office and committee accommodation well matched to the Institution's current needs; and the new 3-storey open-plan annexe provides compact housing for the Library and for all the staff concerned with membership administration and the Institution's publishing work. The ground floor of No. 99 has externally an unusual arcaded front with three arches, and the doorway is flanked by small Doric columns. The principal rooms on ground and first floor are respectively the Committee Room and the Council Chamber.

As was explained in the recent Annual Report, the Institution's housing problem has been a matter of concern to the Council for some considerable time. This move to 99 Gower Street reflects that concern and has been made generally to reduce and stabilize the Institution's housing costs, which have been increasing alarmingly over the past three years, but specifically to avoid a further major increase in rent due under the terms of the Bedford Square Lease in April 1978. The Institution has taken a 25-year Lease on the new building and during negotiations for this Lease a suitable understanding has been reached to cover the Institution's possible future wish to purchase the property when the new Housing Fund is sufficiently well established.

The headquarters move of any active and long established organization is a major venture which like marriage, should not 'be taken in hand unadvisedly lightly or wantonly'. The IERE move cannot be faulted in these terms: the matter has been the subject of much careful debate and the final decision to abandon 8/9 Bedford Square was not taken until it became quite clear that the many problems associated with those premises could be contained no longer. It is hoped, therefore, that the Institution's association with 99 Gower Street will be a long-term success; and since the building is so well matched to the modern needs of the Institution there is justification for confidence on this score.

S.M.D.

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Professor Biswaranjan Nag is on the teaching staff of the Institute of Radio Physics and Electronics, Calcutta University, and at present he is working as a Jawaharlal Nehru Fellow. He received his B.Sc. degree in 1951 and M.Sc. (Tech.) in 1954 from Calcutta University, the M.S. degree from Wisconsin University in 1960, and Ph.D. and D.Sc. degrees in 1961 and 1970 respectively from Calcutta University. Dr Nag has worked on electronic computers, non-linear phenomena in electronic oscillators, wave propagation in ionized media and microwave measurement of semiconductor properties. His present research interests are concerned with microwaves and hot-electron transport in polar semiconductor superlattices. He was awarded the Shanti Swarup Bhatnagar Prize for 1975 for his contributions in the field of electron transport in semiconductors. He is the author of 130 research papers and a book; six of his papers have been published by the Institution and in 1962 Dr Nag received the Sir J. C. Bose Premium for a paper on a two-state circuit using inductively-coupled Colpitts oscillators.

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Dennis Ralphs (Member 1960, Associate 1945) is head of the Research and Development section of the Foreign and Commonwealth Office Communications Engineering Department. He was responsible for the design of the 'Piccolo' 32-tone telegraphy system used on the FCO h.f. radio network. A fuller biography appeared in the December 1976 Journal on the occasion of a previous (joint) paper on a h.f. propagation simulator, which has gained him the Marconi Premium.

George Wilson completed an Electrical Apprenticeship with the National Coal Board in 1961. He obtained a B.Sc. (Class I) and Ph.D. degrees in Electrical Engineering in 1970 and 1973 respectively from Sunderland Polytechnic, and was Research Assistant in the Electrical Engineering Department for several years. In 1974 he was appointed to a lectureship in the Department of Electrical Engineering at the James Cook University of North Queensland. Dr Wilson's research interests are in RC active networks, and he has contributed to date some fifteen papers to British, American and German journals.

* See also page 454

UDC 621.394.4: 621.371.029.55

The application of m.f.s.k. techniques to h.f. telegraphy

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SUMMARY

The paper briefly describes the history and the principles of m.f.s.k. signalling as applied to h.f. telegraphy. It quotes mathematical relationships for the performance of such systems, and derives equations for the occupied bandwidth. These relationships, together with the results of measurements carried out on different experimental m.f.s.k. systems, and practical limitations suggested by experience, are used to suggest an approach to the selection of optimum parameters.

Systems suitable for different telegraphy and telemetry applications are derived and discussed. Practical problems of synchronization and source stability are analysed, and methods of error coding for ARQ and FEC systems are described.

It is concluded that for data links in the range of 20 to 200 bits per second, the use of m.f.s.k. techniques can give an improvement in performance equivalent to about 10 dB in signal-to-noise ratio, or a reduction in error rate of the order of 10/1 in the raw data (or more than 100/1 if an error coding system is used).

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1 Introduction

The technique of signalling by making and breaking a direct current in a wire between the send and receiving station was one which evolved at a very early stage in the history of point-to-point communication, and the simplicity, reliability and sensitivity of simple on/off devices were vital factors in the crude technological environment of the day. The later developments of tone signalling over lines and radio frequency signalling inherited and perpetuated these techniques since the overriding requirement for simplicity of terminal equipment still existed. It is not surprising therefore that the world-wide network of international radio telegraphy links which was built up in the first half of this century should be based entirely on binary concepts, and, having once been firmly established in this mode, that economic considerations and the natural conservatism of a large international industry should discourage the serious investigation of other possibilities. Advances in the theory of communication in the 1950s included investigations into the possibilities of using multi-state systems[1–3] but although it was established that such techniques were capable of producing a considerable improvement in accuracy over an equivalent binary system, they were never seriously introduced into the h.f. sphere and indeed at that time the electronic techniques required to implement complex circuits to the required degree of reliability were not yet available. In the early 1950s Mr. H. K. Robin, then Chief Engineer of the Diplomatic Wireless Service (which later became the Communications Engineering Department of the Foreign and Commonwealth Office) experimented with the use of multi-tone systems for h.f. telegraphy and proved that such a system could indeed give a very marked reduction in error rate. This led to the design and development of the 'Piccolo' 32-tone multiple frequency shift keying system,[4] which was introduced experimentally in 1962 and demonstrated publicly in 1963.

The present-day FCO network of about 60 long-range, low-power point-to-point h.f. telegraphy links operates almost entirely on Piccolo Mk. 2 and Mk. 3[5,6] (about 220 units being in service), and the success of the technique is such that in 1970 planning began for a 'second generation' equipment.

The need for a redesign arose primarily for three reasons:

(*a*) The increased cost of production and maintenance due to component obsolescence.

(b) Changes in requirements of the network, requiring additional facilities which had been provided temporarily by *ad hoc* additional units.

(c) Improvements which it was known could be made to the system in the light of experience.

However, it was appreciated that the original parameters had been derived by a rather empirical approach, and that the theoretical analysis of m.f.s.k. principles available since then indicated that it would be worth while re-assessing the basic parameters. The possibilities of commercial exploitation, the growing application of m.f.s.k. techniques in parallel fields (such as telemetry from unmanned meteorology and marine research buoys,[7]) and the publication of a CCIR Report[8] on such techniques, suggested that a wider and more general analysis would be of interest to other communication organizations in the field. It is this analysis, together with practical comments and advice based on fifteen years of experience that forms the basis of this paper.

In view of the proved success of the existing Piccolo equipment and the close approximation of its measured performance to that of an ideal non-coherent correlation detector,[15] consideration will be limited to m.f.s.k. systems of the same type, in which the transmitted signal consists of a sequence of single-tone elements of constant duration, selected from a finite number of possible frequencies. The transition between elements is assumed to be instantaneous and phase-continuous. This signal is transmitted as an s.s.b. modulation with suppressed carrier so that the radiated signal consists of a single sine wave which changes frequency suddenly by discrete steps at exact multiples of a fixed time interval. This is received in a synthesized s.s.b. receiver and converted to an audio tone of varying frequency, which is applied to a bank of 'integrate-and-dump' matched filters, one for each frequency. At the end of each tone interval a comparison circuit examines the outputs of all the filters and indicates which filter contains the greatest amount of energy. This choice constitutes the output of the system. For dual diversity operation a separate receiver and signal chain is provided ending in a similar bank of matched filters and the comparator selects the highest output of the total quantity of filters. This constitutes a 'channel selection' diversity system operating over each individual character interval.

The principle governing the design of a matched filter for a burst of tone of constant amplitude and known length, and the use of orthogonal spacing, whereby the frequency interval between two adjacent tone frequencies is equal to the inverse of the integration time of the matched filter, are treated in detail in the literature and will not be repeated here. As a matter of practical and theoretical convenience the integration time of the filter may be assumed to be equal to the length of the transmitted element. Mathematical relationships for error-rates for such systems are also derived in many textbooks and some are quoted in Section 12 together with other calculations not essential to the text.

2 Basic Principles

Consider a system which is required to convey H bits per word, and let each word be transmitted by a sequential series of N elements of tone each of duration T seconds, each element being freely selected from a series of M available frequencies. From fundamental communication theory, if a single choice is made from M different alternatives, the amount of information Q in that choice is:

$$Q = \log_2 M \text{ bits.} \tag{1}$$

Therefore

$$M = 2^Q \geqslant 2^{H/N}. \tag{2}$$

In principle N can have any value from 1 to H and therefore, taking $H=10$ bits as an example, a table can be constructed as Table 1.

Table 1. Theoretically possible systems for transmitting a 10 bit word.

No. of elements N	H/N	$2^{H/N}$	Minimum number of frequencies M
10	1	2·00	2
9	1·11	2·16	3
8	1·25	2·38	3
7	1·43	2·69	3
6	1·66	3·17	4
5	2	4·00	4
4	2·50	5·66	6
3	3·33	10·08	10†
2	5	32·00	32
1	10	1024	1024

† 10 bits = 1024 possibilities: 3 elements selected from 10 tones = 1000 possibilities. The two are sufficiently equivalent for present purposes.

The probability of error for an orthogonal m.f.s.k. system in the presence of Gaussian noise can be calculated (see Sect 12.2) and Fig. 1 shows the computed word error probability P_W against the normalized signal-to-noise ratio, W, where W=(signal energy per bit)/(noise power per Hz).

The curve for $M=2$ represents the performance of an 'ideal' non-coherent binary f.s.k. system recommended as a reference system by CCIR.[9] The curves confirm the conclusion of Slepian[2] and others that the error rate decreases as M is increased, but the improvement for very high values of M is small. Note that the curves are normalized to the same data rate.

Similar curves for a Rayleigh fading signal in Gaussian noise, received in single-path and dual diversity, are given in Fig. 2 and it can be seen that the differences (in dB) between the performance of any two systems are of similar order in the three conditions considered, although of course the change in error rate per dB difference is very much greater for a non-fading signal than for a fading signal however received.

For a constant data rate, the element period T must be inversely proportional to the number of elements N per word, and since the frequency spacing is the inverse of the element period, the differences in signal structure may be illustrated as in Fig. 3, which shows four different systems for signalling at 50 bits/second.

The 'necessary bandwidth'[10] of m.f.s.k. signals of this type extends at least 1 tone-interval and preferably 2 tone-

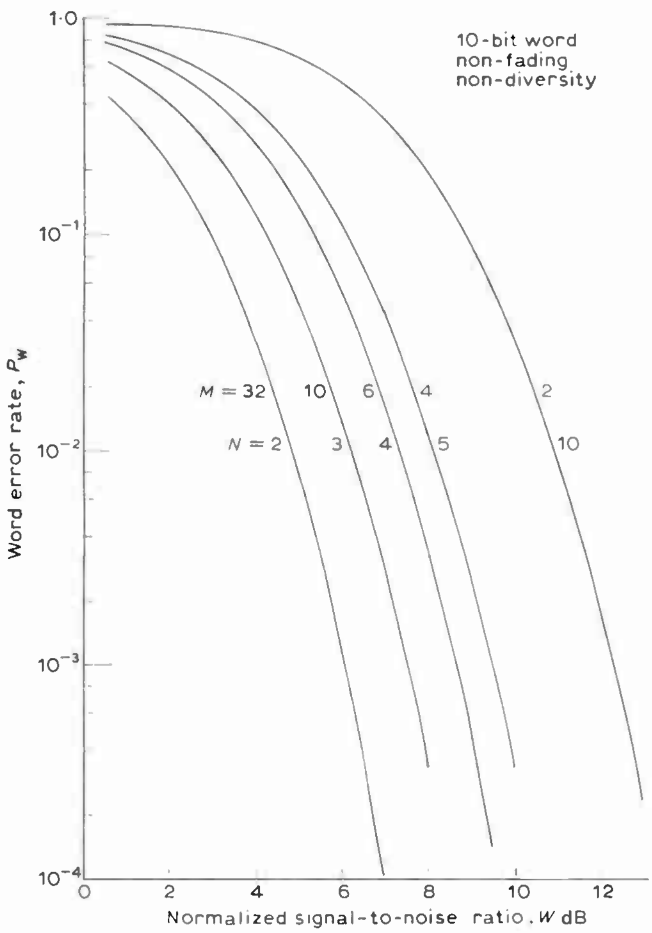

Fig. 1. Word error rate on non-fading signal in noise.

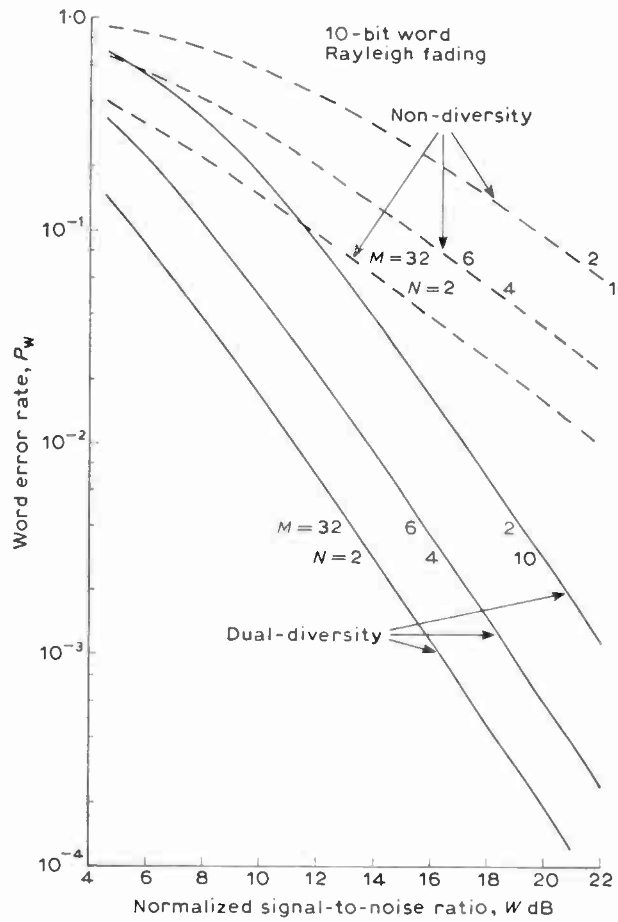

Fig. 2. Word error rate on Rayleigh fading signal in noise.

intervals beyond the extreme tone frequencies. If the 'occupied bandwidth', defined as the bandwidth containing 99% of the total radiated power, is assumed to extend a total of G tone-intervals beyond the extreme tones, it can be shown that the 'normalized occupied bandwidth' is given by

$$B_0 = (M-1+G)/\log_2 M \text{ Hz bit}^{-1}\text{ s}^{-1} \text{ or cycles/bit} \quad (3)$$

This equation is plotted in Fig. 4 for various values of G. The spectrum can be calculated for the worst case, in which the signal is keyed on alternate elements between the extreme frequencies,[11] and on this assumption the occupied bandwidth of a 6-tone system extends about 2 tone-intervals beyond the extreme tones at either end and that of a 32-tone system about $3\frac{1}{2}$ tone-intervals.

Assuming then a practical value of $G=4$, it can be seen that B_0 is at a minimum for $M \simeq 6$, and the minimum normalized bandwidth is about 3·5, giving a maximum specific information density of about 0·3 bits/cycle. This is of the same order as that obtained for comparable binary systems.[12]

The basic philosophy of the choice of an optimum value for M can now be seen by reference to Figs. 1, 2 and 3. As the number of available tones is increased above two the error performance is considerably improved and the bandwidth falls, reaching a minimum at about 6–8 tones. Further increase in the number of tones will continue to improve the performance but at the expense of increasing bandwidth.

3 Choice of Parameters

Although in principle any value of M satisfying equation (2) can be used, if unnecessary redundancy and coding inefficiency is to be avoided the choice is limited to those parameters for which equation (2) approaches equality. (Note that this is in contrast with binary techniques, where the design of the modulation and demodulation system—excluding error coding—can be made without reference to the signal structure of the information source.)

From Table 1 it can be seen that for a 10-bit word the only systems worthy of consideration are: $M=6$, $N=4$; $M=10$, $N=3$; $M=32$, $N=2$. $M=10$ can cause coding difficulties in some cases, and the other two have the advantage of being also directly applicable to the ITA-2 (5-bit) alphabet. Similar consideration of a 7-bit word (e.g. ITA-5 or ASCII) indicates that the choice is restricted to: $M=5$, $N=3$ (with possible coding difficulties); $M=6$, $N=3$; or $M=12$, $N=2$.

The other major parameter to be considered is the element duration T, and here again there are practical

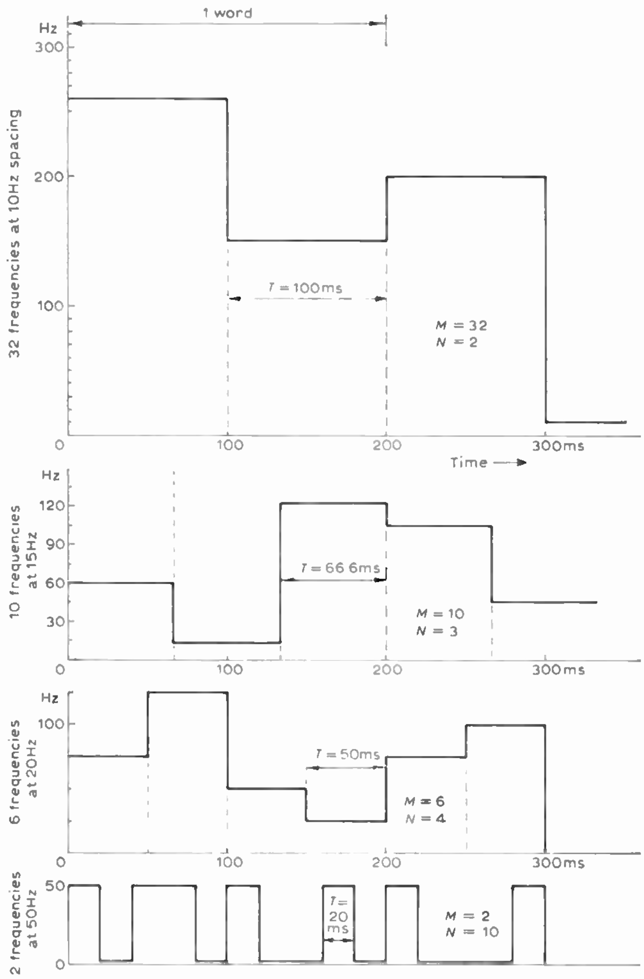

Fig. 3. Four 50 bit/s systems for a 10-bit word.

limitations. The highly destructive effects of multipath propagation on conventional h.f. telegraphy are well known,[12] systems with an element length of the order of 10 to 20 milliseconds tending to collapse completely with path time delays in excess of 3 or 4 milliseconds. The occurrence of multipath propagation with delays considerably greater than these values has tended to be ignored in the literature since they are primarily a phenomenon of short-range communication (100 to 1000 km) but in the opinion of the author it would be unwise to design an h.f. system for poor signal applications with an element length less than 20 milliseconds while a system which is intended primarily for short-range application should use an even longer element length, preferably up to 100 milliseconds. Further increase in element length restricts the immunity of the system to Doppler shifts and makes increasing demands on the absolute stability of the r.f. frequency sources, which can cause difficulties in mobile or emergency applications.

If these limitations are accepted it can be seen that the optimum area of application of m.f.s.k. techniques to poor-signal h.f. telemetry or telegraphy is roughly bounded by the values of $M=6$ to 32, and $T=20$ to 100 ms, giving a range of available data rates of 25 to 250 bits/second per channel.

Table 2 gives the suggested parameters for four systems covering most of the requirements for low-speed h.f. telegraphy and telemetry. Frequency-division-multiplexing of channels is quite practicable and the rapid fall-off in the power spectrum outside the necessary bandwidth enables close packing of channels to be achieved.

Table 2. Suggested practical m.f.s.k. systems.

	A (Piccolo Mk. 3)	B	C1	C2
1 No. of tones	32 (+2)	6 (+0)	12 (+0)	10 (+2)
2 Element duration, ms	100	50	50	50
3 Tone spacing, Hz	10	20	20	20
4 Occupied bandwidth, Hz	370	180	300	300
5 Possible codes	ITA-2	ITA-2	ITA-5	Decimal Telemetry
6 Bandwidth occupancy, Hz $\text{bit}^{-1}\ \text{s}^{-1}$	6·4	3·6	4·3	4·5

Notes:
1 Additional tones in brackets are for synchronizing purposes. (See Sect. 5.)
4 Occupied bandwidth is calculated for 'worst case' conditions. It may be possible to use f.d.m. at about these spacings.
5 Output data rate = 10 characters/second (C2 = 20 digits/s). Note that any ITA-2 system can in theory operate on an ITA-5 alphabet with a 33% reduction in character rate, by recoding two 7-bit characters to three 5-bit characters.
6 Bandwidth occupancy = (occupied bandwidth/true information rate).

4 Experimental Evidence

As part of the development of a 'next-generation' system to replace the Piccolo Mk. 2/Mk. 3 (referred to in the Introduction), and to confirm the theoretical analysis, it was decided to compare in detail two of the systems in Table 2. System A has similar parameters to the existing Piccolos, while System B is a 6-tone system, with two elements per 5-bit character.

Experimental equipments to these standards were built and have been tested in the laboratory on an h.f. channel simulator[13] and on live radio circuits. The measured values for the 32-tone system on a non-fading signal are within ½ to 1 dB of its theoretical curve of Fig. 1 (confirming the findings of the BPO.[14] The 6-tone system performance is about 1 to 1½ dB from its theoretical curve. It is noteworthy that such f.s.k. systems as have been tested by the author have been 2–5 dB worse than the corresponding binary curve.

In the case of non-diversity reception of a Rayleigh fading signal in noise there is a similar degree of agreement with theory except that a residual error rate is obtained at high signal-to-noise ratios. The power spectrum of the radiated carrier is spread by the fading, causing the introduction of energy into adjacent frequency filters. Fast fading therefore destroys the orthogonal

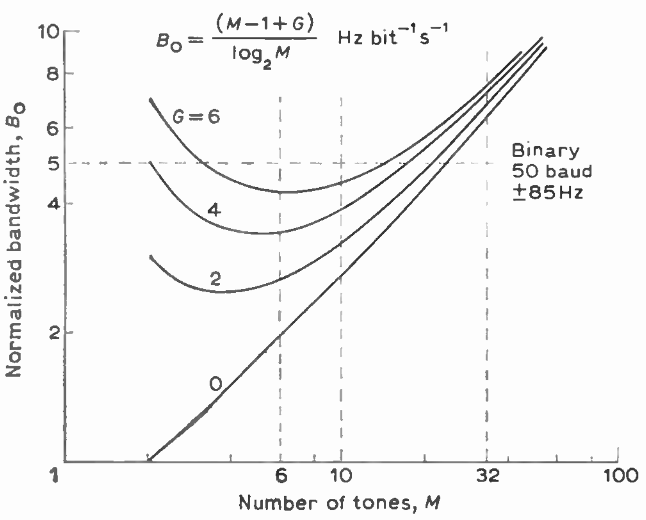

Fig. 4. Normalized occupied bandwidth.

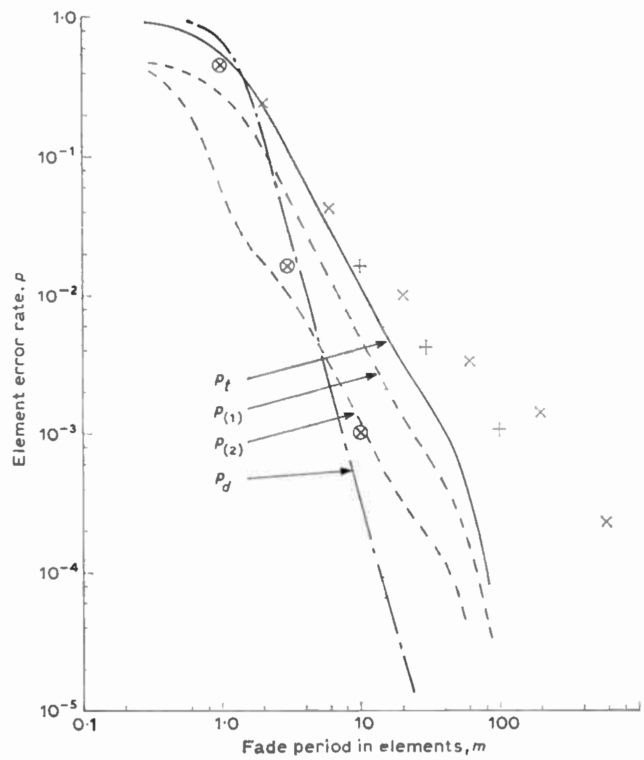

Fig. 5. Element error rate on fast fading signal with no noise. p_1 = probability of selecting one specific adjacent filter; p_2 = probability of selecting one specific filter two tones away; p_t = total probability of error; p_d = total probability of error with optimal selection dual diversity; × = measured points (100 ms system); + = measured points (50 ms system); ⊗ = measured points (100 ms system–dual diversity).

nature of the signal. A complete mathematical analysis of this phenomenon does not seem to be simple but analysis by arithmetical integration on the lines given in Section 12.4 produced the curves shown in Fig. 5 which show the theoretical element error rate on a fading signal with no noise, plotted against the rate of fade (expressed as the number of signalling elements in a fade cycle).

The curve p_1 shows the theoretical probability of selecting one specific adjacent filter in error, and p_2 that of selecting a filter two tone-spacings away. Twice the sum of these gives p_t, the total probability of error (neglecting band-edge effects). By a similar procedure, the curve p_d gives the theoretical total error probability in a dual-diversity mode using optimal channel selection. Measurements have been carried out on the simulator both on a 100 millisecond and a 50 millisecond system and these are shown in the diagram. It can be seen that agreement is very close for very fast fading but at slow fading the theoretical curves tend to be optimistic.

The discrepancy on slow fading between theory and measurement may be explained by observing that the phase of a Rayleigh fading signal varies slowly when the amplitude is high, and rapidly near to a null, when it will give a temporary broadening of the power spectrum and introduce more energy into an adjacent filter. There is therefore a negative correlation between the level in the wanted filter and that in an adjacent filter, causing the measured error rate to be higher than theory. With very rapid fading the two will tend to be uncorrelated and therefore comply with the assumptions.

Empirical 'straight line' approximations to the curves may be used to show that for the two systems under discussion the character error rate in fast fading will be very similar (the 6-tone system possibly having a slight advantage), and also that if the tone separation of any m.f.s.k system is doubled (thus doubling the bandwidth), the improvement in element error rate in fast-fading conditions will be about 4 to 1 for single-channel reception and more than 10 to 1 for dual diversity. The theoretical and measured error rates of the two systems under the various conditions are compared in Table 3.

Radio trials have been carried out totalling nearly 2000 hours of testing on eight or nine different paths between 100 and 14000 km in length, on frequencies between 4·5 and 25 MHz. The vast majority of these tests showed a small superiority for the 32-tone system of 1·3 to 1·8/1 in error rate, thus validating in general terms the conclusions of the laboratory analysis. Although the details of these trials cannot be discussed here, it is interesting to note that specific points, such as the greater superiority of a 32-tone over a 6-tone system for a shallow fading signal buried in noise were confirmed, and the only conditions under which the 6-tone system consistently showed to appreciable advantage was when its narrower bandwidth allowed it to work between interfering signals which caused deterioration in the wider bandwidth 32-tone system. The similarity in performance of the two systems, and the different advantages and disadvantages have led the author to form the opinion that for most long-range applications, the choice of the principal parameters (number of tones and element length) within the ranges indicated above would have negligible effect on the performance of an m.f.s.k. system. For short-range, low-power applications the superiority of the longer element in conditions of non-fading signals in noise and long multipath delays indicate a preference for the higher values of M.

Table 3. Summary of comparative performance of 6-tone and 32-tone systems.

Conditions		A†		dB‡	
		Calc.	Meas.	Calc.	Meas.
Non-fading.	Single aerial.	100	~50	2·5	3
Fading.	Single aerial.	2·3	2	3·5	3
Fading.	Dual diversity.	3	—	3	—
Fast fading.	Single aerial.	0·5–1	0·7–1	—	—
Multipath.	P.t.d. < 5 ms.	—	~1	—	—
Multipath.	P.t.d. > 5 ms.	—	> 1	—	—
Radio—Typical long range		—	1·3–1·8	—	—
Short-range (daytime)		—	~2·5	—	—

† $A = \text{ratio} \dfrac{\text{character errors on 6-tone system}}{\text{character errors on 32-tone system}}$

‡ dB = increase of signal/noise ratio in dB required on 6-tone system to give same error rate as 32-tone.

5 Synchronization and Stability Requirements

As indicated above, it is considered advisable that an m.f.s.k. signal should be conveyed by s.s.b. with suppressed carrier because this gives optimum performance and the ultimate in energy and bandwidth conservation. It has been suggested that the resulting high standard of short and long-term frequency stability required in the radio equipment is a major disadvantage of such a system but in fact it can be shown that the orders of stability required are well within the capabilities of good modern synthesized equipment, although m.f.s.k. systems are not recommended for use with obsolescent or low-performance radio receivers.

The effect of an error in any of the radio-frequency sources in the system is to displace the frequency of the received tone and it can be shown that an error of 10% of a tone interval will give a deterioration in performance equivalent to about 1 dB reduction in signal-to-noise ratio. Thus for a 100 millisecond element an absolute overall accuracy of better than 1 Hz is preferred (1 in 10^7 at 10 MHz).

The theoretical analysis leading to the curves of Figs. 1 and 2 is based on the assumption that the instant at which an incoming element begins is known at the demodulator. In practice this information must be derived from the incoming signal by some form of synchronization In the Piccolo Mk. 2/Mk. 3 design this information is conveyed by a small amplitude modulation of the signal at character rate. In the demodulator a phase-locked loop with a time constant of 10 to 15 seconds locks on to this modulation and the synchronization process is therefore continuous. Although this system has proved reasonably effective under most ionospheric conditions it has been found that it can introduce systematic errors in conditions of fast or selective fading and the use of a.m. introduces engineering complications. A new system of synchronization has been developed using frequency modulation of a synchronizing tone (or between two tones) at character rate or half character rate.

If the basic system is non-redundant (i.e. eqn. (2) is an equality) the synchronizing tone frequency must be additional to the data tone frequencies; for example, the effective bandwidth of the 32-tone system is extended a further one or two tone-intervals by the addition of the synchronizing signal. The 6-tone system can convey 36 different two-tone sequences and so one of the four 'spare' combinations can be used as synchronizing signal, and therefore no increase in bandwidth is required.

The synchronizing signal is detected in the demodulator by a double phase-locked-loop system which has a high immunity to ionospheric effects including fading rates of slower than 1 fade per second and Doppler shifts up to 3 or 4 Hz. The system will pull into synchronization on a clean signal within 5 or 6 seconds, leaving a residual timing error of less than one or two milliseconds, while on a marginally bad signal these figures are roughly doubled. The synchronizing signal is only transmitted when the system is idling, synchronism being maintained by the stability of the send and receive character rates during data transmission. In the system developed, these are derived from the same stable crystal source as is used to synthesize the radio frequencies.

If long periods of continuous data transmissions are expected, consideration must be given as to the length of time for which synchronism will remain adequately accurate. The effect of a timing error can again be interpreted as being equivalent to a deterioration in signal-to-noise ratio, and it can be shown that an error of 5% of an element length is equivalent to a reduction in input signal-to-noise of about 1 dB. A frequency error in the master clock source can therefore affect the error rate in two ways, one of which tends to be of predominant importance when operating at high frequencies, and the other which dominates on long periods of continuous data transmission.

6 Error Coding Techniques

It may be thought that error coding for a multi-level modulation system could be accomplished by conventional binary coding before the signal is assembled into multi-level form, decoding being similarly carried out on the final binary output stream. In fact such a system would be extremely ineffective, since a single element error in the multi-level system can cause corruption of a number of bits. For instance, in a 32-tone system almost half the element errors will cause corruption of an even number of bits and will not therefore be detected with a binary single parity check code. It would seem to be a fundamental principle that the error coding for an M-level channel should be carried out in modulo-M (or higher level) arithmetic. This necessarily involves rather more complex electronics but can lead to a considerable increase in efficiency.

Consider an m.f.s.k. signalling system using M tones in which each tone is allocated an arithmetical 'weight' of 1 to M. Input data are coded into blocks of $(B-1)$ elements, and to each block is added a check element selected so that the total, in modulo-M arithmetic, is zero. If $M=2$ (binary) this code will detect all blocks in which an odd number of elements are in error and will fail to detect all those where an even number of elements are in error. A similar procedure applied to a 32-level system will detect all blocks with a single element error and 97% of all multiple error cases. The curves of Fig. 6 (derived in Sect. 12.5) show the probability of an undetected error block achieved by a 11-element block (0·9 rate) parity check code for $M=2$, 10 and 32 and it can be seen that the improved performance of a multi-level system is maintained at high error densities. This type of code would be adequate for implementing an ARQ† system on an m.f.s.k. circuit.

It would seem to be theoretically practicable to implement in M-level form any binary FEC‡ code based on a number of parity checks, and this has been carried out in computer simulation for decimal versions ($M=10$) of a

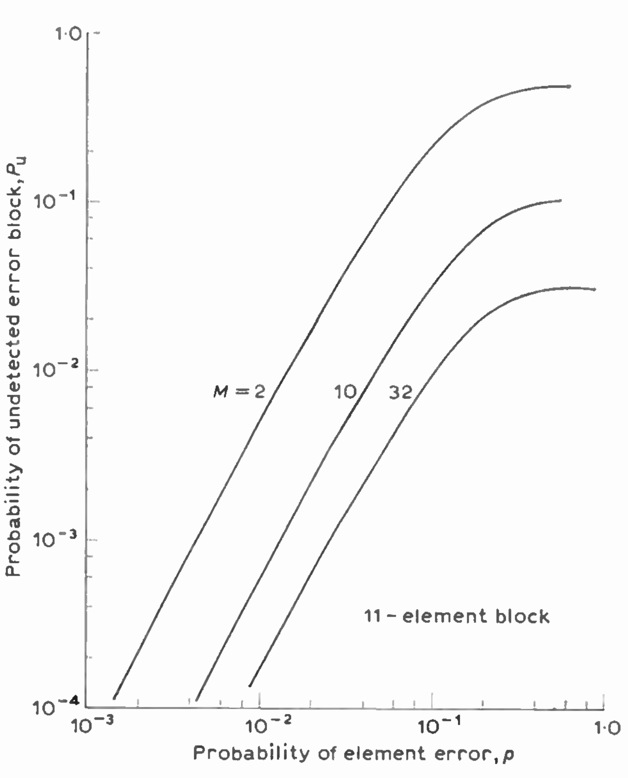

Fig. 6. Effectiveness of multi-level parity check error detection.

† ARQ = Automatic Request for Repeats.
‡ FEC = Forward Error Correction.

Hamming (7, 3) code, a majority-vote code, and a double convolution code. The general conclusion reached was that the use of multi-level coding can result in a very considerable reduction in the number of undetected errors but that there are a relatively large proportion of detected (but uncorrected) errors that can only be reduced by increasing the complexity of the decoding algorithms. It is suggested that the problem of the design of more effective multi-level FEC codes is one which would repay detailed investigation.

7 Comparison of Systems

A previous paper[6] described the performance of the Piccolo on the FCO circuits in general terms, comparing the performance (where possible) with that of a similar binary link. Improvements of circuit availability from about 80% with binary systems to about 95% with Piccolo systems are typical. Schemel[15] describes some practical comparison trials between Piccolo and a 75-baud start/stop f.s.k. system with a total deviation of 340 Hz and indicates an error rate on Piccolo of between 100 and 200 times lower than that of the binary system.

System C2 in Table 2, which was developed for digital (decimal) telemetry,[7] has been in use on the oceanographic research buoy DB1 since August 1975. Trials comparing this system with an equivalent 60-baud binary f.s.k. system have been carried out by the FCO in co-operation with the Christian Michelsen Institute in Bergen, Norway. These trials used automatic test equipment to sample the two signals for approximately 1 minute each once per hour and the tests extended over several weeks. The transmitter was sited near Bergen and transmitter power was low (25 watts), the transmitting aerial being a 6 metre whip. The signal, received at Hanslope, Buckinghamshire, tended to be very weak. The tests showed that the use of the 10-tone system reduced the average word error rate (for a 10-bit word) by a factor of about 10 to 1, which with a simple parity check would have given a reduction of undetected word error rate of about 100 to 1. The circuit availability (at 1% word errors) was about 9% with the binary system and 32% with the 10-tone system. Laboratory measurements on the two systems indicated a superiority of about 9 to 14 dB for the multi-tone technique.

8 Practical Implementation

A complete system design based on integrated circuits and implementing all the various considerations discussed above has been completed by the FCO and that 'know-how' has been purchased from HM Government for commercial exploitation. The first application was the DB1 Data Buoy telemetry system[7] and a commercial version of the 32-tone system for general telegraphy is now available[16] including the synchronizing and e.d.c. systems described. Other versions embodying some of the parameter variations discussed in this paper are planned.

9 Conclusions

It is not practicable in a short paper to do more than outline very briefly the theoretical and practical work on m.f.s.k. techniques. However, it has been demonstrated that the use of such techniques can be a very powerful weapon in medium and low-speed telegraphy over difficult h.f. channels. M.f.s.k. systems are in general less flexible than binary, requiring a simple relationship between the bits per character or word of the signal source and the bits per level of the communication system itself. The demands on the radio equipment are high, particularly in the requirement for high frequency stability, and m.f.s.k. techniques are not recommended for application to existing links based on obsolescent radio equipment. However, the required stabilities are well within the capacity of modern synthesizer techniques. A properly designed and engineered m.f.s.k. h.f. telegraphy link may be expected to give a considerable improvement in performance over its binary equivalent, amounting to 10 dB or more in signal-to-noise ratio, or a reduction of more than 10 to 1 in error rate (or more than 100 to 1 if an error coding system is used). Multi-level error coding is extremely effective but requires further theoretical investigation to realize its full potential.

Consideration of limiting factors in the various parameters suggests that m.f.s.k. techniques may be applied most effectively to data rates of about 20–250 bits/second/channel. The optimum number of tones lies in the range between about 6 tones, which gives the minimum normalized bandwidth and 32 tones, which gives the best performance in the face of long-delay multi-path and shallow-fading weak signals, and which would therefore be indicated for short-range applications. Under 'typical' long-range h.f. conditions the difference in performance of systems within this range is not a major factor.

10 Acknowledgments

The Piccolo m.f.s.k. system was designed and developed by the Foreign and Commonwealth Office (Communications Engineering Department), and the author would like to acknowledge with gratitude the active support and practical assistance of Mr. D. Bayley in the early stages, initially as Deputy Chief Engineer and later as Deputy Director, and the practical support of Mr. R. D. Neale and other colleagues of the Design Section. The radio tests mentioned in Section 7 were carried out with the help of Mr. Jan Strømme of the Christian Michelsen Institute of Bergen University, Norway.

The author is grateful to the Director of Communications of the British Foreign and Commonwealth Office and to the Controller of HM Stationery Office for permission to publish the paper.

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12 Appendix: Mathematical Analysis

Some of the equations given here can be found in various forms in many standard reference books on communication theory. Derivations and references are therefore not given unless the relationship is believed by the author not to be generally available.

12.1 Basic Principles

From Section 2, if:

M = no. of tones available for transmitting data (excluding synchronizing or error coding).

N = no. of elements per word

H = no. of bits per word.

Then

$$M \geqslant 2^{H/N}. \tag{4}$$

A system is non-redundant and therefore most efficient if this is an equality and this will be assumed in all cases.

12.2 Error Rates in Noise

Let:

R = signal-to-noise power ratio in one filter
= signal energy per element/noise power per unit bandwidth

r = an integer

W = signal energy per bit/noise power per unit bandwidth
= 'normalized signal-to-noise ratio'

N = number of tone elements per word (excluding error coding)

T = duration of each tone element (seconds).

The probability of element error for a non-fading m.f.s.k. signal with added white Gaussian noise received by matched-filter orthogonal techniques is

$$P_e = (M-1)! \sum_{r=1}^{M-1} \frac{(-1)^{r+1}}{(r+1)!\,(M-1-r)!} \exp\left(\frac{-rR}{r+1}\right). \quad (5)$$

Since there are $\log_2 M$ bits per element,

$$R = W \log_2 M. \quad (6)$$

The word error probability is

$$P_W = 1-(1-P_e)^N. \quad (7)$$

The curves of Fig. 1 were computed from equations (4)–(7), to express P_w as a function of W and M.

For slow Rayleigh fading:

$$P_e' = (M-1)! \sum_{r=1}^{M-1} \frac{(-1)^{r+1}}{(r+1)!\,(M-r-1)!} \cdot \frac{r+1}{rR+r+1} \quad (8)$$

For slow Rayleigh fading with dual diversity reception

$$P_e'' = \frac{(2+RP_e')P_e'}{2+R}. \quad (9)$$

The curves of Fig. 2 were computed in a similar manner.

12.3 Occupied Bandwidth

If it is assumed that the system keys alternate extreme tones, the output may be considered as an f.s.k. signal with modulation rate $1/T$ bauds, total deviation $(M-1)/T$ Hz and no modulation shaping. Then:[9]

$$A(f) = \frac{2E}{\pi(M-1)(x^2-1)} \quad (10)$$

where E = amplitude of unmodulated carrier

$$x = \frac{2fT}{M-1}$$

$A(f)$ = amplitude of component at frequency f

f = frequency in Hz measured from the mean of the two keyed frequencies.

Then if the normalized occupied bandwidth (specific bandwidth) is assumed to be

$$B_0 = \frac{M-1+G}{\log_2 M} \text{ Hz/bit/s (cycles/bit)} \quad (11)$$

it can be shown that G is typically about 4.

12.4 Fast Fading

The normalized energy spectral density of a Rayleigh fading carrier can be shown to be

$$G(f) = \frac{1}{f_r\sqrt{2\pi}} \exp\left[-\tfrac{1}{2}\left(\frac{f}{f_r}\right)^2\right] \quad (12)$$

where f_r is the 'r.m.s. frequency' of the Gaussian power spectrum. The response of a matched filter of the type described is of the form $(\sin \pi fT)/(\pi fT)$ and the frequency difference between any filter and an incoming signal tone is k/T Hz where k is an integer. ($k=0$ for the 'signal' filter, $k=\pm 1$ for the two 'adjacent' filters, and so on.) Then by reference to Fig. 7 it can be seen that the energy received by a filter from a fading signal is:

$$E(k) = \int_{-\infty}^{\infty} \frac{1}{f_r\sqrt{2\pi}} \times \\ \times \exp\left[-\tfrac{1}{2}\left(\frac{f-k/T}{f_r}\right)^2\right] \cdot \left(\frac{\sin \pi fT}{\pi fT}\right)^2 df.$$

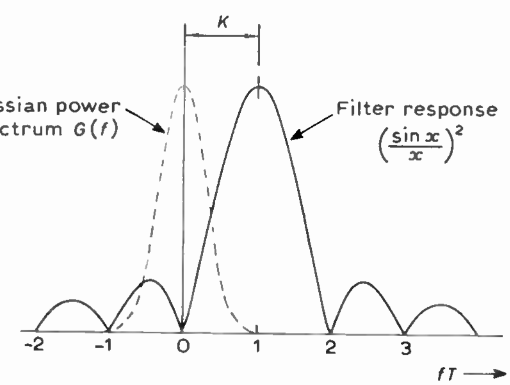

Fig. 7. Derivation of theoretical error rate in fast fading.

It can be proved that f_r is related to the fading rate f_m (as normally defined) by $f_m = af_r$ where $a = 1{\cdot}475$. It is convenient to express the fading rate in terms of the signal element rate, i.e. the number of signal elements per fade period $m = 1/Tf_m$, and then $f_r = 1/amT$. Therefore

$$E(k) = \int_{-\infty}^{\infty} \frac{amT}{2\pi} \cdot \left(\frac{\sin \pi fT}{\pi fT}\right)^2 \times \\ \times \exp\{-\tfrac{1}{2}[am(fT-k)]^2\} \, . \, df. \quad (13)$$

The value of $E(k)$ can be computed from equation (13) and then if $S(k) = E(0)/E(k)$, it can be shown that since each has a Rayleigh amplitude distribution and assuming they are uncorrelated the probability of element error through selecting one specific filter at a frequency k/T from the signal is

$$p_k = \frac{1}{1+S(k)}. \quad (14)$$

Then in a complete system the total probability of element error is

$$p_t \simeq 2p_1 + 2p_2. \quad (15)$$

The computed curves for p_1, p_2, and p_t are shown in Fig. 5, together with measured values of p_t. It can be seen that a convenient empirical relationship for the measured values of p_t is $p_t \simeq 0{\cdot}2/m$, i.e. on a non-diversity m.f.s.k. system on a Rayleigh fading signal, an element error occurs about once every five fade periods.

Similar calculations and measurements for an optimal selection dual diversity system (p_d) suggest that the error rate is approximately $3/m^4$.

12.5 Modulo-*M* Parity Check

Consider an m.f.s.k. signalling system using M tones in which each tone is allocated an arithmetical 'weight' of 1 to M. Input data are coded into blocks of (B-1) elements, and to each block is added a check element selected so that the total, in modulo-M arithmetic, of the block of B elements is zero. If p is the probability of error of each element, the probability of n errors in a block is

$$P_n = C_n{}^B p^n (1-p)^{B-n}. \quad (16)$$

The author has been unable to trace in the literature a general expression for the probability of errors cancelling (i.e. the modulo-M total remaining at zero for multiple errors) but has derived empirically the following expression:

$$P_c = -\frac{1}{(1-M)^{n-1}} \sum_{r=0}^{n-2} (1-M)^r. \quad (17)$$

The probability of an undetected error block is then

$$\begin{aligned} P_u &= \sum_{n=1}^{B} P_n \,.\, P_c \\ &= \sum_{n=1}^{B} \left[\sum_{r=0}^{n-2} (1-M)^r \right] \frac{-1}{(1-M)^{n-1}} \times \\ &\qquad \times \frac{B!}{B!(B-n)!} p^n (1-p)^{B-n}. \quad (18) \end{aligned}$$

This equation is plotted in Fig. 6 for $B=11$ and $M=2$, 10, 32. It has been checked mathematically for various asymptotic limits and also by measurement and simulation on 32-level and 10-level systems.

Manuscript first received by the Institution on 21st December 1976 and final form on 3rd June 1977.
(Paper No. 1783/Comm. 154)

UDC 621.372.553

RC active variable group-delay equalizers with independent Q and ω controls

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SUMMARY

The high-speed transmission of data over voice frequency telephone links creates the need for variable group-delay equalization networks. This paper describes two alternative group-delay sections, the characteristics of which may be continuously adjusted over the approximate bandwidth (300 to 3000 Hz). Both of the proposed networks provide for the desired degree of group-delay adjustment using only two resistive controls neither of which demand high precision components.

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1 Introduction

The high-speed transmission of data over voice-frequency telephone links usually requires some group-delay equalization of the transmission circuit in order to ensure satisfactory data reception. In practice, the degree of equalization required would often be determined by a series of *in-situ* measurements. An equal-ripple type of equalization would then be obtained by dynamically adjusting the characteristics of a cascade connection of non-minimum-phase second-order sections. When it is desirable to avoid frequency-dependent insertion losses, as is the case where the circuit has a satisfactory amplitude characteristic, then it is necessary that each of the equalizer sub-section transfer functions have the general second-order all-pass form:

$$T_v(s) = \frac{s^2 - \dfrac{s\omega_0}{Q} + \omega_0^2}{s^2 + \dfrac{s\omega_0}{Q} + \omega_0^2} \qquad (1)$$

Although the literature contains many examples[1-10] of all-pass networks that offer resistive and therefore continuous control of the frequency parameter ω_0, their value as variable group-delay filters is often limited by considerations such as:

(i) Restricted Q-factor range. This constraint is particularly severe for some single amplifier structures[1-6] where, as noted by Soliman[11], the poles of $T_v(s)$ are confined to the negative real axis.

(ii) Component ganging. The useful practice of ganging variable components controlling the parameter ω_0, can become relatively expensive and mechanically complex when the number of resistors defining the frequency exceeds two[3,6], or, where the variable-parameter Q sensitivities

$$S_x^Q = \frac{x}{Q}\cdot\frac{dQ}{dx} \qquad (2)$$

are high[10], in which case the permissible tracking errors are correspondingly reduced.

(iii) Multiplicity of controls. Since the group-delay characteristics corresponding to eqn. (1) are completely specified in terms of only the two design parameters; ω_0, and Q, only two controls are required. However, the poles and zeroes realized by most well-known structures[7-9] are functions of different resistive elements. In such cases, it will therefore be necessary to manipulate several sets of components in order to maintain the desired all-pass form.

The purpose of this paper is to show that low-sensitivity all-pass structures can be developed having group-delay characteristics that may be simply varied over a wide range using the minimum number of controls.

2 Development of Variable Group-delay Structures

It will be clear from the foregoing that desirable structures should have their natural frequency, ω_0, defined and

B

therefore controlled, by a single pair of resistors. In addition, the networks should provide a variable Q factor capability in an insensitive manner thereby easing the problems associated with ganged-resistor frequency control (by eliminating the costly high-precision alignment requirement), and allowing for the use of relatively low precision capacitors. The simple configurations shown as Fig. 1(a) and 1(b) are both well-suited as basic building blocks and may be recognized as an undamped two-integrator loop circuit and a cascade of first-order all-pass networks respectively.

2.1 Integrator-based Group-delay Equalizer

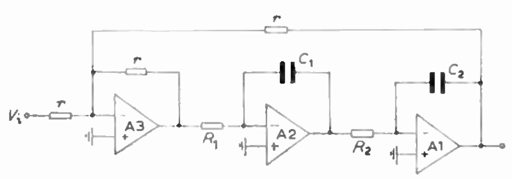
The development of variable frequency/variable Q factor all-pass networks using the two-integrator-loop block of Fig. 1 (a) is straightforward and simply requires that the set of functions

$$T_{v1}(s) = \frac{a_1 \omega^2{}_0}{s^2 + \dfrac{s\omega_0}{Q} + \omega_0{}^2} \tag{3}$$

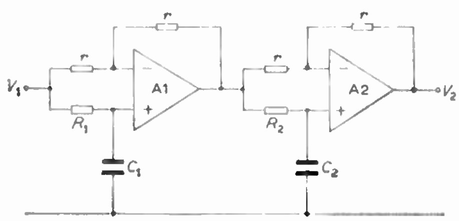
$$T_{v2}(s) = \frac{a_2 \dfrac{s\omega_0}{Q}}{s^2 + \dfrac{s\omega_0}{Q} + \omega_0{}^2} \tag{4}$$

$$T_{v3}(s) = \frac{a_3 s^2}{s^2 + \dfrac{s\omega_0}{Q} + \omega_0{}^2} \tag{5}$$

be generated and thereafter summed with the appropriate weightings. It should, however, be noted that whereas $T_{v1}(s)$ and $T_{v3}(s)$ are orthodox low-pass and high-pass functions respectively, the band-pass function $T_{v2}(s)$ has been assigned a Q dependent numerator (which effectively reduces the number of control variables in the complete filter section). Although the basic undamped structure of Fig. 1(a) simultaneously realizes low-pass, band-pass and high-pass transmissions at the outputs of amplifiers 1, 2 and 3 respectively, the direct application of standard



(a) Undamped two-integrator loop network.



(b) Cascaded first-order all-pass sections.

Fig. 1.

damping techniques does not lead to the desired band-pass form. However, the provision of damping via a potentiometer and unity-gain buffer as shown in Fig. 2(a) gives the transmissions at the outputs of amplifiers 1 and 3 as:†

$$T_{v1}(s) = \frac{V_1}{V_i} = \frac{-1}{s^2C_1C_2R_1R_2 + 3\alpha sC_2R_2 + 1} \tag{6}$$

$$T_{v3}(s) = \frac{V_3}{V_i} = \frac{-s^2C_1C_2R_1R_2}{s^2C_1C_2R_1R_2 + 3\alpha sC_2R_2 + 1} \tag{7}$$

and yields the additional function

$$T_{v2} = \frac{V_2}{V_i} = \frac{\alpha sC_2R_2}{s^2C_1C_2R_1R_2 + 3\alpha sC_2R_2 + 1} \tag{8}$$

in which the parameter α appears as a coefficient of the s term in both numerator and denominator. By comparing the objective function, eqn. (1), with eqns. (6), (7) and (8) it will be evident that the required all-pass form may be obtained as the weighted sum:

$$T_v(s) = -T_{v1}(s) - 3T_{v2}(s) - T_{v3}(s) \tag{9}$$

The complete variable group-delay configuration, incorporating this summation is shown as Fig. 2(a) and realizes the transfer function

$$T_v(s) = \frac{V_0}{V_i} = \frac{s^2C_1C_2R_1R_2 - 3\alpha sC_2R_2 + 1}{s^2C_1C_2R_1R_2 + 3\alpha sC_2R_2 + 1} \tag{10}$$

From eqns. (1) and (10) the natural frequency, ω_0, and selectivity, Q, can be readily identified as

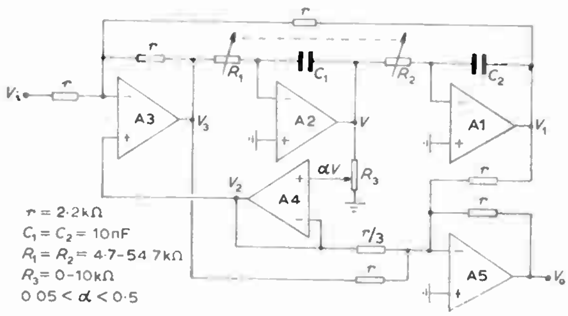
$$\omega_0 = \frac{1}{\sqrt{(C_1C_2R_1R_2)}} \tag{11}$$

and

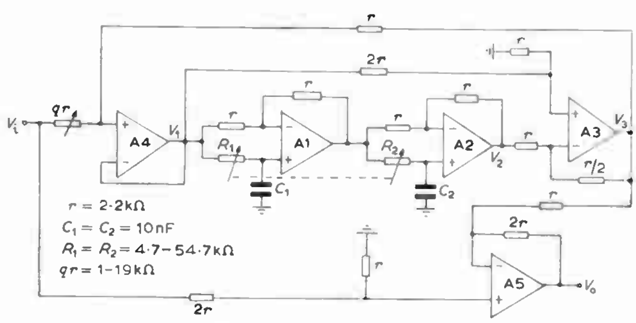
$$Q = \frac{1}{3\alpha}\sqrt{\left(\frac{C_1R_1}{C_2R_2}\right)} \tag{12}$$

In designing for an adjustable group-delay characteristic it is obviously convenient to select equally valued fixed capacitors ($C_1 = C_2 = C$) and employ a ganged pair of nominally equal variable resistors ($R_1 = R_2 = R$) as the means of controlling ω_0 (thereby varying the position of the group-delay curve along the ω axis). For this choice, the selectivity or Q factor is virtually independent on the RC products and effectively becomes: $Q = 1/3\alpha$. Since the potentiometric fraction α can be varied within the range $0 < \alpha < 1$, the Q factor, which governs the shape and maximum value of the group-delay curve, may be regarded for all practical purposes as being completely adjustable. It will be evident that the foregoing design meets the specified objectives in that it provides a realization of eqn. (1) in which ω_0 and Q can be independently and simply varied over a wide range in an insensitive manner using only two controls.

† These expressions and those subsequently derived, assume the operational amplifiers to be ideal devices. This assumption is useful in that it simplifies and clarifies the presentation and may be justified on physical grounds for the application under consideration.



(a) Integrator-based variable group-delay section.



(b) Cascaded unit group delay section.

Fig. 2.

2.2 Cascaded Unit Group-delay Equalizer

The use of the cascaded all-pass unit of Fig. 1(b) as the basic building block requires a different synthesis technique to that previously described. A somewhat indirect but nonetheless effective procedure can be summarized as follows:

(a) Construct a negative-real-pole band-pass function with a gain at the resonant frequency of unity.
(b) Use positive feedback to enhance the selectivity of the low-Q section (retaining the unity gain feature).
(c) Subtract twice the enhanced-Q-factor bandpass function from unity.

From an analysis of the basic cascaded all-pass section of Fig. 1(b), which gives the overall transfer function as

$$T_1(s)=\frac{V_2}{V_1}=\frac{s^2C_1C_2R_1R_2-s(C_1R_1+C_2R_2)+1}{s^2C_1C_2R_1R_2+s(C_1R_1+C_2R_2)+1} \quad (13)$$

it may be seen that step (a) of the synthesis scheme proceeds by implementing the subtraction

$$T_2(s)=\tfrac{1}{2}(1-T_1(s)) \quad (14)$$

as shown in Fig. 2(b), to give

$$T_2(s)=\frac{V_3}{V_1}=\frac{s(C_1R_1+C_2R_2)}{s^2C_1C_2R_1R_2+s(C_1R_1+C_2R_2)+1} \quad (15)$$

It will be noted that the coefficients of the s term in the numerator and denominator of eqn. (15) are identical. As will be demonstrated, this feature enables the subsequent enhancement of the Q factor, as required in step (b), without incurring sensitivity penalties. The enhanced Q capability is secured as shown in Fig. 2(b), by feeding the output of amplifier A3 back to a resistive summing junction (buffered by the amplifier A4) via the fixed and variable resistors r and qr respectively. The p.f.b. loop so formed, yields the function:

$$T_3(s)=\frac{V_3}{V_1}=\frac{1}{q+1}\cdot\frac{T_2(s)}{1-\dfrac{q}{q+1}\cdot T_2(s)} \quad (16)$$

Therefore from (15):

$$T_3(s)=\frac{\dfrac{s(C_1R_1+C_2R_2)}{q+1}}{s^2C_1C_2R_1R_2+\dfrac{s(C_1R_1+C_2R_2)}{q+1}+1} \quad (17)$$

The third and final step in the all-pass function synthesis simply requires the subtraction

$$T_v(s)=1-2T_3(s) \quad (18)$$

which is implemented as shown in Fig. 2(b) and gives

$$T_v(s)=\frac{V_0}{V_1}$$

$$=\frac{s^2C_1C_2R_1R_2-s\left(\dfrac{C_1R_1+C_2R_2}{q+1}\right)+1}{s^2C_1C_2R_1R_2+s\left(\dfrac{C_1R_1+C_2R_2}{q+1}\right)+1} \quad (19)$$

From eqns. (1) and (19):

$$\omega_0=\frac{1}{\sqrt{(C_1C_2R_1R_2)}}, \quad (20)$$

and

$$Q=(q+1)\frac{\sqrt{(C_1C_2R_1R_2)}}{C_1R_1+C_2R_2} \quad (21)$$

Whilst eqn. (20) shows the natural frequency ω_0 to be dependent on the RC products in the usual manner, eqn. (21) indicates a special relationship exists between them and the Q factor. If a sensitivity analysis is performed as given by eqn. (2), the following results are obtained:

$$S_{C_1}{}^Q=S_{R_1}{}^Q=-S_{C_2}{}^Q=-S_{R_2}{}^Q$$

$$=\tfrac{1}{2}\left[\frac{C_2R_2-C_1R_1}{C_2R_2+C_1R_1}\right] \quad (22)$$

Thus for the attractive design choice $C_1=C_2=C$, and $R_1=R_2=R$, the above sensitivities are all zero; the Q factor being simply defined by the variable resistor qr. Although zero sensitivities are desirable it will be appreciated that sensitivities of unity order, as exhibited by the integrator based configuration, are quite acceptable.

It will be clear from the form of eqn. (19), and the preceding design details, that the cascaded unit group-delay equalizer described, also offers adjustable characteristics using only two independent controls.

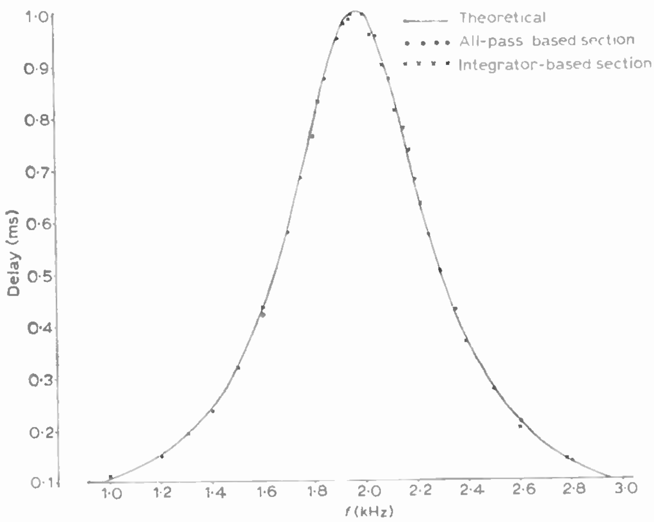


Fig. 3. Theoretical and experimental group-delay responses.

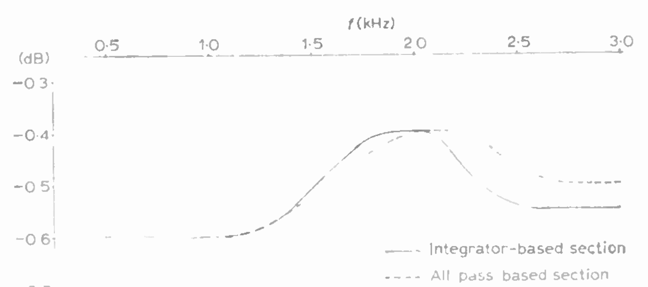


Fig. 4. Amplitude responses.

3 General Considerations

In the preceding Sections two alternative multi-amplifier structures have been described both of which offer controllable group-delay characteristics. It will be of interest to consider briefly the factors which could influence the selection of one network in preference to the other.

Although both configurations employ the same number of amplifiers, the integrator-based section is more economical in its use of fixed resistors, requiring only seven compared with thirteen. This advantage is reinforced from a consideration of the sensitivities $S_{r_i}{}^{Q}$ (r_i representing any fixed resistor) since a detailed analysis will show that these measures are uniformly low for the Fig. 2(a) structure but range up to a value Q for the cascaded unit configuration. However, as previously noted, the latter circuit exhibits outstandingly low values for $S^{Q}{}_{C_1,C_2,R_1,R_2}$ (in that they approach zero in practice). Moreover, it can be shown that the Q factor realized by the cascaded unit circuit is a factor of Q times less sensitive to variations in the amplifier gain-bandwidth products than is that of the integrator type network. It should be emphasized that because of the relatively modest range of Q and frequencies involved ($Q < 5$, $\omega_0/2\pi < 3{\cdot}4$ kHz)† neither network should suffer undue performnace degradation given reasonable quality components.

Practical versions of both equalizers have been constructed and tested, using readily available components including polystyrene capacitors ($\pm 5\%$ tolerance), metal oxide resistors ($\pm 2\%$ tolerance) and 741-type operational amplifiers. The component values shown against Figs. 2(a) and 2(b) were chosen to permit a Q variation within the range $1/\sqrt{2} < Q < 5$ and a frequency excursion 290–3400 Hz. The practical results obtained have generally demonstrated that the group-delay responses of either network can be adjusted simply and efficiently as required, via the ω and Q controls. The group-delay responses presented as Fig. 3 typify the close agreement found for the experimental and theoretical curves and correspond to a peak delay of 1 ms at 2 kHz (for which case $Q=\pi$). The corresponding amplitude responses (ideally independent of frequency) which are shown as Fig. 4, indicate a gain variation of only 0·2dB (together with an inconsequential constant flat loss of approximately 0·5dB) which is consistent with the component tolerances previously noted.

† This range covers that specified by the CCITT Rec. M102.

4 Acknowledgment

The author wishes to acknowledge the financial support of the Radio Research Board of Australia for this work.

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Manuscript first received by the Institution on 25th October 1976, in revised form on 31st January 1977 and in final form on 13th April 1977. (Paper No. 1784/CC275).

UDC 681.326.3

Design of self-checking and fault-tolerant microprogrammed controllers

IAN WILLIAMSON, M.A.*

Based on a paper presented at the Conference on Computer Systems and Technology held at the University of Sussex from 29th to 31st March 1977

SUMMARY

A realization of a self-checking and fail-safe programmable controller which uses a new control memory organization to give a simple and elegant implementation is described.

Using this self-checking controller as an example, a new approach to fault-tolerant design referred to as Dual-Fail-Safe (DFS) is presented which utilizes two self-checking modules. The resulting fault-tolerant system is shown to be less costly and significantly more reliable than a conventional Triple-Modular-Redundancy (TMR) system.

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1 Introduction

Perhaps the major difficulty in the assessment and understanding of fault-tolerant design techniques is that virtually all the active research work has been aimed at a technology which is no longer applicable. The early work by Von Neuman,[1] Tryon[2] and Armstrong[3] clearly illustrates this point as it primarily concerns computer systems constructed of discrete components. The current use and understanding of Triple Modular Redundancy (TMR) is due to its continued applicability regardless of advances in semiconductor technology.

Recently, research work has turned towards the design of easily-maintained sub-systems recognizing that, in the majority of applications, availability is of greater importance than mean time to first failure. This research work covers many aspects of digital design, of which improved fault-diagnosis is a major example, but of particular relevance to this paper is the design of *self-checking* digital circuits. A self-checking circuit is one which is tested by its functional inputs alone, and fails in a secure and safe manner. Self-checking circuits can be designed to detect, reliably, the class of failures which occur in l.s.i. circuits. They are therefore relevant and important to current and foreseeable semiconductor technologies.

In this paper, we define the self-checking properties of a circuit, and justify their relevance to current semiconductor technologies. We then describe as an example a self-checking microprogrammable controller, with a novel memory architecture. Finally, we describe a new fault-tolerant design technique and, using the microprogrammable controller as an example, we compare its cost and performance with an equivalent TMR design.

2 Self-checking Circuits

Self-checking circuits were first described by Carter and Schneider[4] in their discussion of dynamically checked structures. Anderson[5] provides a complete and thorough discussion of self-checking properties, and put forward the following definitions in the context of a combinational circuit.

Definition 1:

A circuit is *self-testing* if, for every fault from a prescribed set, the circuit produces a non-code-space output for at least one code-space input.

Definition 2:

A circuit is *fault-secure* if, for every fault from a prescribed set, the circuit never produces an incorrect code-space output for code-space inputs.

Definition 3:

A totally *self-checking circuit* is both self-testing and fault-secure.

It is clear from these definitions that both the inputs and outputs of self-checking circuits must be encoded, and the choice of error detecting code determines the range of faults, i.e. the prescribed set of faults, for which the circuit retains its self-checking properties.

The most useful class of error detecting code for this application is the fixed weight or k-out-of-n code. With this class of code, circuits can be designed with self-checking properties for both single and unidirectional s-a-1 or s-a-0 faults.

Definition 4:

A unidirectional fault is a multiple fault in which all faulty lines are stuck at the same logical value, s-a-1 or s-a-0.

In a paper on the design of self-checking microprogrammable controllers, Cook *et al.*[6] conclude that the unidirectional fault model covers a large majority of the faults which occur in a semiconductor memory. Reference 7 includes an independent analysis of l.s.i. failure modes, based upon the RADC reliability prediction equation (8). This analysis suggests that whilst 30% of chip failures are catastrophic or linked, producing multiple faults within the circuit, only 1% of chip failures produce faults which are not unidirectional. Thus we conclude that the unidirectional fault model is soundly based upon existing knowledge about integrated circuit failure modes.

Diaz in two recent papers[9,10] has extended the concept of self-checking circuits to cover sequential as well as combinational circuits, and has specifically addressed the design of self-checking microprogrammed controllers.

3 Microprogrammed Controller

We use as our reference example, the microprogrammed controller shown in Fig. 1. This circuit is representative of a data bus controller in a fault-tolerant multiprocessor configuration. It provides a field of 15 TO instructions, 15 FROM instructions and can test up to 14 external conditions. Included in the controller is an 8-bit counter which serves as a watchdog timer. The r.o.m. output is 24 bits wide, which is subdivided as follows:

Next Address	8 bits
Condition Vector	4 bits
Instruction Fields	2×4 bits
Flags (internal and external)	4 bits
	24 bits

The address modifier (AM) increments the next address (NA) when the tested external condition is true. Thus, a program branch has only two destinations which are stored at the current NA, and NA+1.

The microprogrammed controller described above is a sequential machine with several feedback loops. To introduce self-checking properties, each feedback path must be designed in such a way that circuit failures cause faults which are detectable by the encoding. We therefore encode the contents of the r.o.m. using an error detecting code—such as a k-out-of-n code. The remainder of the controller must then be constructed from self-checking circuits which accept k-out-of-n encoded inputs, and generate similarly encoded outputs. By observing Anderson's rules of self-checking design,[5] namely, that each circuit be securely located, fully exercised and code disjoint, then a completely self-checking controller can be assembled.

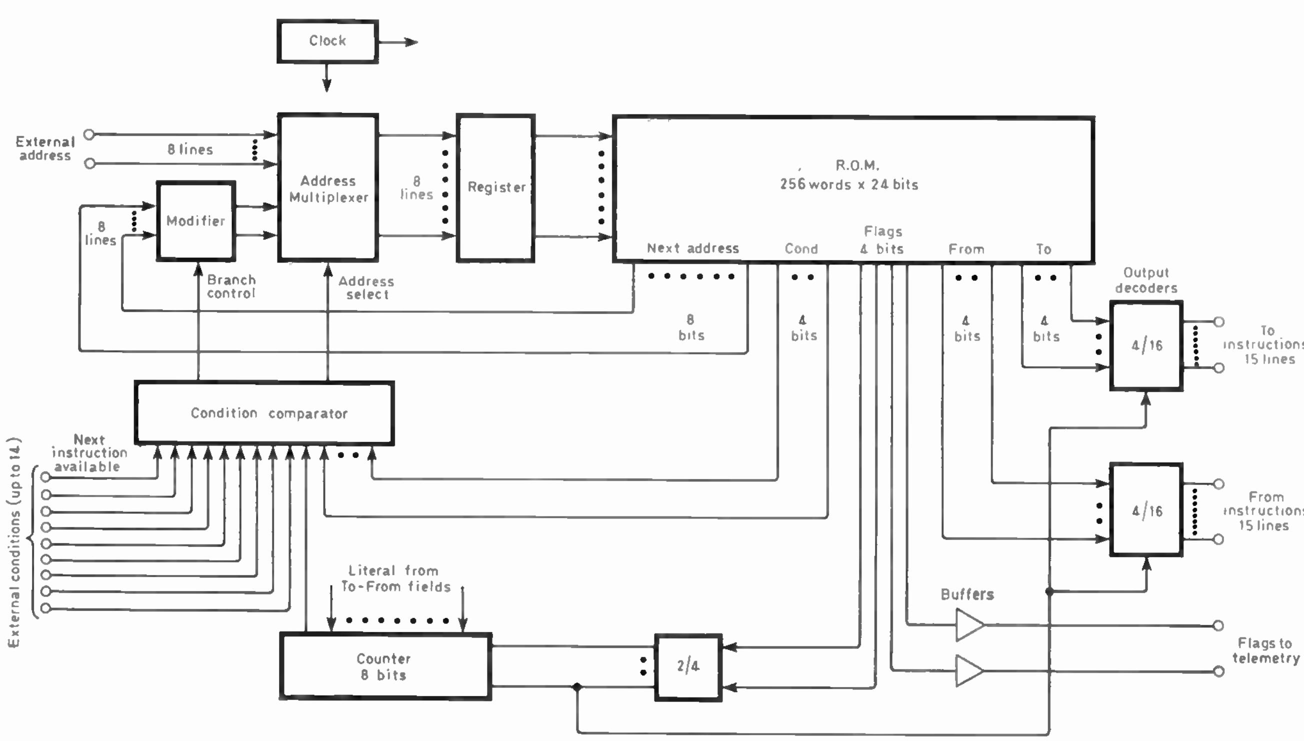


Fig. 1. Microprogrammed controller.

In practice, the most difficult loop to protect is that containing the Next Address, Address Modifier, and r.o.m. address decoder. If the NA is incorrect, owing to a failure in the loop, then at the next step a correctly encoded line of data may be read from the wrong part of the r.o.m. Subsequent steps will also be erroneous, but undetectable, and the controller may enter a permanently erroneous cycle. Because of the combinational mapping in the r.o.m., failure in the NA circuits does not necessarily fit the unidirectional fault model defined above. We must therefore derive some means of ensuring that NA failures are reliably detectable at the r.o.m. output—by examining carefully the architecture of a semiconductor memory.

An i.c. r.o.m. is generally organized as shown in Fig. 2. For optimum performance the r.o.m. matrix must be kept as square as possible. For example, a 2048-bit memory organized as 256 by 8 bits would preferentially contain a matrix arranged 64 × 32 cells, not 256 × 8 cells; the output word being selected from the 32- or 64-line output of the matrix under the control of part of the address word.

Diaz[10] has suggested a self-checking r.o.m. organization where the address is specified by a k-out-of-n code. Unfortunately, a k-out-of-n code does not factorize, as does a binary codeword, and therefore a 5-out-of-10 codeword must be fully decoded to define the 252 possible codewords. Thus, each codeword must correspond to a single line of the r.o.m. matrix, resulting in a matrix which is far from square in most applications. This is illustrated in Fig. 3 for a 3-out-of-6 code address, and a 4-bit output word.

An alternative to the Diaz organization is to use two separate fixed-weight codewords addressing the x and y lines of the r.o.m., thereby achieving an efficient implementation. For example, a 100-word by 1-bit r.o.m. can be constructed with a 10 × 10 square matrix using a 2-out-of-5 code decoder, and a 2-out-of-5 code multiplexer, which are standard self-checking circuits well documented in reference 7.

In the microprogrammed controller example described above, we found that a two-part address encoded as 4-out-of-9 code and 1-out-of-2 code produced a minimum cost self-checking design. A single part 5-out-of-10 code

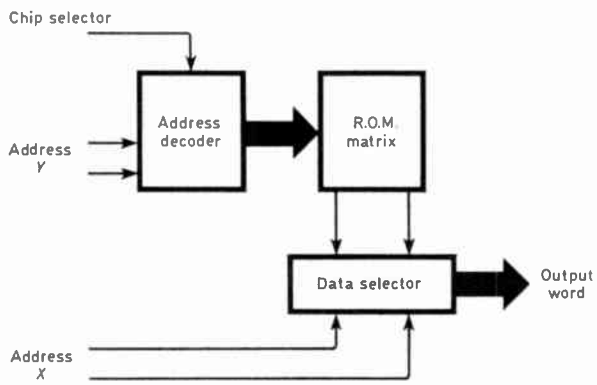


Fig. 2. Conventional semiconductor r.o.m.

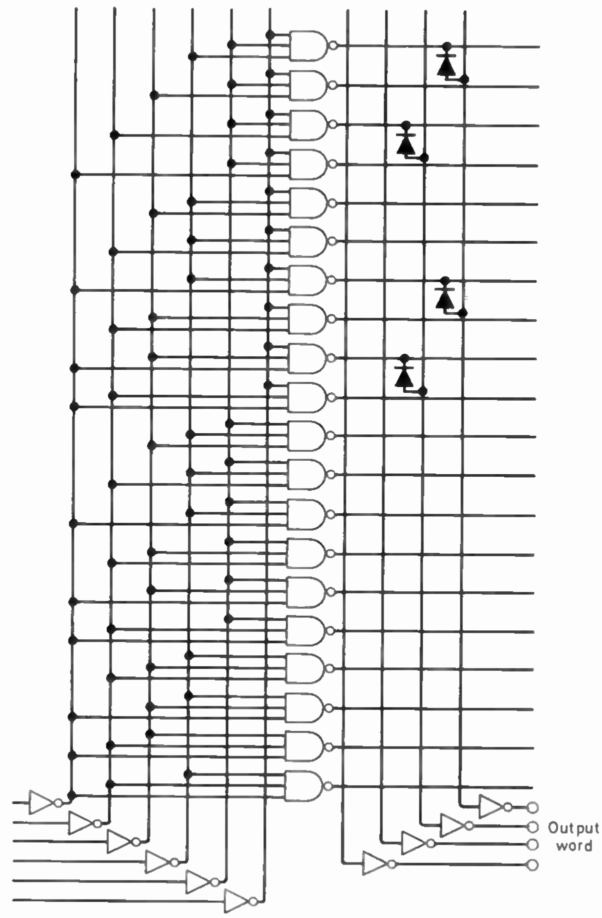


Fig. 3. 3-out-of-6 code self-checking r.o.m.

address was not practicable with current semiconductor technology.

We can now define the fixed-weight encodings for the self-checking microprogrammed controller.

Next Address	11 bits	4-out-of-9 code and 1-out-of-2 code
Condition Vector	5 bits	2-out-of-5 code
Instruction Fields	2 × 6 bits	2 × 3-out-of-6 code
Flags	8 bits	4 × 1-out-of-2 code
	36 bits	

The NA is encoded as a 4-out-of-9 codeword, and a 1-out-of-2 codeword. Modification of the address for branch control is conveniently carried out on the 1-out-of-2 codeword, and a known self-checking circuit is available for the mapping. The reduction circuit for checker outputs (RCCO), sometimes referred to as a two-rail checker, has the following logical function:

$$f = a_1 b_2 + b_1 a_2$$

$$g = a_1 a_2 + b_1 b_2$$

If f and g are the NA bits corresponding to a_1 and b_1, and a_2 and b_2 are the branch control bits encoded 10 for branch, and 01 for no branch, then this function correctly modifies the NA as shown in the truth-table

a_1b_1 \ a_2b_2	00	01	10	11
00	00	00	00	00
01	00	01	10	11
10	00	10	01	11
11	00	11	11	11

A 00 output will fail to select any line of r.o.m. giving an all 0's output at the next step; a 11 output will activate two or more lines of r.o.m., and the r.o.m. output will not be a permitted codeword. The circuit implementation of this function requires 6 NAND gates.

Modifier circuits for larger codes are described in reference 7, but their cost rapidly becomes prohibitive. In general, a k-out-of-n codeword can be modified into another k-out-of-n codeword using a self-checking address modifier containing $2(n/k)$ AND gates and n OR gates. For example, a 4-out-of-9 code address modifier requires 252 5-input AND gates and 9 126-input OR gates.

The remaining circuit elements of the self-checking microprogrammed controller are constructed from standard self-checking circuits which are collected together as a convenient anthology in reference 7. To function as a stand-alone self-checking and fail-safe controller, self-checking k-out-of-n code checkers[5] are applied to the instruction fields of the control r.o.m. Note that the codes which are used within the controller itself, such as the NA, need not be checked since faults will always propagate to the output fields where they are easily detectable.

The complete self-checking controller is approximately 40% more expensive in terms of gate count and power consumption than its simplex counterpart.

4 Dual Fail Safe Design

It is almost axiomatic that duplication is necessary for full fault detection, and triplication is necessary for full fault tolerance. However, the self-checking controller described above achieves fault detection for some 99% of l.s.i. chip failures with only a 40% increase in cost. We now describe a new approach to fault tolerant design, called Dual Fail Safe (DFS) design, which provides greater reliability enhancement than Triple Modular Redundancy (TMR) at a lower cost, by duplicating the self-checking controller described above.

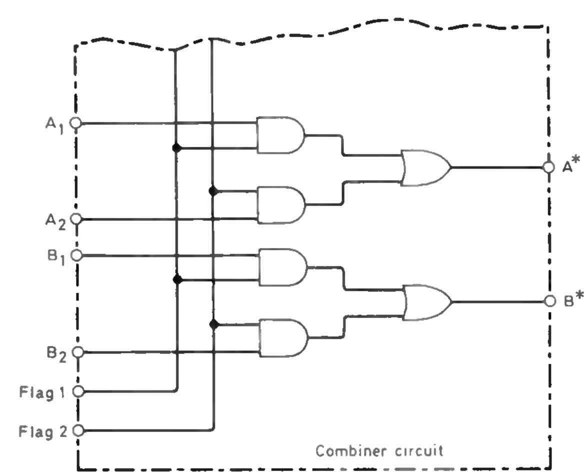
If we take a combinational circuit module, for example, a decoder, which produces a set of outputs subject to a certain set of inputs, then we can define a set of output errors which is generated by a defined and limited set of faults within the circuit.

In DFS design, we take two such circuit modules and combine corresponding outputs from each module using a Boolean function AND or OR. Now, dependent upon the logic function of the combiner, we can define a fail-safe state for the circuit module. Thus if the combiner is an AND gate, then the fail-safe state is a 1. Part of the error set which we are trying to mask will therefore be fail-safe and masked by the combiner, and it is only necessary to detect the non-fail-safe state.

We therefore define critical and sub-critical errors at the output of the module, sub-critical errors being those masked by the logic function in the combiner. If we design a checking circuit to detect the critical errors, then we can force the module output to a sub-critical error condition. We have, therefore, a *reduced-complexity checker* because only a fraction of the output errors need be detected. It is this reduced complexity in the checking function which enables DFS to compete successfully with TMR.

The practical implementation of this scheme is shown in Fig. 4. The two 'fail-safe' modules generate, in addition to their normal outputs, a flag which indicates the status of the module. The outputs and two flags are fed to a combiner which is constructed of standard AND-OR select gates—which are available in t.t.l. or c-m.o.s.

Since the combiner is an OR gate, the sub-critical error is s-a-0, and the critical error is s-a-1. Critical errors are inhibited at the preceding AND gate by the flag which is a 0 when a critical error has occurred, and a 1 when no error or a sub-critical error has occurred.



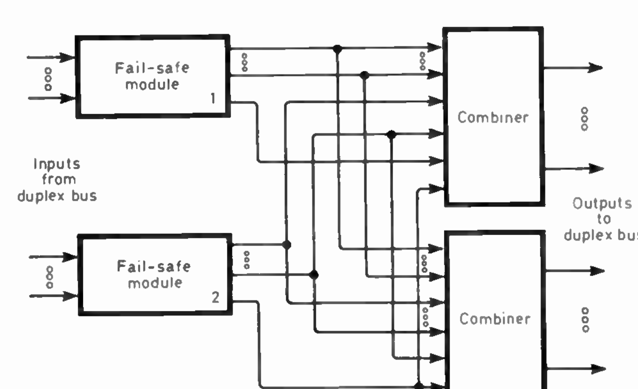


Fig. 4. Dual-Fail-Safe system.

The operation and properties of DFS are best illustrated by an example. Figure 5 gives the circuit of a self-checking 2-out-of-5 code decoder. This circuit generates 10 outputs in negative logic corresponding to the 10 codewords of the 2-out-of-5 code.

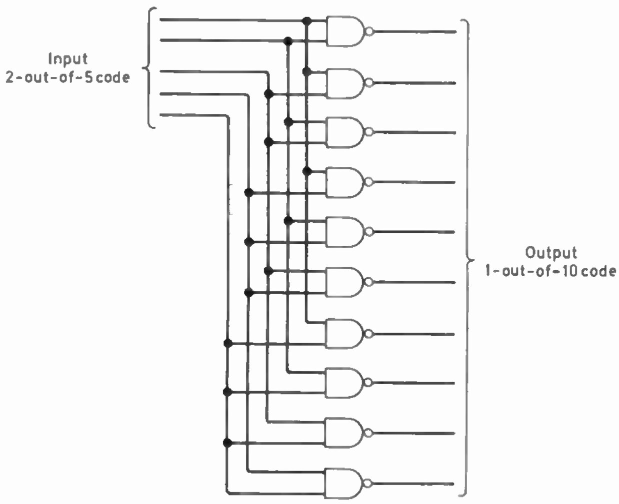


Fig. 5. Self-checking 2-out-of-5 decoder.

Let us examine the error conditions on the outputs corresponding to fault conditions in the circuit.

(i) *Gate output s-a-0*

If one or more NAND gates are s-a-0, then either the correct output plus additional incorrect outputs, or the correct otuput alone, will be activated.

(ii) *Gate output s-a-1*

If one or more NAND gates are s-a-1, then either the correct output is activated, or no outputs are activated.

(iii) *Gate input s-a-1*

As gate output s-a-0.

(iv) *Gate input s-a-0*

As gate output s-a-1.

From the above analysis, we see that a fault (and unidirectional faults) generate easily detectable error conditions at the output, and furthermore, if we consider the output to be in 1-out-of-10 code, then any correctly encoded output is indeed a correct output. Thus the important error condition which is avoided is a correctly encoded, but erroneous, output.

If we combine two such decoders using the AND-OR combiner shown in Fig. 6, then the critical error condition is a s-a-1 output. Examination of the error conditions shows that s-a-1 errors are generated by condition (ii) and occur only when no outputs are active. Hence, all 10 decoder outputs are in the logical 1 condition. The reduced-complexity checker which generates a logical 0 flag from such a condition is simply a 10-input NAND gate.

The circuit shown in Fig. 6 is fault tolerant to single (and unidirectional) faults in the decoder modules.

5 Comparison of DFS and TMR

Two copies of the self-checking microprogrammed controller described above can be combined, using the circuit arrangement shown in Fig. 6, to form a fault-tolerant DFS system. We have compared the reliability and cost of this DFS system with an equivalent TMR design formed by triplicating the simplex controller shown in Fig. 1.

The analysis is conducted using the two extremes of:

(i) designs using standard chips where possible, and

(ii) designs which are constructed entirely of custom chips, including custom voter elements in TMR and custom combiners in DFS.

Table 2.

(1) Standard Chips

	Simplex	TMR	DFS
Package count	1	10·7	8·4
Pin count	1	7·9	6·1
Power dissipation	1	3·5	1·9
Lifetime at 99·5% confidence	1	7·34	9·36
Lifetime at 95% confidence	1	2·49	3·53

(2) Custom Chips

	Custom simplex	Standard simplex	TMR	DFS
Package count	1	7	13·5	8
Pin count	1	1·6	5·8	3·6
Power dissipation	1	1·5	3·6	2·85
Lifetime at 99·5% confidence	1	0·66	5·08	9·25
Lifetime at 95% confidence	1	0·66	2·68	3·23

Reliability has been estimated and optimized using a prediction equation based upon the MIL HDBK 217B

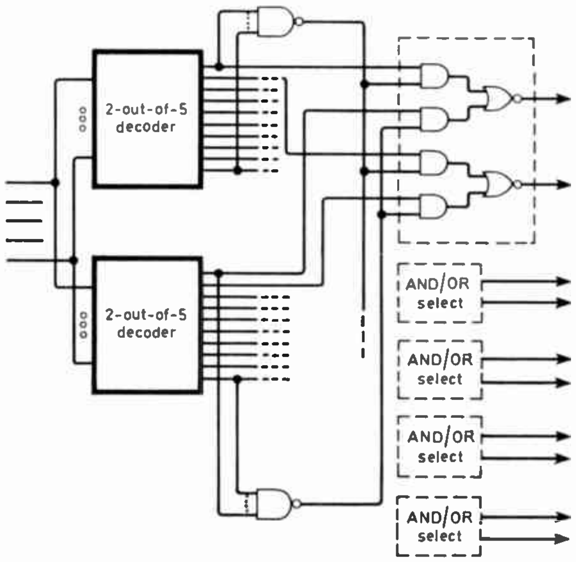


Fig. 6. Dual-Fail-Safe decoder.

model.[11] Therefore, the baseline reliability of the simplex represents the best that can be achieved with existing standard technologies.

We have also assumed that the controller is embedded within a larger fault tolerant system, and therefore redundant output circuits are necessary. The analyses of the two designs are compared in Table 2.

The DFS design is consistently cheaper, and more reliable, than the corresponding TMR design. It is interesting to note that the DFS design retains its reliability enhancement with respect to the simplex design, regardless of the actual reliability required. In contrast, TMR will only provide a reliability enhancement at high levels of reliability. Thus the lifetime of TMR at 50% confidence is always less than the simplex lifetime, whereas the lifetime of DFS is greater than simplex at this level of confidence.

6 Conclusions

In this paper, we have described a new approach to fault-tolerant design, Dual-Fail-Safe-Design, in which two copies of a self-checking circuit are combined to provide tolerance to failures in a single copy of the circuit. We note that the technique can be extended to provide N-fault-tolerance.

We have described an improved memory organization which enables self-checking microprogrammed controllers to be implemented using current and foreseeable semiconductor technologies.

Finally we have used such a self-checking controller as the basis of a comparative evaluation of DFS and TMR as fault tolerant techniques. In this example, DFS yielded twice the lifetime enhancement available from TMR at some 20% to 30% lower cost. We conclude that DFS is a useful innovation in the design of highly reliable systems.

7 Acknowledgments

The work reported in this paper was funded by the European Space Agency under ESTEC contract no. 2258/74 AK. Technical management by M. A. Perry, ESTEC, Nordwijk, Holland.

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Manuscript received by the Institution on 1st December 1976. (Paper No. 1785/Comp. 183)

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UDC 681.3.06

Developing interactive systems for the small machine environment

GILL RINGLAND, M.Sc., M.B.C.S.*

Based on a paper presented at the Conference on Computer Systems and Technology held at the University of Sussex from 29th to 31st March 1977

SUMMARY

This paper describes the design methods, and tools, for developing interactive systems which will run in the small machine environment. The constraints of this environment are discussed, and it is shown that the requirements of users in this environment are rather stringent, but for systems with a well engineered interface rather than for high performance. The solution adopted is to provide realistic simulation early in the design process, and to provide mechanisms for divorcing the detailed screen layout design from program design.

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1 Introduction

In the first fifteen years of computing, computers were the expensive tools of mathematicians. In the second fifteen years, most belonged to commercial data processing departments. In the third fifteen years, computers will migrate into offices and factories and be used by non-experts.

Certainly the first two sentences are true, and we have the hardware capability to make the third true. But the ability of us—the computer 'experts'—to design systems for the non-experts has developed less rapidly, and we are in many ways still designing systems to meet the constraints and user demands of the last fifteen years.

This paper describes a technique, set in the context of the CAP on-line design methodology, which explicitly uses the processing power available, to improve the flexibility of the user interface. This is especially important for non-expert users who often, for reasons we will discuss, use small computers outside the normal data processing environment.

2 The Small Computer Environment

Since everyone's definition of a small computer is different (and these definitions are changing rapidly), small computers will here be categorized primarily by their typical users. The range of typical users is a direct consequence of the price and performance band being considered.

Some small machines are the bottom end of an upwardly compatible range—such as IBM 370/115 or Honeywell and (it seems likely at the time of writing) the ICL 2903. Others come from the minicomputer manufacturers—DEC, Data General, etc. Others are 'single user' machines, like the IBM S/32, the Hewlett Packard 9830. The price range of these machines is large—from £5,000 to £150,000—but they have in common their extensive use by 'first time users' in relatively small businesses or in user departments of large organizations. Many small businesses have used computer bureaux or visible record computers to perform standard functions, such as payroll, or accounting in the past. There are clear signs that these companies are now acquiring small computers not just to do these standard functions in-house, but primarily for applications which are *central* to their business.

So, for instance, a distribution company will want to handle scheduling and the production of order and delivery notes by computer. A travel agency will incorporate a computer-based register of holidays. A motor retailer will use a computer system for stock control and spares ordering. These examples show that, in each case, the computer is being used by people whose skills and experience are in areas other than computing, as a tool to aid them in their work. These staff will know the operational constraints and special conditions of the company, and will expect the computer system to reflect these.

These users are therefore very different from users such as data entry operators, in that they have a *range* of tasks to perform and high expectations from the system. The systems analysis for such systems is lengthy because the business requirements tend to be specialized and complex. Providing a system to meet these demands is likely to be expensive and time consuming, and when the system is built, the maintenance environment is different from that we are used to. The users are also likely to be remote from a DP department, so the software must be reliable and the user interface must be well engineered. Thus, the emphasis is now not the machine *per se*, but the commonality among the users' requirements.

The main topic of this paper is the methods used in designing and implementing the software to ensure that the user interface remains flexible, and can be altered. This can be achieved without reprogramming in many cases, in other cases through introducing new and separate sub-systems written and tested elsewhere.

This paper therefore does not cover in any detail the important areas of file design for on-line systems, or testing interactive systems, since these are not so particular to the small machine environment. It is also not concerned with techniques for packaging code for use in similar applications and systems, or with high-level 'program generators'. These techniques are being widely introduced in the small systems area, as an intermediate step between completely standard packages (e.g. payroll) and completely tailored software, and are used in tandem with the design techniques which are discussed below.

3 Dialogue Design

It is clear from the preceding discussion, of the users of small machines, that their needs can best be met by an interactive system. In some cases the need is for information retrieval, as for the travel agents. In other cases, information is provided either to add to the data base or to enable documents, such as the delivery notes, to be produced. While the time scale is different for these different needs, the common factor is that the user will engage in some sort of dialogue with the computer.

The principles of structuring this dialogue, so that at each stage of the processing the user is presented with a list of options for the next step (the 'menu') is now well established (see, for instance, ref. 1). The design process for the dialogue, within this structure, can be considered at several levels.

Using a top-down approach, the highest level is that at which interactions are enumerated. Each interaction consists of the output of a message by the computer, and the operator reply. At this stage, the content of the message is not specified in any detail, e.g. 'help text' or 'start' is defined.

† There are many incompatible terminologies in use. The author hopes that this definition of transaction will be acceptable for the purposes of this discussion.

A meaningful unit of work within the system will consist of a sequence of interactions, called a transaction.† The next level of design is to group the interactions into transactions, and identify all the possible responses at each step. At this level, each screen format is given a unique identifier and the data which will appear on it (by output from the computer or input by the user).

The discussion to date has been independent of the I/O device used for interactions with the system. In the discussion below, however, we will assume that v.d.u. screens are the I/O device. This is because, in practice, they are becoming the standard device. The reasons for this are the relative speed advantage they have over mechanical devices, which increases the information that can be given to the user at each stage. The use of the whole screen (typically 12–24 lines of 50–80 characters) enables the operator to scan information visually, and to be guided effectively by detailed instructions from the system. These advantages are thought to outweigh the disadvantages of the lack of hard copy, as long as system designs make additional provisions for the logging of the message in this environment.

At the lowest level of design then, the detailed screen layouts are defined. The use of the term 'lowest level' does not imply in any way that this is a less skilled process than the system design. It is clear that different skills are required, namely the 'human science' skills of identifying how best to guide the users through the system. 'Computer people' may find it difficult to anticipate the problems of the first-time user. But, providing that the essential prerequisite of a reliable system has been achieved, the content and layout of the messages will be the aspect of the system most directly affecting the users.

Therefore, in this area we have found it is important to provide flexibility. The technique used to achieve this involves separating the program specification and coding completely from the screen layout definition process. Screens are defined by means of an off-line program which produces a table. This table is then accessed at run-time by the on-line system, to write messages and to format data, in a process called run-time mapping.

4 A Sample Application

The discussion so far has outlined the approach to designing interactive systems. We consider in more detail below the aids available at each level, the information which is defined at each level (and hence the degree of flexibility) and discuss the implications for program structure. To make this discussion more concrete, an example is given, based loosely on a personnel records system.

The system has to allow staff to be added to the records, or deleted from it. Access for addition and deletion of information needs to be controlled, by passwords or some other method. And the amount of guidance and text given in the use of the system needs to be variable, so that as staff become used to the system they bypass the prompts and help sequences and key in information directly. It is necessary to be able to access staff directly by name, or by skills (e.g. which are our Coral programmers?). And occasional hard copy summary reports are required—for instance, the monthly staff list.

5 Interaction Inter-Relationships

The sample application (the personnel records system) can be described at the highest level by one chart, the Interaction Interrelationship (II) chart (Fig. 1).

In this chart, every message output by the computer is represented by a rectangle. Connecting lines show different types of operator reply. The importance of this level of chart is in identifying the units of processing, as well as the interactions—one interaction processor subprogram (IP) will be implemented for each message rectangle. This equivalence makes the estimating of effort required a low-error process, since in practice IPs are found to cover a fairly narrow size range (e.g. 5 to 8 kbytes).

The specification of one IP for each interaction provides a structure for the application, so that even staff without on-line experience can implement systems. It is interesting to note that each IP needs to have the *inverse* structure to a batch program. Whereas a batch program conventionally reads data, performs some processing and ends by producing some output, an IP consists of two logically distinct parts: the output function, followed by the input function. The output function is initiated by the previous IP, and sets up the message to the user on the screen. He then reacts, for instance by keying in data, and this is analysed and processed by the input function. The input function then ends by passing control either to a supervisor, or directly to another IP.

Without the imposition of a structure, it is very difficult for inexperienced application programmers to write online code: this particular structure will be shown to relate closely to the techniques used at lower levels of the system design.

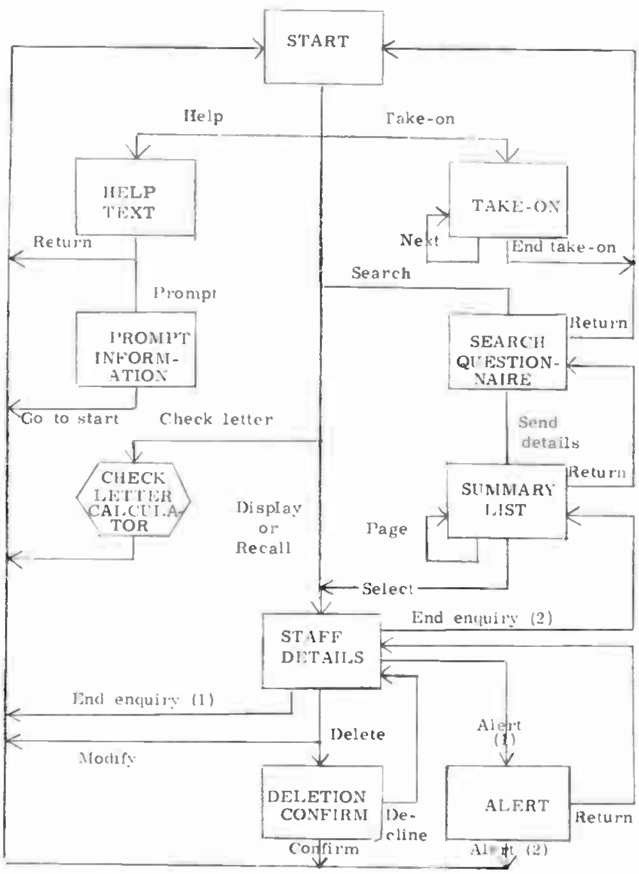


Fig. 1. Personnel records interaction interrelationship chart.

Notes: The return, end and decline functions correspond to the RETURN action sequence. The delete and alert (1) functions correspond to the DELETE and ALERT action sequences respectively. The remaining functions relate to the use of the SEND key.

6 Transactions

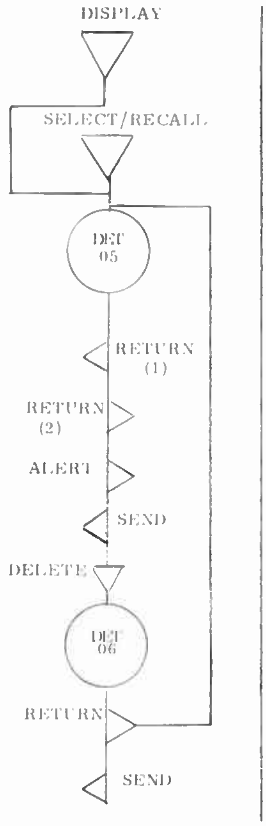
Once the interactions and their interrelations have been specified, the system transactions can be identified. These will correspond to different applications—in the example, 'search' transactions and the 'take-on' (of new staff details) are obviously distinct. Consider, however, the transaction shown in Fig. 2, which is called a 'terminal procedure' chart. This inspects and optionally modifies staff details. This transaction contains parts (the modification for instance) which are privileged. It would have been possible to identify instead two transactions, accessible by different classes of operator; one to inspect staff details, and a separate one able additionally to modify staff details. Thus, there is a measure of judgement and convenience in the choice of transaction definitions.

The convention used on these charts is simple—messages are shown as circles and small triangles are operating replies. Large triangles are used to mark connectors to other parts of the system, for instance messages output by other transactions.

It will be noticed that the validation of data is not shown, either on the II chart or on the terminal procedure chart. This is because a standard mechanism for handling errors is always used, which has the effect of returning to the position before the invalid data was entered, for simple errors. For errors which need to be detected in the application—for instance, by reference to data held on files—control is also returned to the point at which the user is requested to enter data. Thus in all cases there is an *implied* loop, back to the user.

The other point worthy of comment is that the terminal procedure chart shows neither file accesses nor processing, although it is clear from the earlier discussion that one IP is needed between each circle representing a screen of data and messages.

The screens are identified by a reference number on the terminal procedure chart. This reference number is also used on the interaction description form (see Fig. 3 for example) to identify the data which is used in this message. It will be noticed that the text is not specified at this time,



The operator enters 'D' as a transaction code on the start screen;

or may have selected a record during the search transaction or may have recalled a record.

The staff details screen is output, containing public information and, if the operator is privileged, private information.

The operator may terminate the enquiry if display transaction - or indeed the recall transaction - see note 1;

or return to the summary screen during a search transaction;

or alert another operator;

or, if privileged, modify any details except staff code, (see note 2);

or, if privileged, request the record to be deleted, in which case...

The deletion questionnaire is output.

The operator may decline the deletion by returning to the staff details screen;

or confirm by keying the word 'DELETE', (see note 3).

Fig. 2. Inspection and modification.

Notes:

1. The start screen is displayed with a completion message stating that the display of staff member, nnna, is complete.
2. The start screen is output with a completion message stating that the record for staff member, nnna, has been modified.
3. The start screen is output with a completion message stating that the record for staff member, nnna, has been deleted.

nor is the layout yet defined. This form is the reference document used by programmers in defining the data input to, and output by, their programs.

The interaction description form has two parts, variable fields and replies. The types of variable are O (output), I (input) and U (output for optional updating and re-input). The replies listed correspond to the names entered on the terminal procedure chart, and the entry in the NEXT column is an interaction name corresponding to a screen shown on the terminal procedure chart.

The methodology described so far enables the design to be analysed from the user's point of view, isolating unnecessary message sequences and ensuring that the data needed for processing are specified to the program. The next Section considers how, within this structure, the detailed user interface can be implemented independently of the program development.

7 The Video Format Program

Decoupling program design from the design of the detailed screen layouts involves two steps. The first step uses an off-line utility, the video format program, to produce a table specifying the data and its layout on the screen, for use by IPs. The table is then included with the IP (by link-editing or use of a COPY library) and drives the code sequence at run-time.

The data supplied to the video format program consist of a layout—essentially a mock-up of the screen as the user sees it—and a set of descriptors for each field. This layout can be used, in conjunction with the terminal procedure chart, to talk through the system with the users at an early stage of the project. This has proved to be an effective way of enlisting the support (and help) of at least the senior staff involved in getting the design right. For a good discussion of this neglected area see ref. 2.

The program can produce modules for use during testing —when hard copy devices are used—or for live running. In either case, the layouts correspond to the formats which will be actually seen by the user.

The descriptors for each field connect the data format details with the programming details. The data format details cover such information as the type (I, O, U or, constant), the format (e.g. fixed or variable length) whether the field may be repeated or suppressed and which standard validation checks are to be performed. The

Interaction Description

Reference
Page
Version Date
Author

NAME DEL   MODULE: 06

TITLE Deletion confirmation

TYPE (RNAME)	DESCRIPTION	NOTE (NEXT)	COMMENTS
0	Staff code		
I	Deletion confirmation		The word 'DELETE' must be keyed.
REPLIES			
SEND & 'DELETE'	Confirm deletion	ZST (1)	The completion message 390 is output.
RETURN	Decline deletion	DET	The deletion request is declined by the operator

Fig. 3. Interaction description form.

Note 1. Message 390 states that the staff member, nnna has been deleted from the file.

program label, by which the field is known in the IP, is also specified to the video format program.

This enables the layouts to be changed without re-programming, and in fact it has been found that the procedure can be followed by completely untechnical staff. This has obvious advantages in this environment where there may be no DP department available.

This technique of providing a table to control the run-time use of the screens is called run-time mapping. It has of course, one disadvantage—like all essentially interpretive schemes, it uses more computer processing than tightly bound or compiled schemes. Historically this has been an overwhelming disadvantage. However, the decreasing cost of hardware and the increasing cost of people means that the balance is now very firmly in favour of using processing power in this way. This technique is thus seen as one small part of a larger trend—at last, to use computers to improve the quality of life.

8 Conclusions

The techniques described in this paper have been widely used within CAP on numerous projects. It has been shown that they are particularly applicable to the small machine environment, through two main factors. The first is the overall emphasis on dialogue design and the consequent structuring of the programs. The second is the use of run-time mapping to allow operational flexibility in the user interface.

9 Acknowledgments

The author would like to stress that the ideas and techniques described in this paper have been the result of work by many colleagues in CAP, in particular Esmond Hart and John Johnston. Any errors or omissions in the paper are however entirely the author's.

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The following two references provide the background for this work and are included for completeness:

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Manuscript received by the Institution on 30th November 1976.
(Paper No. 1786/Comp. 184)

UDC 621.372.851/2

A method for the determination of the T-equivalent-circuit parameters of microwave structures

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SUMMARY

A method is described for the measurement of the equivalent circuit parameters of waveguide-mounted structures, such as a post or an inductive iris. Experimental results for post structures centrally mounted in X-band waveguides are also given. The values obtained are found to agree closely with those predicted from theory and the estimated accuracy is about 5%.

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1 Introduction

The active devices in solid-state microwave oscillators and amplifiers, such as Gunn diodes and avalanche diodes, are often post-mounted in waveguides. The equivalent circuit parameter values for the post structure are required to be known for designing the oscillator or amplifier properly. These values may be theoretically calculated[1–3] and experimental values for some particular dimensions of the post are also given by Marcuvitz[4] for some wavelengths. However, the values depend critically on the dimensions and location of the post and the frequency of operation. The values are also significantly affected by the characteristics of the bias filter which is connected at one end of the post. For the evaluation of the dependence of the performance characteristics of an oscillator or amplifier on the post parameters, directly measured values are useful. The method used for obtaining the parameter values quoted by Marcuvitz was not described by him. Apparently, the results were obtained by Pickering *et al.*[5] using a method, the description of which has been given by Huxley.[6] Essentially, the same method has been used by White,[7] to determine the post reactance for a Gunn diode oscillator, neglecting the effect of the series reactances.

We describe in this paper a new method for the determination of the reactive as well as the resistive parameters of an inductive obstacle in a waveguide. In order to check the suitability of the method, we have measured the parameters of two posts centrally mounted and completely spanning an X-band waveguide at two frequencies. These results are presented in Section 3 along with the theoretical values given by Marcuvitz[4] for the same structures.

2 Method of Measurement

The equivalent circuit for the post used is shown in Fig. 1. X_a is the normalized shunt reactance (usually inductive) produced by the post and X_b is the normalized series reactance (usually capacitive) introduced by the delay action due to the non-zero diameter of the post. The normalized resistance R takes into account the losses associated in series with the inductance X_a of the post. Since the structure has symmetry the series capacitors are equal and we are required to determine three circuit parameter values.

In the method of measurement described in this note the post structure was terminated by a precision short

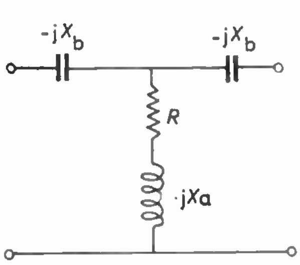

Fig. 1. Equivalent-T-circuit parameters of a post.

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(Microwave Components Type 16/11) and the input impedance of the structure was determined by a bridge, shown in Fig. 2, employing a precision attenuator (Hewlett Packard X382A) and another precision short of the same type as the first one. The input impedance was determined for different lengths of the terminating precision short and the data so obtained were used as discussed below for the determination of the circuit parameter values described earlier.

The input impedance of the terminated post structure at the post plane (determination of its position is discussed later) is given for the assumed equivalent circuit by the following expression,

$$Z_{in}=-jX_b+\frac{[-jX_b+j\tan\beta(d+l_b)][R+jX_a]}{[R+jX_a]+[-jX_b+j\tan\beta(d+l_b)]}, \qquad (1)$$

where $\beta=2\pi/\lambda_g$, λ_g being the guide wavelength and $\tan\beta(d+l_b)$ is the reactance presented by the precision short at the plane of the post, d being the distance of the precision short from its front face and l_b is the distance of the rear face of the post structure from the post plane (see Fig. 2). The resistive and reactive components of the input impedance are, respectively,

$$R_{in}=\frac{R(1-x/X_b)^2}{R^2/X_b{}^2+(1-X_a/X_b-x/X_b)^2} \qquad (2)$$

and

$$X_{in}=-X_b\left[1+\frac{(1-x/X_b)\{R^2/X_b{}^2-X_a/X_b(1-X_a/X_b-x/X_b)\}}{R^2/X_b{}^2+(1-X_a/X_b-x/X_b)^2}\right] \qquad (3)$$

where

$$x=\tan\beta(d+l_b). \qquad (4)$$

We find from equation (2) that R_{in} would have a maximum value of

$$R_{in(max)}=R[1+X_a{}^2/R^2] \qquad (5)$$

and from equation (3) that

$$X_{in(x\to\infty)}=X_a-X_b. \qquad (6)$$

Considering the expression for X_{in} we also find that as x varies from $-\infty$ to $+\infty$, X_{in} will change from an asymptotic value of $X_{in}(x\to\infty)$, reach a high value at $x=-(X_a-X_b)$, then change sign and become zero at

$$x\approx\left[\frac{1-2X_a/X_b}{1-X_a/X_b}\right]\cdot X_b,$$

for small values of R, then rise asymptotically to the value $X_{in(x\to\infty)}$. It can also be shown (see Appendix) that the plot of $[X_{in(x\to\infty)}-X_{in}]^{-1}$ against x would be a straight line near $x=0$. The slope of this straight line would be

$$m=\frac{1}{X_a{}^2}\left[1-\frac{R^2}{X_b{}^2}\cdot\frac{2X_a/X_b-1}{(X_a/X_b)^2(1-X_a/X_b)^2}\right]. \qquad (7)$$

The features of the input impedance noted above may be used to obtain the position of the post plane and the equivalent circuit parameters of the post as discussed in the following section. The unknown parameters X_a, X_b and R may be obtained, in principle, from the measured values of R_{in} and X_{in} for some chosen values of x. However, the analysis is much simplified and the errors are also smoothed out if we use equation (7).

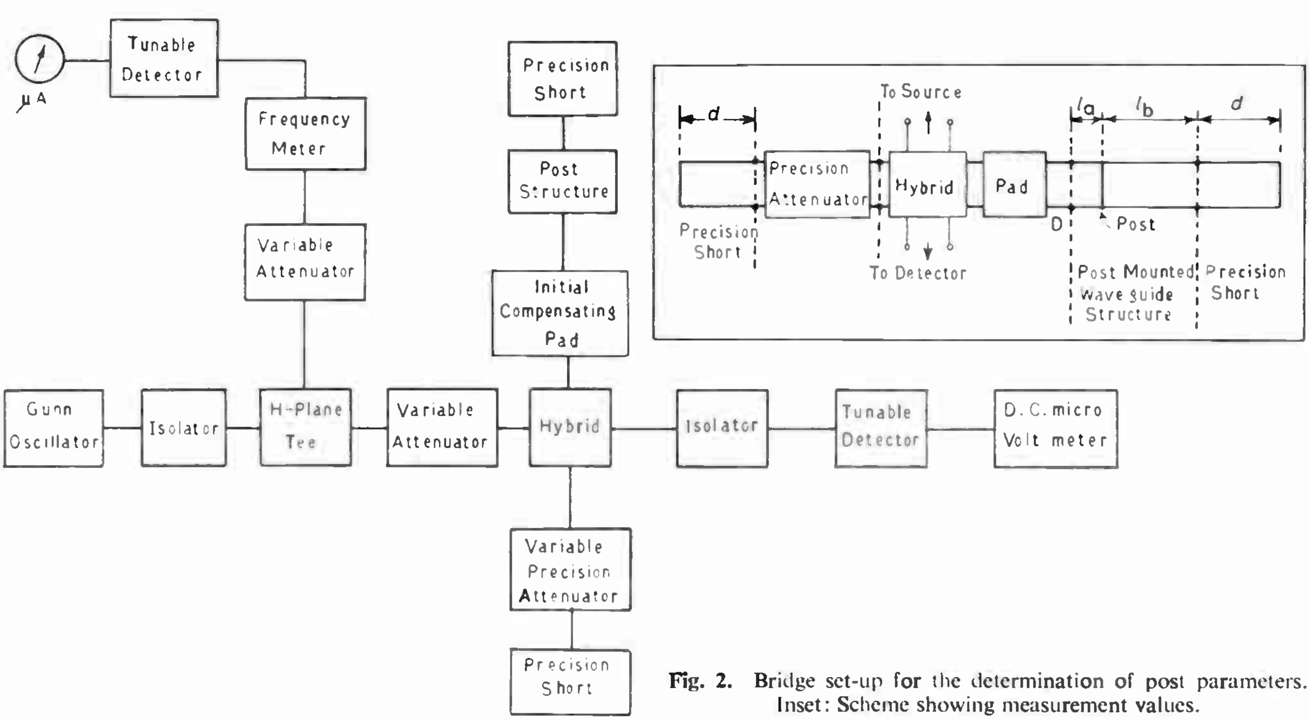

Fig. 2. Bridge set-up for the determination of post parameters. Inset: Scheme showing measurement values.

2.1 Determination of the Position of the Post Plane

The position of the post plane is often determined by mechanical measurement. However, as the post has a non-zero diameter and it may not be exactly vertical and also since other structures are connected at one end of the post, the location of the post plane by mechanical measurements may introduce errors. We may determine the post plane from electrical measurements using the property that $X_{in(x\to\infty)}$ attains an asymptotic value. This would mean that if d_1 (see Fig. 2 inset), is plotted against d the curve should show an asymptotic value for d_1. In Fig. 3 we present some measured values and indeed we find that for some values of d, d_1 remains constant. If we now take two sets of measurements with the two faces of the post structure terminated by the precision short in the two sets, the difference in the two asymptotic values of d_1 would be equal to the difference in the distances of the post plane from the two faces of the post structure, l_a-l_b (see Fig. 2). The total length of the post-mounted structure (l_a+l_b) may be determined with precision mechanically or electrically (before inserting the post). The distances l_a and l_b are evaluated from these measured values of (l_a-l_b) and l_a+l_b.

2.2 Determination of X_a, X_b and R

With the position of the post plane determined as discussed above, the input impedance of the post structure at the post plane was calculated from the measured values of the attenuation, α, and length of the precision short, d_1, using the following equations,

$$R_{in}=\frac{1-\exp(-4\alpha)}{1+2\exp(-2\alpha)\cos 2\beta s+\exp(-4\alpha)} \tag{8}$$

$$X_{in}=\frac{2\exp(-2\alpha)\sin 2\beta s}{1+2\exp(-2\alpha)\cos 2\beta s+\exp(-4\alpha)} \tag{9}$$

where $s=d_1-(d_0+l_a)$; d_0 is the value of d_1 for a null in the bridge with the post structure replaced by a short circuit and l_a, as defined earlier, is the distance of the post plane from the front face of the post structure.

We show in Figs. 4 and 5 experimental plots of X_{in} and R_{in} respectively against the values of x calculated by using equation (4). We find that the nature of the plots agree very well with the expected behaviour which was discussed earlier. This may be taken as an experimental justification of the equivalent circuit used in the analysis. In Fig. 4 is also shown a plot of $[X_{in(x\to\infty)}-X_{in}]^{-1}$ against x, which again is a straight line near $x=0$. The values of R, X_a and X_b are now evaluated using the values of $R_{in(max)}$, $X_{in(x\to\infty)}$ and the slope m given in equations (5), (6) and (7) respectively. As R is a small quantity these equations were solved by successive approximations. The R-dependent term in equation (3) is first neglected and X_a and X_b are obtained from equations (6) and (7). The value of X_a is used in equation (5) to obtain R. The corrected values of X_a and X_b are then obtained from equations (6) and (7), using this value of R.

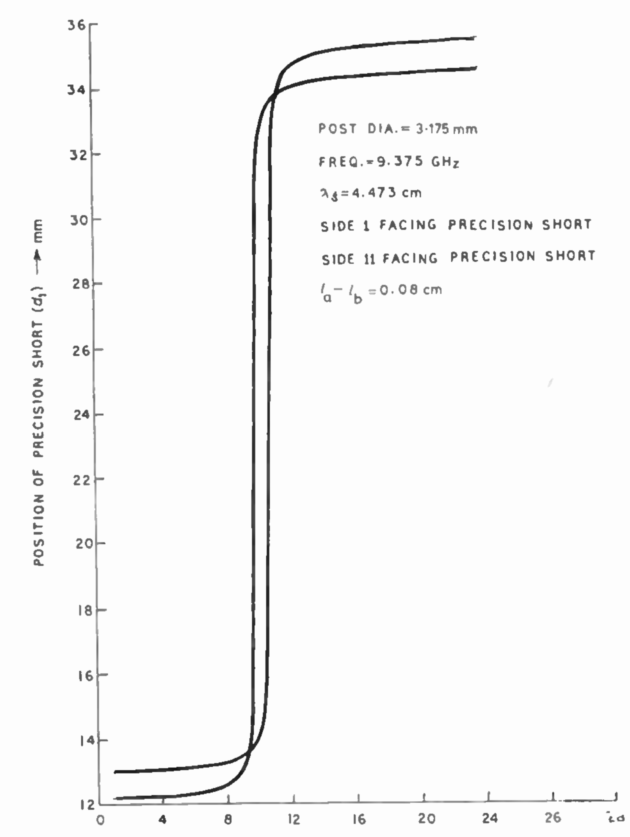

Fig. 3. Variation of the position of precision short, d_1 with the position of precision short, d.

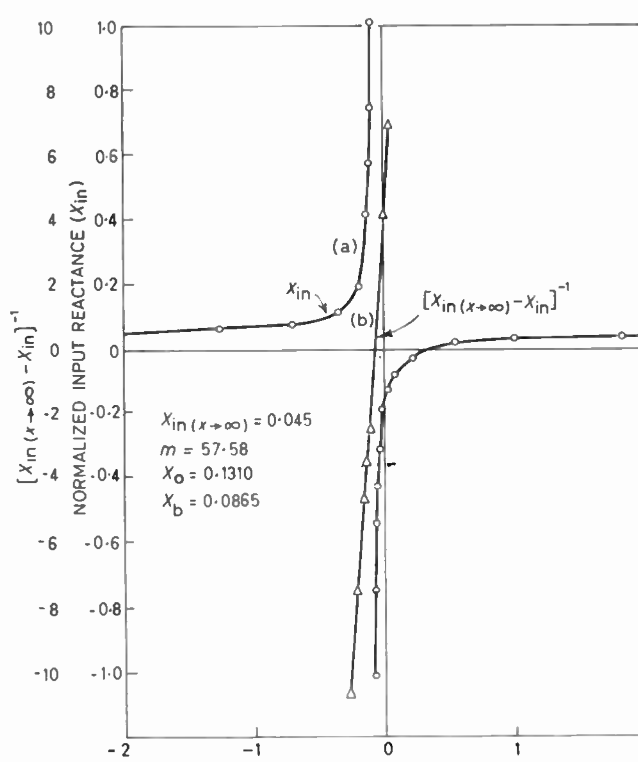

Fig. 4. Experimental plots showing the variation of (a) the normalized input reactance, X_{in}, (b) $[X_{in(x\to\infty)}-X_{in}]^{-1}$.

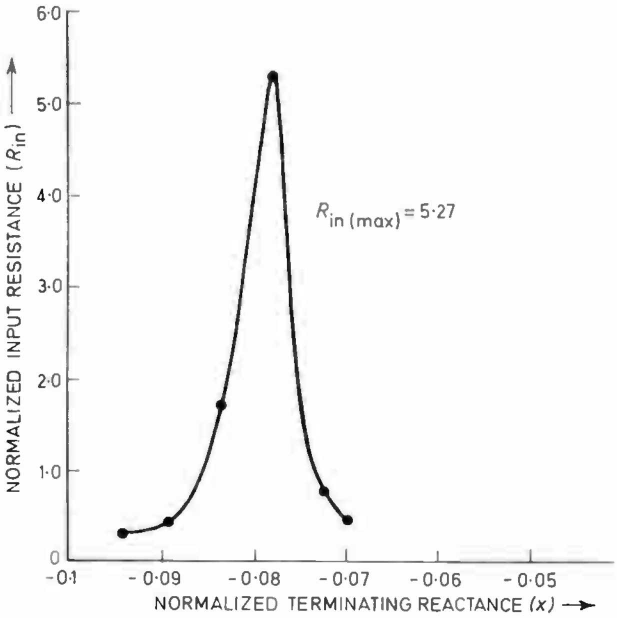

Fig. 5. Variation of the normalized input resistance, R_{in}, against the normalized terminating reactance, x.

3 Experimental Results

We have applied this method for determining the circuit parameter values for different dimensions of a centred post completely spanning an X-band waveguide at different frequencies. The results for two different diameters of the post at two different frequencies are given in Table 1 along with the theoretically computed values.[4] We find that the experimental values agree with theory quite closely.

Table 1. Experimental and theoretical values of post parameters

Post diameter (mm)	Frequency (GHz)	X_a Experiment	X_a Theory	X_b Experiment	X_b Theory	R Experiment
3·175	9·375	0·131	0·122	0·087	0·089	0·0032
	10·000	0·161	0·144	0·107	0·101	0·0030
1·5875	9·375	0·294	0·273	0·019	0·024	negligible
	10·000	0·327	0·312	0·022	0·029	negligible

4 Accuracy of the Method

The accuracy of the method is essentially determined by the accuracy of the precision short, symmetry of the magic-T and change in phase-shift introduced by the precision attenuator. We checked the effect of the magic-T symmetry by interchanging the unknown and known impedance arms. No measurable change in values was detected, which lead us to conclude that the error contribution of asymmetry, if there be any, was insignificant. If, however, the asymmetry is found to be significant under certain conditions, one may use the method of van Iperen and Tjassens[8] to eliminate the errors arising from this source. In order to assess the effects of attenuator phase shift and of the error in precision short lengths, we calculated the parameter values assuming an error of about 0·02 mm in the position of the shorts and a change in attenuator phase-shift of 0·5 degree for a change in attenuation of 2 dB. The final values were found to be altered by less than 5%. We also made measurements on different dates and found that the values of $X_a - X_b$ and X_a agreed to within 5%, whereas those of X_b were within 10%.

5 Conclusion

The method described here is based on very simple measurements, requiring two precision short circuits, a precision attenuator and a null detector such as are commonly available in a laboratory. Although results are given for post structures, the method should be suitable also for other structures for which a T-equivalent circuit representation is possible.

6 Acknowledgment

The authors acknowledge with thanks the support provided by the Ministry of Defence, Govt. of India and the free computer service of the Computer Centre, Calcutta University. One of the authors (B.R.N.) acknowledges the support of the Jawaharlal Nehru Memorial Fund.

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8 Appendix: Derivation of equation (7)

Combining equations (3) and (6) we obtain

$$[X_{in(x\to x)} - X_{in}]^{-1} = \frac{R^2 + (X_b - X_a - x)^2}{R^2(X_a + X_b - x) - X_a^{\,2}(X_b - X_a - x)}. \quad (10)$$

The slope, m of $[X_{in(x\to\infty)} - X_{in}]^{-1}$ against x, is obtained by differentiating equation (10) with respect to x. We have thus:

$$m=\frac{1}{X_a^2}\left[\frac{2}{1-\frac{R^2}{X_a^2}\cdot\frac{X_a+X_b-x}{X_b-X_a-x}}-\frac{\left(1+\frac{R^2}{(X_b-X_a-x)^2}\right)\left(1-\frac{R^2}{X_a^2}\right)}{\left(1-\frac{R^2}{X_a^2}\cdot\frac{X_a+X_b-x}{X_b-X_a-x}\right)^2}\right]. \tag{11}$$

Near $x\to 0$, equation (11) gives

$$m_{(x\to 0)}=\frac{1}{X_a^2}\left[\frac{2}{1-\frac{R^2}{X_a^2}\cdot\frac{X_a+X_b}{X_b-X_a}}-\frac{\left(1+\frac{R^2}{(X_b-X_a)^2}\right)\left(1-\frac{R^2}{X_a^2}\right)}{\left(1-\frac{R^2}{X_a^2}\cdot\frac{X_a+X_b}{X_b-X_a}\right)^2}\right]. \tag{12}$$

Equation (12) may be re-arranged and written as,

$$m_{(x\to 0)}=\frac{1}{X_a^2}\left[1-2\cdot\frac{R^2}{X_a^2}\cdot\frac{X_b+X_a}{X_b-X_a}-\frac{R^2}{X_a^2}\cdot\frac{X_b(2X_a-X_b)}{(X_b-X_a)^2}+\frac{R^4}{X_a^2(X_b-X_a)^2}\right]\cdot\left[1-\frac{R^2}{X_a^2}\cdot\frac{X_a+X_b}{X_b-X_a}\right]^{-2}. \tag{13}$$

Since $R^2/X_b^2 \ll 1$, the last term may be expanded binomially. Considering only the first order terms we thus have

$$m_{(x\to 0)}=\frac{1}{X_a^2}\left[1-\frac{R^2}{X_b^2}\cdot\frac{2X_a/X_b-1}{(X_a/X_b)^2(1-X_a/X_b)^2}\right]. \tag{14}$$

Manuscript first received by the Institution on 27th July 1976 and in revised form on 1st February 1977. (Paper No. 1787/CC 276)

UDC 621.396.62:621.372.56

Automatic input attenuator for h.f. communication receivers

DURK VAN WILLIGEN*

SUMMARY

The final selectivity of a communication receiver is generally determined by the bandwidth of the i.f. filters. When the a.g.c. of the receiver is determined only by the signals passing these filters, severe overloading of the relatively broadband h.f. gain stages may occur.
Reducing this risk of overloading requires a broadband attenuator at the input of the receiver. The attenuation should be dependent on the signals passing the i.f. filters as on all other signals reaching the h.f. gain stages, but not the i.f. amplifiers.
Such an automatic attenuator system is discussed, and its design philosophy presented.

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1 Introduction

There can be a great difference in amplitudes of the wanted and unwanted signals for frequencies in the h.f. band. If the wanted signals are strong, there is no problem. But in the opposite case there may arise real problems if the power level difference between the wanted and the unwanted signals exceeds the dynamic range of the receiver's front end, as non-linear effects will be introduced.

A possible definition of the dynamic range of a receiver is:

$$DR = P_{\max}/kTBF,$$

where $P_{\max}$ is the maximum input power level generating the maximum permissible amount of intermodulation product level, kTB is the available input noise power, and F is the noise figure of the front end. If an antenna is connected to the receiver, the noise output of the antenna should be just over that of the noise power level of the receiver itself. Only then are we able to use the full dynamic range of the receiver.

If the antenna noise exceeds the noise of the receiver, a shorter antenna can be applied or an input attenuator can be placed between the antenna and the input terminal of the receiver. The latter is a more elegant way, because the amount of attenuation is easy to vary. The same is valid if the received wanted signals are stronger than the receiver noise.

An operator can determine the optimum attenuation at the input by trial and error. Too much attenuation at the input terminal will cure the overloading, but may also decrease the signal-to-noise ratio of the wanted signal to an unacceptable level. If the attenuator is to react automatically, the problem is where the information for control is to be derived. The i.f. signal gives only information about the wanted signal, but not about all other signals reaching the front-end stages. If the information is taken from the front-end signals, then it is valid if we want only to prevent overloading the front end. But there is no information about the resulting signal-to-noise ratio of the wanted signals. A better result can be achieved by using both sets of informations to optimize for best reception.

A unit to control the input attenuator and the i.f. gain would find application in the unattended receivers of automated h.f. radio links. The design described in this paper takes into account the overloading risk of the front end of the receiver, the signal-to-noise ratio and the type of modulation (c.w., a.m., s.s.b. or f.m.) of the wanted signals.

2 Design

The automatic input attenuator unit controls two gain control circuits. The first is the wide-band input attenuator, the second the in-band i.f. gain control. The latter is well-known, and in most receivers is externally controllable. Therefore, we will only discuss the wide-band input attenuator.

Many devices are available for gain control purposes: diodes, transistors, p-i-n diodes, f.e.t.s, reed switches, etc. All the linear devices behave rather poorly under strong signal conditions and the only suitable devices for our purpose are diode or reed-switched attenuators. They are discrete attenuation devices, but their strong signal behaviour is excellent. We must keep in mind that it should be possible to use receivers on board aircraft and ships, where the front ends of the receivers are tortured by strong transmitting signals coming from the same craft and the attenuator must be able to withstand at least 10 V r.m.s. out of 50 Ω.

In our experimental automatic input attenuator, we used three reed-switched attenuators with values of 6 dB, 12 dB and 24 dB, respectively, Logic control can give every attenuation between 0 and 42 dB in steps of 6 dB. The value of 6 dB is a compromise between many steps, with a small increase of attenuation per step and only a few steps with inherently a higher step value of attenuation. Small step values allow an accurate gain control of the front end, but it may cause frequent stepping of the attenuator, and this will disturb the received signal too much. The stepping attenuator is a switching device that can cause problems in the in-band i.f. automatic gain control because it takes some time for the gain of the i.f. amplifier to be regulated to the correct value after a change of input attenuation. An elegant method of over coming this problem is to gate the audio signal off until the i.f. a.g.c. is stabilized, which can be accomplished within 10 ms. Audio interruptions of 10 ms are hardly perceptible if they occur at random.

We use a wide-band amplifier, followed by an amplitude detector, in parallel with the front end of the receiver. The output of the detector is fed into the two level detectors, a and b in Fig. 1. They can detect the level P_{max} and a level 9 dB lower (the $P_{max-9\ dB}$ level). These level detectors give information about the amount of overloading of the front end.

A second amplifier with fixed gain is located in parallel with the normal i.f. amplifier at the output of the i.f. bandwidth limiting filter and is also followed by an amplitude detector. The output of the detector goes to the level detectors c and d, Fig. 1. When the receiver input terminal is disconnected from the antenna, the output power N from the i.f. bandpass filter is the noise internally generated in the front end of the receiver and amplified by the i.f. stages. When the receiver is operating, the output from the i.f. filter is a combination of the front-end generated noise, received atmospheric noise and the wanted signals. This combined signal is called S. While N is fairly constant for a given type of receiver, it is possible to measure S/N by measuring S only. Level detector d detects the minimum allowable S/N ratio, and detector c indicates a level 9 dB over $(S/N)_{min}$, or $(S/N)_{min+9\ dB}$.

Both amplitude sensors are followed by two level detectors having an amplitude difference of 9 dB. That is one and one-half step value of the above mentioned attenuator. This allows us to introduce the necessary hysteresis, as will be explained in the discussion of Table 1.

Table 1.

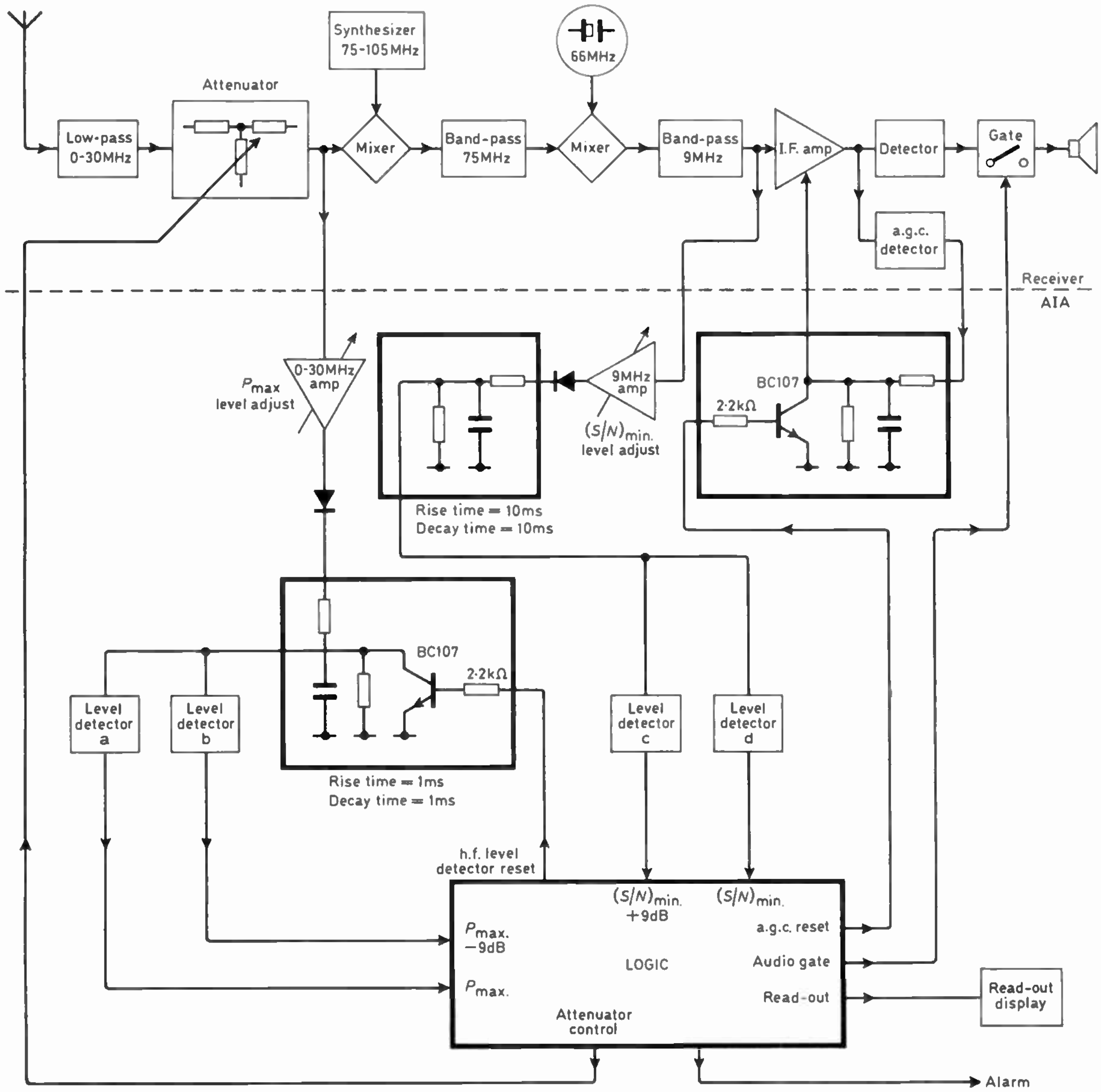
Output of level detectors a	b	c	d	Combination
0	0	0	0	1
0	0	1	0	2
0	0	1	1	3
1	0	0	0	4
1	0	1	0	5
1	0	1	1	6
1	1	0	0	7
1	1	1	0	8
1	1	1	1	9

The resistive input attenuators are switched by reed relays whose switching speed is adequate for use in h.f. communication receivers with a relatively narrow bandwidth. The switching is controlled by a digital up-down counter which is commanded to increase attenuation at 'on' pulses and to decrease attenuation at 'off' pulses. An 'on' pulse (logical 0) is generated if the input level is over the P_{max} value, and the i.f. S/N ratio can be decreased without going below the $(S/N)_{min}$ level. An 'off' pulse (logical 1) is given if the input level is less than $P_{max-9\ dB}$ or when the i.f. S/N falls below the pre-determined minimum level $(S/N)_{min}$. (See also Table 1.)

Problems may arise with the choice of the reaction time of the decision. Overloading of the front-end circuits of the receiver can also be due to spikes etc. They are characterized by short duration, seldom longer than 10 μs. This type of noise overloading can be handled best by the so-called 'pre-i.f. noise gating' techniques. In our design overloading of the input circuits must exist for at least 10 ms before an 'on' pulse will be generated.

If the overloading situation ceases, a longer time should be taken before generating an 'off' pulse. One must be sure that there is not just a gap in a c.w. or s.s.b. interfering signal. The release time for decreasing input attenuation is chosen to be 1 second if the receiver is used in the c.w. or s.s.b. mode. In the a.m. or f.m. mode, 'off' pulses can be generated within 100 ms if the i.f. S/N ratio falls below the minimum acceptable level $(S/N)_{min}$. (See Table 1.)

The switching action of the input attenuator will give unwanted interaction with the normal a.g.c. system of the i.f. amplifier. In general the rise-time of the i.f. a.g.c. system is short enough to follow the decrease in input attenuation. But when the attenuation is increased, the long decay time of the a.g.c., especially in the c.w. and s.s.b. modes, keeps the receiver insensitive for a rather long time. So here the wanted audio signal is disturbed too much. This dead time, however, can be brought almost to



zero if the 'on' pulse is also used to discharge the integrating capacitor of the a.g.c. line. During the 10 ms the audio signal is gated off, the a.g.c. is able to stabilize to the new situation due to the much shorter rise-time of the a.g.c. system. (See Fig. 1.)

The 10 ms off-gating can introduce errors on digital communications. Decreasing the audio gating and the a.g.c. settling time to 1 ms will be necessary. However, this was not investigated.

The nine combinations used are shown in Table 1.

1 'on' pulse is generated after 10 ms
2 'alarm' signal
3 'off' pulse after 100 ms in a.m. and f.m. mode and 'alarm',
'off' pulse after 1 s in s.s.b. and c.w. mode and 'alarm'
4 don't care position
5 don't care position

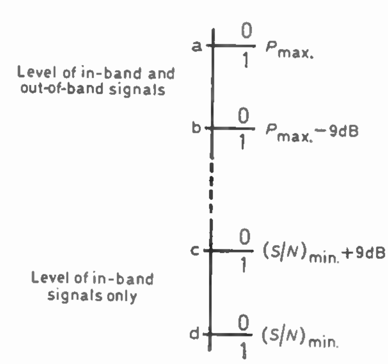


Fig. 1. Block diagram of the complete system.

6 'off' pulse after 100 ms in a.m. and f.m. mode,
'off' pulse after 1 s in s.s.b. and c.w. mode
7 'off' pulse after 2 s
8 'off' pulse after 2 s
9 'off' pulse after 100 ms in a.m. and f.m. mode,
'off' pulse after 1 s in s.s.b. and c.w. mode.

Generating an 'on' pulse in position 2 may cause the i.f. S/N ratio to fall below the minimum acceptable level $(S/N)_{min}$. This is already the case in position 3. Here an 'off' pulse must be given to prevent losing the wanted signal in the noise, in spite of the overloading distortion that may arise. We prefer a distorted signal to a lost signal. But in positions 2 and 3 an alarm signal is given to indicate that the receiver is operating under uncontrolled conditions. This alarm signal may be used to steer a diversity system.

3 Circuit Description

Previous discussions may have given the impression that the complexity of a communication receiver will increase drastically, but that is not so as may be seen from Figs. 1 and 2.

The reed switch attenuator is built in a small screened box, which contains the 6 reed relays and the three attenuators of 6, 12 and 24 dB. The π attenuator configuration is chosen because of screening simplicity. (See Fig. 3.) Standard low inductance resistors are used.

The 0–30 MHz amplifier is realized by a couple of wide-band integrated circuits such as CA 3028 or an equivalent. The gain is set to a comfortable level so that the $P_{max-9\ dB}$ level can be detected without difficulties and so that at the same time the P_{max} level signals will not saturate the amplifier. The amplifier must have good pulse saturation characteristics to prevent blocking effects. Care should be taken to avoid degenerating the noise figure of the receiver by loading the input of the front end with a too low impedance from the sensing amplifier. If the ratio of the impedances of the sensor and the mixer is 20, the increase in the noise figure will be only 0·2 dB.

An almost identical circuit is used for the i.f. sensor amplifier. As the noise level N is rather low, much higher gain is necessary, about 120 dB. The noise figure of the sensor should be less than 10 dB to make a good S/N ratio measurement possible.

The four level detectors are of the 710 type.

In the integrating filter of the a.g.c. circuits, the transistor is shown which will discharge the capacitor to reduce the dead-time of the receiver after an 'on' pulse is generated.

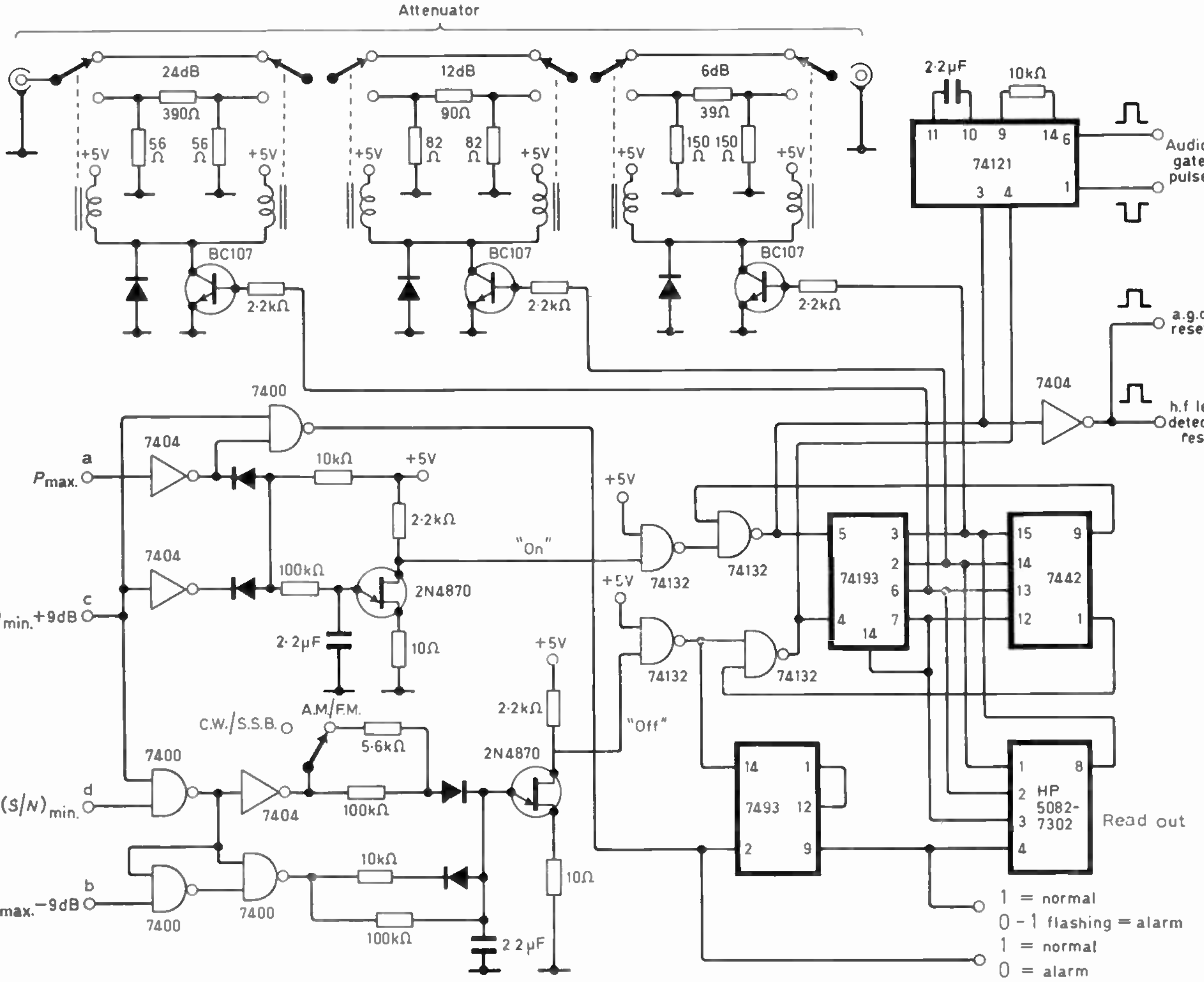


Fig. 2. Logic circuit.

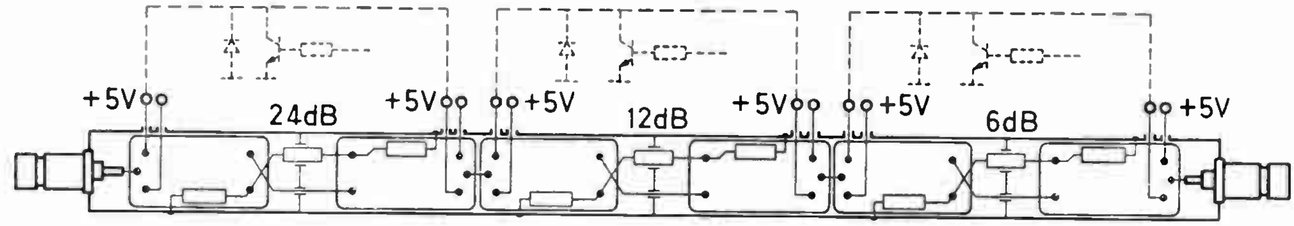


Fig. 3. Reed-controlled attenuators.

The 'on' and 'off' pulses are generated by a couple of uni-junction transistors. Simple logic circuitry is used to control the pulse oscillators. The 'on' and 'off' pulses are fed through Schmitt trigger circuits to avoid spurious pulse generating due to the long rise and decay times of the u.j.t. oscillators.

The up-down counter SN 74 193 is counting between 0 and 7 by control of the b.c.d.-to-decimal converter SN 7442. A l.e.d. read-out display shows the number of attenuation steps in use and a flashing point in case of alarm. The pulse duration of the one-shot generator SN 74121 is set to the minimum possible time interval when the audio signal is gated, and the receiver a.g.c. is returned to stable gain.

4 Conclusion

When high-level, off-frequency signals disturb normal h.f. communications, the automatic input attenuator can give a real improvement in wanted signal quality. This type of interference problem will frequently be present on board ships and aircraft with high communication density.

5 Acknowledgment

We thank Ir. C. M. Ligtvoet for fruitful discussions and help in the practical realization of the prototype.

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Manuscript first received by the Institution on 8th July 1976 and in revised form on 21st February 1977. (Paper No. 1788/Comm 155).

Letters

From: J. H. Roberts, B.Sc., F.I.M.A.
J. D. Ralphs, B.Sc., C.Eng., M.I.E.R.E.

Harmonic Summation

The paper by Ralphs and Sladen[1] on the simulation of fading data while describing a novel approach casts a shadow over the well-known harmonic summation method, describing it as expensive, time consuming, and inconvenient.

Such criticism can largely be agreed when signal generators and filters are assembled in the laboratory, but they are of much less relevance when a computer-based simulation is undertaken using the theory around which multi-oscillator fading machines have been designed in the past.

The purpose of this letter is to provide a few words in defence of an old and well tried analytical tool and to add a comment on the general method used in Ref. 1 for generating normal distributions.

Studies of the probability distributions of sums of tones go back some years[2,3] and if such a sum is written

$$X(t) = \sum_{m=1}^{M} V_m \cos [\omega_m t + \phi_m]$$

where the ϕ_m are a set of pseudo-random phase angles lying in $-\pi$ to $+\pi$, then the distribution of $X(t)$ (for general t) approaches the normal law as M increases, independently of how the ω_m cover the band from ω_1 to ω_M.

The power spectrum of $X(t)$ consists of a set of lines which will be of equal height when V_m is constant (equation (2) of Ref. 1) and $V_m = \exp(-k\omega_m{}^2)$ is the choice that models a Gaussian power spectrum.

$X(t)$ can represent the in-phase component of a carrier tone received over a Rayleigh fading link and a computer simulation will require $Y(t)$, the corresponding quadrature component. $Y(t)$ is obtainable by adding $\pi/2$ to each ϕ_m and this fact, together with a uniform frequency step being permissible, means that $\sqrt{[X^2(t)+Y^2(t)]}$ can be generated with both speed and economy for large M if necessary ($M = 50$ or 100 say[4]) so realizing a distribution that is very close to Rayleigh and free from the graininess caused by quantizing.

In contrast, the general method of Ref. 1 utilizes sums of discrete random variables such as

$$S = \sum_{n=1}^{N} \omega_n x_n \quad \text{where} \quad x_n = \pm 1 \text{ with equal probability.}$$

The ω_n chosen by Messrs. Ralphs and Sladen follow a Gaussian curve and are shown to provide distributions with only small departures from normality and so are fully acceptable in many modelling applications. However, an instructive example for which the distribution of S decreases much faster than a normal law is $\omega_n = 1/n$ (Ref 5).

Simply adding random variables (continuous or discrete) is not a panacea for generating a normal distribution.[6] A bell shape is nearly always achieved but the precise behaviour in the tail region is a matter for investigation case by case and could be important especially when the simulation is to model rare events.

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I am answering Mr. Roberts' letter in the absence of Mr. Sladen, who has, however, been consulted and is in agreement in principle with the reply. The letter raises two points, firstly Mr. Roberts' doubts as regards the accuracy of simulation of the Rayleigh fading method we propose, and secondly his preference for a model based on addition of sine waves to generate a Rayleigh distribution. As Mr. Roberts implies, our statement (in Sect. 11.1) that the Gaussian distribution will be approached for 'arbitrary weighting' is incorrect, in that the Central Limit Theorem breaks down if one weighting is comparable with the sum of all the others (as in the 'inverse weighting' case investigated by Rice which he quotes). While we accept that this is mathematically true, it is irrelevant in the context of the rest of the paper, since we recommend that the weightings should approximate to a Gaussian pulse; therefore they occur in symmetrical pairs and the requirements of the Central Limit Theorem are adequately fulfilled for even a small number of stages. If the number of stages is increased, the differences between the larger weightings decrease (which is not true of the inverse weighting case) and so there would seem to be no case for assuming a limitation on the accuracy obtainable.

Mr. Roberts' remark that 'simply adding random variables...is not a panacea for generating a normal distribution' is somewhat puzzling when it is considered how much of Communication Theory is based on the assumption that a normal distribution is generated by the sum of random events (as in the generation of Gaussian noise from a pseudo-random bit stream) and that a Rayleigh fading signal is itself generated in practice by the sum of a number of signals of random amplitude and phase.

It is not completely clear from his letter whether it is this (or some other) limitation of the Central Limit Theorem which is Mr. Roberts' principal objection to the method, or (in view of his reference to the 'tail region') the 'graininess' caused by amplitude quantization. In the particular realization we described, using 10-stage registers, the minimum amplitude step is about 7% in each of two dimensions in quadrature, giving minimum changes of the order of ¼% (−52 dB on peak) near to maximum amplitudes. I would submit that this is more than adequate accuracy for most purposes and could be improved considerably by even a small increase in the number of stages.

As regards his defence of the multi-oscillator principle, it must be emphasized (as has been evinced in correspondence) that Mr. Roberts is considering the computerized mathematical modelling of a complete signalling chain, including the 'modems' under test. In effect, the aim of his work is to derive a mathematical model for an 'optimum' communication system (by some particular criteria), whereas ours is to design a hardware implementation of an optimum system and to prove its practical performance.

Our criticisms on practical grounds of the multi-oscillator simulators known to us (Refs. 1 and 2 of the paper) are apparently accepted as valid by Mr. Roberts. Our criticism of the incorrect frequency spectrum is answered by postulating a considerable increase in the number of components and a weighting of each component to give a Gaussian power spectrum. Indeed, the model used by Mr. Roberts has virtually nothing in common with the systems described in the references (other than the basic idea of the summing of sine waves to give a Rayleigh envelope).

In addition to the two factors mentioned above, Mr. Roberts generates two 'quadrature' summations of sine waves before modulating his signal with each and summing. While accepting that this model may be mathematically valid, it is not amenable to hardware realization and is strictly a product of computer techniques, and so can hardly be considered an 'old and well-tried analytical tool' but must be judged on its own merits.

On these grounds (and considered in the mathematical concept alone) the system would seem to have little to recommend it. The 'graininess' in the amplitude plane which seems to worry Mr. Roberts (although he presumably carries out A/D conversion on his signal) is replaced by a 'graininess' in the frequency plane, which is less obvious to casual analysis but could still give misleading results. For instance, assume 50 components spaced evenly in frequency between the 1% points of the Gaussian cumulative distribution curve. At a fading rate of 50 fades/second (the maximum recommended by CCIR for h.f. simulation) the frequencies would be spaced 4·6 Hz apart. If this is used to measure the performance of a modern high-stability communication system such as Piccolo, which uses an array of matched filters with 10 Hz effective bandwidths, the results could be most misleading.

I would also disagree most strongly with Mr. Roberts' assumption that 'a uniform frequency step...(is)...permissible'. Taking the example suggested above of 50 components spaced evenly between the 1% points, the spacing is about 0·08 of the fading rate. Since the fading pattern will repeat after a period equal to the lowest frequency beat the total 'machine cycle' is no more than 12 fade periods. This is hardly a long enough test to capture the the 'rare events' that Mr. Roberts requires—in the simplest Piccolo tests it represents no more than 2 or 3 element errors. Taking a more orthodox example, if a binary FEC or ARQ system with a code block length of 100 bits, is to operate at 100 bauds and be tested under 'flutter fade' conditions, say a fading rate of 10 Hz (all these figures are reasonable and by no means extreme), then in order to test over 10^4 blocks, a total test run of at least 10^5 uncorrelated fade cycles is required—which would seem to be completely beyond the range of Mr. Roberts' techniques.

A series of short tests run with different initial values of ϕ_m is a nuisance and in any case runs the risk (unless the values of ϕ_m are specifically computed to avoid it) of repeating (or closely following) the fade pattern of a previous test. Reference to the original Law paper establishes without doubt that the system described there relied on the 'random' frequency drift of crystal oscillators to obtain an adequate test period (and so by implication lacked precision of definition of the fading rate with slow fading!).

It is my opinion that, while Mr. Roberts' system may be adequate for his own requirements, it could not be recommended as a general-purpose technique, either on accuracy of simulation (in the frequency plane) or length of fading pattern.

A problem that can arise with computerized mathematical modelling is that of 'real-time' operation. Although I have little experience of computing techniques, I would doubt whether Mr. Roberts' simulation could be carried out in 'real-time' at any reasonably fast fading rate and therefore its use is strictly limited to mathematical modelling of complete systems, with no facilities for the introduction of 'hardware' equipment at any point. This is a massive restriction on its use, limiting it to the narrower field of quasi-research. If a computerized Rayleigh fade simulation is required, it may be worth investigating a direct modelling of the fading technique described in the paper. I have modelled this in 'slow time' on a desk calculator (HP 9820A) with no difficulty, and the process is basically extremely simple so that real-time computing may be possible.

I am grateful to Mr. Roberts for the interesting correspondence and conversations which have resulted from his letter—which I am sure have caused a considerable amount of revision of preconceived ideas on both sides. It is always instructive to see something of 'how the other half lives'.

This may be an appropriate opportunity to correct two minor errors in the paper. In section 1.2, third paragraph, fourth line, the expression should read $1/2T$, and not $\frac{1}{2}T$. In section 11.1, equation (12) should read:

$$f_r = \frac{0{\cdot}728 f_c}{(n+1)}$$

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14th September 1977

General Interest Paper

Recent Trends in Automotive Radar

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The possible advantages to be gained from the development of vehicle radar systems are outlined and the factors influencing the design of a suitable system are discussed. The various types of system which are commonly considered are introduced and their relative merits and associated problems are explained. Reference is made to typical experimental systems of each type. It is concluded that significant problems remain to be solved before practical automotive radar can become a reality for the general motoring public.

Introduction

Driver error plays a significant role in a very large proportion of road accidents. In particular, following too closely and lack of anticipation of vehicle movements ahead are highly prevalent errors.[1] If vehicles could be fitted with a system to measure range and relative velocity of vehicles or obstacles ahead, and this information used to provide an indication of impending hazards, or to activate emergency braking when required, it is possible that a significant number of these accidents could be prevented. This is supported by a recent study[2] of 215 accidents in the USA which predicted that a hypothetical idealized collision-avoidance radar system would definitely have prevented 18% of these accidents with some possibility of prevention of up to 42%.

It is this potential for accident prevention or mitigation which has led to the development of a number of experimental vehicle radar systems over the last decade.

Meaurement Techniques

Various methods of proximity measurement have been proposed for use in such systems. Those most commonly considered employ sonar, infra-red or microwave radiation. Sonar has the disadvantage that the speed of sound in air is severely dependent on wind speed and temperature, whilst infra-red transmission provides little advantage in penetrating power over visible light when used in the presence of rain and fog.[3]

Although a few experimental systems employing sonar and infra-red have been constructed in the past,[4,5] microwave radar is now generally regarded as the most suitable method of range measurement.

Health Hazards

There is, however, considerable disagreement about the effect of exposure to microwave radiation as a health hazard. The safety standard in the UK and the USA for prolonged exposure is 10 mW/cm² which is 1000 times greater than that permitted in the USSR[6].

A typical experimental automotive radar operating at 35 GHz with a 5° beamwidth and a detection range up to 100 m will employ a c.w. source power of the order of 40 mW. This corresponds to a power density of approximately 0·4 mW/cm² uniformly distributed over the antenna aperture. This power density will fall off rapidly as distance from the antenna increases.

Although 0·4 mW/cm² is well within the UK and USA safety standard, in order to meet the more exacting USSR limit of 0·01 mW/cm² it would be necessary to disable the radar below certain vehicle speeds to ensure that it is not possible to approach sufficiently close to the antenna to be subject to harmful exposure. The effect of radiation from several radars simultaneously must also be considered.

Propagation

Whilst the attenuation of microwave radiation which occurs due to atmospheric absorption and scattering by rain and fog is negligible over the short ranges required for automotive radar,[7] backscatter from rain can be large enough to interfere seriously with c.w. radar operation by masking the return from distant targets. Backscatter increases with radar operating frequency and with one c.w. automotive radar system it has been found necessary to reduce the operating frequency from 36 GHz to around 22 GHz in an attempt to reduce this effect.[8] The use of circular polarization can also aid rejection of rain clutter, but it is possible that c.w. systems will have to be inhibited, for example by a windscreen wiper interlock, during conditions of heavy rain.[9]

Range and Beamwidth

The choice of maximum detection range and beamwidth is essentially a compromise between detection capability, false alarm rejection and antenna size. In order to allow braking to a standstill to avoid a stationary obstacle from an initial speed of 31 m/s (70 mph), under favourable road conditions, a detection range of some 70 m is required.[8] Under wet or icy road conditions this distance is considerably increased. A detection range of at least 100 m is therefore desirable for a collision avoidance radar; longer operating ranges are impracticable because of the problems of false alarms caused by the increasing coverage of the diverging radar beam.

A beamwidth must be chosen which is wide enough to provide adequate detection over the required range on curved and undulating highways, but which does not cause excessive false alarming due to traffic in adjacent lanes, road signs and bridge parapets etc. Experimental collision avoidance radars have been constructed with beamwidths ranging from $2\frac{1}{2}°$ to 10°.

The size of the antenna must be suitable for vehicle mounting. For example, at 22 GHz a 5° beamwidth can be achieved with an 18 cm diameter parabolic dish. At a given operating frequency a narrower beamwidth requires a larger antenna, although antenna size can be reduced by increasing operating frequency. At present, most experimental automotive

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radars operate between 10 and 35 GHz. Higher operating frequencies are not technically or economically attractive at present; this is particularly true if it is anticipated that microwave integrated circuit techniques will be used in an attempt to achieve low cost volume production.

Dynamic Range

Ideally, the performance of an automotive radar could be predicted by considering a fan-shaped radar beam and applying the radar equation which indicates that received echo power varies inversely with the fourth power of range.[10] In practice, the situation is complicated by the presence of side-lobes in the antenna beam pattern and by the occurrence of severe signal nulls at particular ranges. These nulls are caused by destructive interference due to the path difference between signals taking a direct line between transmitter and target, and those travelling via a reflection at the road surface. This interference can be minimized by vertical polarization of the radar antenna, but even so, variation in signal strength of some ±20 dB from the mean value can result.[11]

Additional problems are caused by the wide variety of vehicle radar cross-sections which can be encountered. At 10 GHz rear automobile cross-sections typically lie within the range 10 m² to 100 m², whilst a large truck can register a value of some 2000 m².[12] This represents a possible dynamic range of vehicle radar cross-sections of 23 dB. The radar cross-section of a signpost with a 1 m × 1 m aluminium sign can be some 18 dB greater that that of a typical automobile.[13]

Types of System

All the above factors must be taken into account in the design of an automotive radar system. Four types of system are commonly considered:

Automatic headway control
Headway warning
Automatic braking
Pre-collision sensing

Automatic Headway Control

Headway control systems are designed to maintain a safe following distance from the preceding vehicle. This is achieved by automatic actuation of throttle and brakes based on a knowledge of own vehicle speed (s), and radar measurements of target range (h), and relative velocity (v). A suitable actuating signal may be obtained from a linear control law of the form

$$e = G(h + kv - as)$$

where a and k are headway and relative velocity constants (seconds) and G is a gain constant. A linear law of this type is generally employed for ease of computation and has proved adequate in practice.[14,15]

For steady-state following a range of 31 m is required to allow a headway of 1 second (i.e. 1 m per m/s) for vehicle speeds up to 31 m/s. However, most early workers aimed at a maximum detection range of 100 m in order to achieve reasonable protection against stationary obstacles in the vehicle's path.

Automatic headway control systems have been considered as an addition to the cruise control systems commonly fitted to medium and large-size automobiles in the USA and which are now increasingly found in Europe. One system[15] developed specifically for this application uses a c.w. Doppler diplex technique whereby two closely-spaced frequencies are sequentially transmitted and received. The target range is proportional to the phase difference between the two Doppler signals which are obtained by mixing the transmitted and received signals. Either of the Doppler signals can provide relative velocity information directly.

This homodyne system with a single antenna and a single ended mixer minimizes cost but has the disadvantage that no Doppler signals are present when lead and following vehicle speeds are equal. This problem is overcome by storing the last received radar information until there is a change in conditions. Under conditions of zero Doppler the system is designed to provide such a change by accelerating the car slightly. The system described operates at a frequency of 36 GHz with a beamwidth of 3° to 4°. A range cut-off of approximately 90 m is employed to minimize the masking effects of radar returns from large trucks etc. at long range which can swamp returns from small targets at closer range. This cut-off is achieved by sequentially turning the transmitter on and off.

Another early experimental headway control system, again operating at about 35 GHz, used a frequency modulated continuous wave (f.m.c.w.) radar with sawtooth frequency modulation.[14]

Headway Warning

Some incidence of false alarms due to non-hazardous targets entering the radar beam is inevitable with systems of the type described so far. In order to avoid the possibly severe consequences of such a false alarm in a fully automatic system, a number of workers have therefore developed warning-only systems. These provide an audio or visual warning to the driver if he is following a vehicle too closely or is approaching a slowly-moving or stationary obstacle.

The minimum safe headway under steady-state following conditions is generally set to be about one second which is equal to a typical driver's reaction time. This will allow sufficient time for the driver to react to the radar's warning so that he can avoid a collision if the preceding vehicle undergoes sudden braking.

An f.m.c.w. vehicle radar of this type using triangular frequency modulation has been developed[16] and the way in which range and relative velocity information can be derived using this type of modulation is illustrated in Fig. 1. A four-bit microprocessor is used to perform the required signal processing. A block diagram of the 0·9 cm homodyne radar is shown in Fig. 2. The radar head unit employs a small dual printed plate antenna and is shown fitted to a vehicle in Fig. 3.

A modulation frequency of 50 Hz and a frequency sweep of 200 MHz allow theoretical resolutions of 0.4 m and 0·2 m/s for range and relative velocity respectively. In practice the

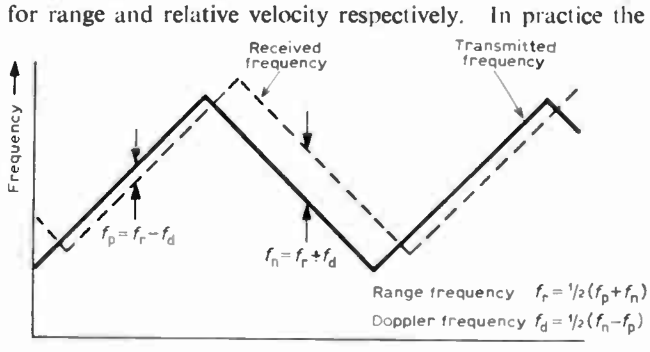

Fig. 1. Triangular frequency modulation waveforms (closing geometry).

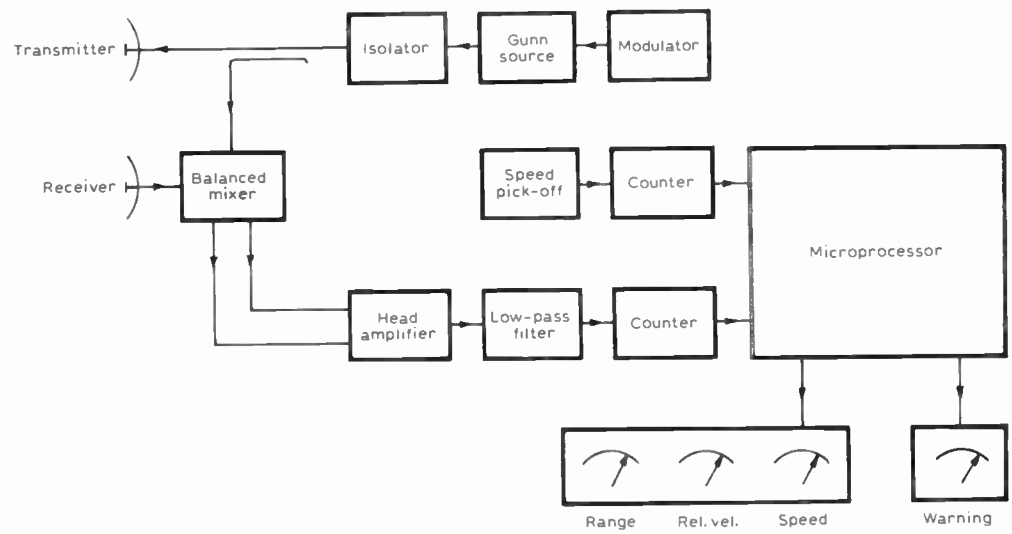

Fig. 2. F.m.c.w. radar headway warning system—block diagram.

complex nature of many targets, which can exhibit a number of reflecting surfaces at different ranges, results in a reduction in accuracy. The maximum detection range of the radar is approximately 100 m for typical target cars.

One major disadvantage of a simple f.m.c.w. radar of this type is that it is possible for the radar return from a large object at long range to swamp that from a smaller but closer, possibly hazardous, vehicle. A typical situation is shown in Fig. 4.

Warning systems using pulse radar techniques have also been developed and are particularly favoured in West Germany.[17,18] These have the advantage that signals due to the nearest target in a multiple target situation can easily be selected; however it must be realized that, particularly on curves, the nearest target is not necessarily the target of interest. The main disadvantages of a pulse system are the high cost incurred by the need for i.f. components, and the relatively wide bandwidth required.

A microprocessor-based f.m.c.w. radar system using sawtooth modulation has also been reported from West Germany.[18]

All the radars described so far have been of the primary (non-cooperative) type, i.e. they will respond to any target

Fig. 3. 0·9 cm f.m.c.w. radar head unit mounted on vehicle (radome removed).

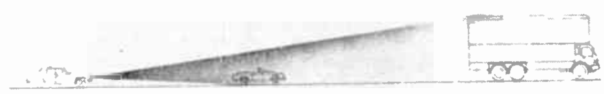

Fig. 4. Primary f.m.c.w. radar—masking of radar return from small target by reflection from large object at long range.

with the result that some incidence of false alarms due to non-hazardous targets entering the radar beam is inevitable. A secondary (cooperative) radar system has been demonstrated[19] which is designed to minimize such false alarms. A passive transponder mounted on the back of vehicles returns the second harmonic of the frequency transmitted from the trailing vehicle. The radar is largely immune to clutter since its receiver is tuned to the second harmonic frequency only. It is also immune to blinding by cars travelling in the opposite direction.

However, harmonic radars have a number of disadvantages.[20] Firstly, all vehicles must be equipped with a special transponder in order to obtain maximum benefit, and secondly, detection range is limited by the poor efficiency of the frequency doubler.

In order to obviate these difficulties, a dual mode (primary and secondary) radar system has been proposed.[21] The secondary mode of the radar is based on tagging cooperating vehicles and other potential highway hazards with modulated fundamental frequency reflectors rather than harmonic reflectors. The range of the secondary system is approximately 100 m and requires much less transmitter power than a harmonic secondary radar. Targets that do not carry tags are also detected by the radar, but at a much shorter range in order to minimize the false alarm problem.

Automatic Braking

A major disadvantage of a warning-only radar system, either primary or secondary, is that the system must always allow for the significant distance which will be travelled during the driver's reaction time after a warning is given before the driver initiates braking. It is possible to design an automatic braking system with a response time significantly shorter than that of a typical driver and the detection range required for the same degree of protection is therefore considerably less. If the radar detection range is reduced below 100 m in an attempt to reduce the problem of false alarms, it can be argued that the delay in brake actuation due to the driver's reaction time in a warning-only system can no longer be tolerated, and that an automatic braking system is imperative if significant accident reduction or mitigation is to be obtained.

Whereas work in Europe has been confined mainly to warning systems, recent developments in the USA have centred on these automatic braking systems. One such system is a development of the Doppler diplex primary radar headway control system described earlier. Extensive road testing with this system showed[9] that false alarms could be effectively suppressed by employing a narrow beamwidth (2·5°) and by limiting the maximum radar detection range to 45 m.

A statistical analysis of accident data indicated that, even with the restrictions on performance imposed by the narrow beamwidth and limited detection range, over 20% of US road accidents could be prevented by a fully automatic radar braking system of this type.

A further automatic braking system employs a primary f.m.c.w. radar with triangular frequency modulation[22] and has been fitted to an Experimental Safety Vehicle (ESV). The system operates at X-band with a frequency deviation of ±25 MHz and a sweep rate of 1 kHz. A 5° × 10° printed circuit antenna with extremely low sidelobe levels is employed and the maximum detection range is 25 to 30 m. A microprocessor is used to perform the demodulation, and target discrimination is achieved by rejecting signals which do not give consistent data over a number of consecutive modulation cycles.

The system is designed to operate in headway control and headway warning modes as well as providing automatic braking. The last is initiated only if a severe set of constraints is satisfied, namely vehicle speed over 10 m/s; range less than 25 m; relative velocity greater than 16 m/s; steering wheel angle zero; driver not already braking. The philosophy behind imposing these constraints is that false alarms must be prevented at all costs: the aim is to achieve accident mitigation when a collision is inevitable rather than complete accident prevention. It is felt that this is a realistic design goal with the present state of technology.

Pre-collision Sensing

A further application of automotive radar is as a predictive crash sensor for use in the automatic deployment of passive restraints. Passive restraints are devices such as air bags which do not require passenger effort or cooperation to provide protection.

At present such devices are usually triggered by a mechanical deceleration sensor set to detect the decelerations of some 5*g* which occur under conditions of impact. The total response time of such sensors is often in the region of 20–40 ms which seriously compromises the effectiveness of the restraints. Pre-sensing impending collisions by radar would allow earlier deployment of the restraint resulting in increased occupant protection.

In order to avoid false alarms, the detection zone must be limited to approximately 2 m but must extend over a large proportion of the width of the vehicle. This is usually achieved by means of a bistatic configuration, i.e. separate transmit and receive antennae mounted on either side of the front of the vehicle and angled inwards towards a point ahead. For simplicity a c.w. system is commonly employed.[23]

The range can be limited more accurately by means of a pulsed Doppler system.[24] Such a system has been used in an ESV to tighten passengers' seat belts automatically in the event of an impending collision. An ultra-short pulse radar has also been proposed for pre-collision sensing.[25]

Clearly false alarms must be avoided at all costs in any air-bag activation system. This is particularly difficult to achieve on account of the problem of assessing the degree of hazard of a particular target. For example, at some angles of approach the radar returns from relatively harmless targets such as shopping trolleys or road sign poles can exceed those from many vehicles. Target discrimination techniques aimed at assessing the degree of hazard of a target have been studied in considerable depth.[26]

An example of such a technique is radar signature analysis whereby the variation in amplitude of radar return as a target is approached is compared with a stored library of signatures from targets of known hazard. However, most of these techniques are severely dependent on angle of approach. Although some techniques appear promising the problem of automotive target discrimination remains to be solved in practice.

Other Work

An alternative approach to vehicle radar has been taken in the design of a system to allow emergency service vehicles to be driven under conditions of poor visibility.[27] A scanning f.m.c.w. technique is employed in conjunction with a frequency-sensitive antenna. The radar scene ahead of the vehicle is displayed to the driver on a television monitor. This scheme has the considerable advantage over other systems that interpretation of the radar signals is left to the very powerful human brain. Although this system has been demonstrated experimentally it is not aimed at general highway use.

In another even more complex development[28] an automatic headway control system based on the use of a microwave telemetry link between vehicles has been constructed. Each vehicle transmits data concerning its longitudinal position measured with respect to a datum in the road, its acceleration, lane number and other information. Two vehicles have been suitably equipped and automatic following successfully demonstrated, but work has now been curtailed due to economic considerations.

Microwaves have also been employed in vehicle identification and location systems. Such systems employ identity transponders which can be either active[29,30] or passive[31] and which, when interrogated with microwave radiation, reply with a coded signal indicating their identity. These identification systems can be used in such areas as traffic control, anti-theft protection, and location of public service vehicles etc. A transponder has been described[21] which fits in the space of a vehicle licence plate and which can act both as an identity transponder and as a transponder for use in secondary radar collision avoidance systems.

Work has also been directed towards the development of microwave vehicle speedometers.[7] Although c.w. Doppler radars are extensively used for aircraft speed measurements, the very short range from vehicle to road surface poses a different set of problems for a vehicle speedometer. In particular, severe amplitude modulation nulls can result due to vehicle vibration or rough road surfaces. Furthermore the radar return from very flat surfaces such as wet, smooth concrete can be very small.

Another use for c.w. Doppler radar has been found as a backing radar to provide a warning to drivers of large trucks if they are reversing towards an obstacle in the blind spot behind their vehicle. Such a unit is now being marketed in Sweden.

Conclusions

It can be seen that there has been considerable activity in the field of research into automotive radar. However, with the exception of the backing radar, all the systems described have been of an experimental nature only. Whilst the progress reported by some workers is encouraging it is clear that significant problems remain to be solved before practical automotive radar systems become a reality for the general motoring public.

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Manuscript received by the Institution on 8th August 1977. (Paper No. 1789/AMMS 82)

IERE News and Commentary

New Chairman for MacRobert Award Committee

The MacRobert Award Evaluation Committee, which annually assesses entries for Great Britain's major engineering award, is to have a new Chairman, Mr John C. Duckworth, Chairman of IDJ Investment Services Limited. Mr Duckworth, a member of the Evaluation Committee for some years, replaces Lord Hinton of Bankside, now President of the Council of Engineering Institutions, who has been its chairman since 1969 (when the award was established).

John Clifford Duckworth, M.A., F.Inst.P., C.Eng., F.I.E.E., was educated at Wadham College, Oxford. He has a background of war-time radar research at TRE Malvern, nuclear physics research in Canada and at Harwell guided, missile development with Ferranti and from 1954 to 1959 he was responsible within the then British Electricity Authority (now Central Electricity Generating Board) for design, construction and operation of the first civil nuclear power stations. From 1959 to 1970 he was Managing Director of the National Research Development Corporation (NRDC) and from 1970 to 1973 Adviser to N. M. Rothschild & Sons Limited.

The MacRobert Award, consisting of £25,000 and a gold medal is made annually by the Council of Engineering Institutions on behalf of the MacRobert Trusts. It is presented in recognition of an outstanding contribution by way of innovation in engineering or the physical technologies or in the application of the physical sciences which has enhanced or will enhance the prestige and prosperity of the United Kingdom. Past awards made include those for the Pegasus engine used in the Harrier VTOL aircraft, the EMI-Scanner, the Dunlop Denovo tyre, railway vehicle suspension for the Advanced Passenger Train, and the rotor system and gearing for the Westland Lynx helicopter.

Change of Name for Mullard Research Laboratories

On 1st June 1977, the Mullard Research Laboratories, Redhill, were renamed the Philips Research Laboratories. This is to mark the Laboratories' status as the corporate research establishment of the Philips Industries group of companies in the UK. Mr Norman Goddard, the Laboratories' Director, will continue to be responsible to Dr P. E. Trier, the Philips Industries Director with special responsibility for Research and Development.

As in the past, a substantial part of the research programme will be concerned with electronic devices and materials for the consumer, industrial and government markets served by Mullard Limited and research support will be provided for the Pye Group of companies, MEL and other operating divisions of Philips Industries.

Software Reliability Problems

The development of reliable computer software is still well behind the development of reliable hardware. But the gap is narrowing, albeit slowly. That is the main conclusion in the latest State of the Art Report on Software Reliability by Infotech International.

The Infotech Report examines software reliability concepts and its economies; requirements analysis and specification; methods of reliable software design; software project management; computer-aided software development; software validation; program testing tools and techniques; and software fault-tolerance.

Hardware faults arise from four main causes according to the Report: design error, production error, life expiry and environmental disturbance. While manufacturers have made great strides in mitigating their effects, developments in software reliability have shown only a marginal improvement.

The balance, however, is now starting to be redressed, largely as a result of developments in the following areas:

Requirements analysis—Because mistakes in original requirements specification are so expensive to correct at a later date, much significant work has been done in recent years on techniques for expressing requirements in a self-consistent notation that can be verified.

Design methodologies—Three criteria of well-designed software are brevity, simplicity and reliability. Design methodologies, formalized approaches to describing the arrangement of components part that communicate through well-defined interfaces, facilitate the process.

Computer-aided software development—Several research and development projects are currently working towards producing executable software for particular computing environments directly from a set of functional requirements.

Other areas where developments are taking place include software validation; reliability modelling; fault-tolerance techniques; and architectural support for reliable software.

Reliability can nearly always be achieved, the Report stresses, but at a price. This price may take the form either of actual increased cost or of degraded performance. Notwithstanding this, the most important factor for the user must ultimately be that the software should perform the required tasks, no more and no less.

A prime example of the former attitude still lurks in the hearts of many programmers whose desire to write efficient programs all too often overwhelms the requirement to write correct programs. Correctness remains, above all other factors, the foundation on which reliable software can be built. To achieve this, it is pointed out, major changes will be required in the practices and attitudes not only of programmers, but also of systems analysts, administrators and managers.

Invited papers in the report come from experts all over the world. The price of the Report is £95 or £72 if bought as part of the complete 'State of the Art' series, from: Infotech International Limited, Nicholson House, Maidenhead, Berkshire SL6 1LD, England.

'Trade Unions and the Professional Engineer'

Although traditionally among the more placid groups in the community, professional engineers have recently shown signs of discontent. They are worried by the contrast between the increasing power of trade unions and the apparent impotence of the professions.

But for many there is a problem. How does trade unionism fit the professions, especially engineering? Many fear that the cautious voices of a few thousand engineers would be overwhelmed among the clamouring thousands of non-engineers in a big union, and the particular professional attitudes they hold would go unheeded. Many fear that on joining a union they could be forced to strike or take up other kinds of industrial action, regardless of their professional responsibilities to their clients or the public. A dilemma exists; but how real is it?

A 76-page booklet* written by Will Howie, an engineer who is currently editor of *New Civil Engineer* (the ICE members' publication) sets out clearly the objectives of trade unions, and the methods they can and do use to achieve those objects. A large section of the book deals with recent legislation on employment, particularly the problems of the engineer and the potential conflict between his professional responsibilities and membership of a trade union. The policy of the CEI on trade union membership is covered, and the organization and operation of the TUC, especially in relation to its political aspects, are dealt with in detail. The booklet ends with brief descriptions of some of the unions which are of particular interest to engineers.

New Presidents of the Institutions

On 1st July, Mr Derek Kimber, O.B.E., M.Sc., F.C.G.I., D.I.C., succeeded Admiral Sir Horace Law, G.C.B., O.B.E., D.S.C., as President of the Royal Institution of Naval Architects. Mr Kimber is Chairman of Austin and Pickersgill Limited, the Sunderland shipbuilders.

The Council of the Institution of Electrical Engineers has chosen Mr J. M. Ferguson, C.B.E., B.Sc.(Eng.), to take office as President on 1st October. Mr Ferguson who will succeed Mr E. S. Booth, C.B.E., F.R.S., is an independent consultant; from 1969 to 1973 he was Director of Engineering with GEC Power Engineering Ltd.

Mr Handel Davies, C.B., M.Sc., who is Corporate Director, responsible for Corporation Technical Policy and for co-ordinating all research design-development and test work in British Aircraft Corporation, now part of British Aerospace, took office as President of the Royal Aeronautical Society on 12th May. He succeeded Mr Charles Abell, O.B.E.

Professor Sir Hugh Ford, D.Sc.(Eng.), Ph.D., F.R.S., is the new President of the Institution of Mechanical Engineers, having taken over from Dr Ewen M'Ewen, C.B.E., D.Sc., F.R.S.E., the retiring President, at the Annual Meeting on 27th April. Sir Hugh is Head of Department and Professor of Mechanical Engineering at Imperial College, University of London.

Professor Robert Nelson Pryor, A.R.S.M., B.Sc., Head of the Department of Mining and Mineral Technology at the Royal School of Mines, London, has assumed the presidency of the Institution of Mining and Metallurgy.

Mr Peter Dunican, C.B.E., elected President of the Institution of Structural Engineers for 1977-78, is the Chairman of Ove Arup Partnership and a Director of Ove Arup and Partners Consulting Engineers, London. He has served since its inception in 1964 as a part-time Director of the National Building Agency.

Electronics Technicians Salaries in 1977

The results of a survey of technicians' salaries recently published by the Society of Electronic & Radio Technicians show that technicians in industry have fallen behind their opposite numbers in other areas, notably government service and education, as far as salary increases are concerned.

The report makes clear that technicians' salaries, overall, mostly moved by the statutory amount (i.e. about £300) during 1976. However, those whose salaries are geared to an incremental scale, as in the Civil Service and education, have had nett increases well above the £300 p.a. Members from nationalized industries, for example, reported receiving over three times the percentage increase compared with those in the field of industrial electronics (17·3% as against 5·3%). Also the average gross salary for technicians in the former categories is over £350 p.a. in excess of their industrial colleagues.

*Published by Thomas Telford, P.O. Box 101 London EC1P 1JH £1.15

Computer Aided Production Management Scheme for Small Firms

The Department of Industry has launched a new scheme to help small manufacturing firms to determine whether computer aids for management of production are likely to be viable. This will finance consultants to carry out feasibility studies in the firms.

The new scheme aims to promote the use of computer aided production management (CAPM) where it is likely to lead to an increase in efficiency. There are two main objectives—to encourage serious consideration of the use of CAPM by small to medium-sized manufacturing companies with up to about 500 employees and—to accelerate growth in the availability of competent consultancy able to help such small firms both to study and implement CAPM systems.

Many small companies may not have experience of using computer systems, perhaps other than for accountancy purposes, and thus be unaware that CAPM systems could be applied to their production management problems. The basic aim of the scheme is to relieve the company of some of the risk entailed in carrying out a feasibility study to explore if a CAPM system could be of benefit, by funding a consultant to carry out this task.

The scheme which is financed by the Department of Industry will last for three years and is administered and monitored by the Blacknest Production Control Group (PCG). This is a small group of engineers with considerable expertise and practical experience of CAPM, whose parent organization is the Atomic Weapons Research Establishment, Aldermaston.

Under the scheme an independent consultant with experience in the production management field identifies a prospective client company which wishes to consider the practical values of installing a CAPM system. The consultant approaches PCG who will assess the proposal in association with the consultant and the company. If the proposal is accepted, PCG will let a contract with the consultant for a feasibility study to be carried out in the company at a maximum cost of £2,000. If during the 24 months following the completion of the study, the consultant is commissioned by the company to instal a CAPM system, he will refund the money received for the study to PCG.

Those eligible to apply are independent consultants who are not connected with computer manufacturers or the prospective client company. Prospective user companies can also seek advise on the scheme directly from PCG. Enquiries should be addressed to: The Secretary, Production, Control Group, Blacknest Centre, Brimpton, nr. Reading Berks RG7 4RS. (Tel: 07356 4111, Ext 5951).

PCG has implemented several CAPM systems in a number of small companies, including a system developed by the Group For the past five years the Group has been working for the Department on a project for organizing demonstrations at companies to show visitors CAPM systems in operation. The experience gained from the project showed the need for the new scheme.

Airborne Microwave Relay System for Nigeria

Following its recent £6·7 million Bolivian contract (see June Journal, page 296) GEC Telecommunications Limited of England has announced further major export orders worth nearly £13 million for transmission systems in connexion with the current National Development Programme in Nigeria. Since completing the £10 million Nigerian nationwide telecommunications transmission network in 1975 GEC Telecommunications Limited has received more orders from Nigeria totalling £28 million. This is a significant contribution to the company's export drive in which the Transmission Division has been particularly successful with some 50% of its orders being for export.

For one of the latest Nigerian contracts, microwave-radio and associated equipment will be supplied for a new high-density television and telephony 'satellite' communications system being provided by the TCOM Corporation of the USA, a subsidiary of the Westinghouse Corporation. Five pairs of aerostat balloons, carrying sophisticated electronic equipment and tethered at a height of some 4000 m (12 000 ft), will provide communications between a number of ground stations covering practically the whole of Nigeria. The new system will create a nationwide colour television service relaying programmes from six regional studios to television broadcast transmitters mounted in the balloons. National telephony circuits will also be provided. The GEC microwave-radio equipment will be mounted in a gimbal assembly underneath the balloons which will be aerodynamically stable and designed to withstand hurricane-force winds. GEC microwave-radio equipment will also be installed at the ground stations.

In addition to the equipment for the aerostat system, a 306 km (190 mile) microwave-radio link will be installed between Ilorin and Lokoja to extend the existing GEC microwave-radio network which connects all Nigeria's large centres of population.

The microwave-radio equipment for all these contracts will be GEC 6·8 GHz 150-M equipment which embodies sophisticated microwave integrated circuit technology to enable the apparatus to be mounted in compact individual modules which plug into racks only 150 mm wide. The contracts for the Ilorin-Lokoja link and the aerostat system ground stations include the provision of antenna towers as well as microwave-radio equipment.

Final adjustments being made to a 6 GHz antenna at Ibadan, one of the 100 radio stations supplied and built throughout Nigeria for the 2500 mile nationwide microwave-radio trunk transmission network installed by GEC Telecommunications Limited.

The equipment will be made at the company's Coventry factories and installation will commence early in 1978 for a period of 2½ years.

Automated Printed-Circuit Design

A system for translating a designer's freehand layout of a new printed-circuit design into the finished artwork needed to commence manufacture of printed-circuit boards, has been developed by Ferranti Cetec Graphics Limited. The complete system is known as PCS (Printed Circuit System) and incorporates a Ferranti Cetec System 4 digitizer operating in conjunction with a mini-computer and a 430 artwork generator.

A designer first draws, freehand, the pattern of the printed-circuit required on graph paper at twice or four times full scale. This circuit layout is then placed on the table of a System 4 digitizer. The various points on the circuit diagram are digitized and relayed to a mini-computer together with other associated data relevant to particular detail such as connection track width, pad size and drilling requirement. Under the control of the PCS program the computer processes the data and produces a check plot. This is returned to the designer to confirm that the circuit layout is correct. Any errors, corrections or circuit amendments are incorporated into the original data in a second computer run. When the design is confirmed as being accurate the computer generates the necessary control information needed to operate the EP 430 artwork generator. Supplied with this information, the artwork generator produces the final artwork of the printed-circuit on photographic film, at the correct scale required for manufacture of the printed-circuit boards.

The EP 430 was designed by Ferranti Cetec specifically for printed-circuit work. It has a working area of 61 cm × 91·5 cm (24 in × 36 in), and a maximum drawing speed of 10 cm/s. It incorporates a light-spot projector with automatic selection of any of 64 apertures for land or symbol flashing and track drawing on photographic film. The choice of symbols can be changed by fitting replacement aperture disks, or by replacing individual apertures. In addition to printed-circuit masks, PCS can be used to prepare other finished manufacturing artwork as long as it can be sketched on graph paper and is within the repertoire of the symbols provided on the aperture disks of the EP 430.

Conference on 'Programmable Instruments'

Organized by THE INSTITUTION OF ELECTRONIC AND RADIO ENGINEERS with the association of the Institution of Electrical Engineers, The Institute of Quality Assurance, The British Computer Society and the Institute of Electrical and Electronics Engineers.

National Physical Laboratory, Teddington, Middlesex

Tuesday 22nd to Thursday 24th November 1977

PROVISIONAL PROGRAMME

Tuesday 22nd November

Session I. PROGRAMMABLE INSTRUMENTS CONCEPTS

'The totally programmable instrument'
By Dr. J. E. BRIGNELL and Dr. C. J. BUFFAN (*City University*)

'The case for a simple, fast programmable interface converter'
By R. YOUNG and Dr. J. E. BRIGNELL (*City University*)

Session II. MICROPROCESSORS

A. SIGNAL SOURCES

'A digitally controlled sinusoidal signal generator'
By Dr. J. F. W. BELL, J. Y. F. CHEN (*University of Aston*) and J. O. M. WILLIS (*SINTRA, Paris*)

'An investigation of linear interpolation applied to analogue function synthesis'
By W. A. EVANS and N. NAVAB (*University College, Swansea*)

'A programmable sinewave signal source'
By W. A. EVANS (*University College, Swansea*)

'A novel technique for programmed generation of multifrequency composite sinusoids'
By S. M. JAWAD (*UWIST, Cardiff*) and Dr. P. A. PAYNE (*University Hospital of Wales*)

'A digitally controlled frequency synthesized signal source'
By W. A. EVANS and R. W. JENKINS (*University College, Swansea*)

Wednesday 23rd November

Session II. MICROPROCESSORS (continued)

B. ANALYSERS

'A microprocessor based instrument for rapid frequency response measurement'
By Dr. D. REES and D. DOYLE (*Polytechnic of Wales*)

'Retino-tectal mapping using a microprocessor controller'
By J. M. SOWDEN, D. B. EVERETT, J. H. SATCHELL and J. E. LEWIN (*MRC*)

C. DESIGN AND ERGONOMICS

'Front panel design'
By D. E. O'N. WADDINGTON (*Marconi Instruments*)

'Instrument/operator interfacing—a case study'
By Dr. R. MURRAY-SHELLEY (*Polytechnic of Wales*) and Dr. J. SHAW (*University College, Cardiff*)

Session III. INTERFACING

'The remote control of instrumentation'
By H. A. DOREY (*Solartron*)

'Design of an IEC interface for a digital voltmeter'
By Dr. H. KÖHLER (*University of Erlangen*)

'Programmable interfacing for microwave gain and noise figure measurement'
By N. M. HOSSEINI, D. A. ABBOTT and Dr. H. V. SHURMER (*University of Warwick*)

Session IV. TEST, MAINTENANCE AND CALIBRATION

'Evaluation of electronic test equipment for service application'
By Major B. REAVILL (*MOD*)

'The use of programmable test aids in calibration areas'
By K. R. BLACKMORE (*Marconi*)

'Debugging microprocessor systems'
By Dr. D. B. EVERETT (*MRC*)

Discussion—'Programmable Instruments—Master or Servant?'

Thursday 24th November

Session V. APPLICATIONS

'An economical intelligent plotter'
By J. D. MARTIN and C. T. DALZELL (*University of Bath*)

'A microprocessor-based digital automatic measuring equipment—DAME'
By J. B. WATSON, P. J. DODDS and A. RAWLINGS (*IBA*)

'The use of a microprocessor in routine cardiac assessment'
By Dr. F. K. HANNA, Dr. M. DIPROSE, M. FITZGERALD (*University of Kent*) and D. J. E. TAYLOR (*Kent and Canterbury Hospital*)

'The application of a microprocessor to provide accurate and reliable temperature measurement'
By A. C. CARVER and P. J. SPREADBURY (*University of Cambridge*)

Further information and registration forms from: Conference Department, IERE, 99 Gower Street, London WC1E 6AZ. Telephone: 01-388 3071, ext. 17 or 18.

Forthcoming Conferences and Colloquia

Fifth International Conference on Gas Discharges

The conference is organized by the Science, Education and Management Division of the Institution of Electrical Engineers (IEE) in association with the IERE, the Institute of Electrical and Electronics Engineers and the Institute of Physics. It will be held at the University of Liverpool from 11th–15th September 1978.

Scope

The Conference will cover the engineering applications of gas discharges and relevant fundamental processes such as:

Applications

Arc furnaces
Compress-gas insulation
Display devices
Fuses
Gas-blast switchgear
Gas-discharge lasers
Gas-filled valves
High-voltage technology
Lamps
Vacuum switches
Welding, cutting and machining

Fundamentals

Corona
Electrode-less and radio-frequency discharges
Electrode phenomena
Glow discharges
High and low pressure sparks
Laser produced plasmas
Lightning
Long sparks
Plasma diagnostics and measurement techniques
Pre-breakdown phenomena
Radiation from discharges
Spectroscopy
Transient and steady-state phenomena in high and low-pressure arcs

(Discharge chemistry, MHD generation and fusion will *not* be included.)

The Organizing Committee invite intending contributors to submit for consideration 25 copies of a synopsis (about 250 words) to the Conference Department on or before 8th November 1977. The subject area should be indicated to assist the grouping of sessions.

Selected contributions, of not more than 3,200 words (including illustrations) will be required for assessment by 10th April 1978. The working language of the Conference is English and will be used for all printed material. Simultaneous interpretation will not be provided.

Full details on the conference are available from the IEE Conference Department, Savoy Place, London WC2R 0BL.

Non-linear Identification

Experimental methods of process identification have become well established over the past few years, but they are by and large appropriate for systems which are linear. Recently, work has been started on identifying non-linear systems, in which the non-linearity is characterized either mathematically (in terms of Volterra integrals, for example) or specifically, with the method being tailored to the particular non-linearity.

Non-linear system identification is to be discussed at a Colloquium organized jointly by the IEE by Professional Group C2 (Control Methods and Computing) and the IERE Specialized Group on Automation and Control Systems to be held at Savoy Place, London on Tuesday, 16th May 1978. Workers in this field who wish to contribute to the Colloquium are invited to submit an Abstract to the chairman of the organizing committee, Dr. K. R. Godfrey, Department of Engineering, University of Warwick, Coventry CV4 7AL, not later than 13th January 1978.

Microprocessing Applications

Following two successful conferences on microprocessors, the Yorkshire Section is organizing a third to be called 'Microprocessing '78—The Ins and Outs of Micros'; it is to be held on Tuesday and Wednesday, 11th and 12th April 1978, at the Old Swan Hotel, Harrogate.

The Section Committee welcomes offers of papers for the Conference from engineers and organizations who have applied microprocessors.

Further information may be obtained from Mr. P. F. Davies, 26 Durham Way, Harrogate, North Yorkshire HG3 2TB (Telephone Harrogate 702842 (office), Harrogate 62198 (home).

A.F. Power Amplifier Design

Professional circuit designers, control engineers, applications engineers and hi fi enthusiasts amongst others will be interested in the Thames Valley Section's Half-day Colloquium on 'The Design of Hi-Fi Audio Power Amplifiers' which will be held on the afternoon of Wednesday, 25th January 1978 at the J. J. Thomson Physical Laboratory, University of Reading.

Experts will recognize that the requirements for power amplifiers for control systems and audio systems have converged in recent years (or have they?) and device technology has improved greatly. Specifications are calling for direct coupling, high bandwidth, low drift, good linearity, stable operation in feedback configurations, high power outputs (particularly under transient conditions), and current overload and back e.m.f. protection.

Three main contributors will each give an account of his own design methodology, and there will be a discussion session opened by Professor E. A. Faulkner (Reading University) who will review the presentations from the standpoint of theoretical circuit design.

The speakers will be Peter J. Baxendall (now an independent electro-acoustical consultant) J. L. Linsley Hood (British Cellophane Ltd.) and J. Vereker (Managing Director of Naim Audio).

The accommodation in the lecture theatre is limited, and entrance is free to members but *by ticket only*. Members of the Thames Valley Section will be circulated nearer the event but members are advised to apply for their ticket in good time to avoid disappointment. *Each applicant will be limited to one ticket.* Apply in writing to Mrs. E. R. Atkinson, Department of Cybernetics, University of Reading, 3 Earley Gate, Whiteknights, Reading RG6 2AL, Berks.

Third International Conference on Software Engineering for Telecommunication Switching Systems

Organized by the Electronics Division of the Institution of Electrical Engineers (IEE) in association with the Finnish Association of Electrical Engineers (SIL) and the Institution of Electronic and Radio Engineers, the conference will take place at the University of Technology, Helsinki, Finland from 27th–29th June 1978.

The Conference will provide a forum for exchange of ideas on software for telecommunications systems in respect of future trends and practical experience. Emphasis will be on achievements of technical interest and advanced ideas rather than on reports of a routine nature. The programme will be limited to a maximum of 30 papers to ensure that there is time for discussion.

The Conference will cover the following areas:

The impact of new technology
- Software implications of distributed control
- Software implications of data protection (privacy)
- The impact of firmware and microprogramming
- The need for software
- Interpretation versus compilation

The role of the operation user (e.g. operating company, data network administration)
- Provision of generic software
- Portability
- In-service modification up-grading
- Programming skills required

Tools, methods and standards

Applications and systems design
- Practical experience

A tutorial session on CCITT high level language will be arranged and papers are invited on the implementation difficulties.

The working language of the Conference to be used for all printed material is English. Simultaneous interpretation will not be provided.

Full details of the conference are available from the IEE Conference Department, Savoy Place, London WC2R 0BL.

STANDARD FREQUENCY TRANSMISSIONS

(Communication from the National Physical Laboratory).

July 1977	Relative Phase Readings in Microseconds NPL—Station (Readings at 1500 UT)		
	MSF 60 kHz	GBR 16 kHz	Droitwich 200 kHz
1	4·1	5·2	37·9
2	3·9	5·0	36·5
3	3·7	5·1	35·7
4	3·9	5·1	35·1
5	4·2	3·8	34·5
6	4·0	4·3	34·3
7	4·0	4·9	33·5
8	4·2	5·1	32·6
9	4·2	5·1	32·4
10	4·2	5·1	32·2
11	4·4	5·3	32·0
12	4·4	5·2	31·5
13	4·4	5·5	31·0
14	4·4	5·4	30·3
15	4·3	5·6	29·4
16	4·1	5·2	28·6
17	4·3	5·3	28·0
18	4·0	5·3	27·7
19	4·3	5·5	27·5
20	4·3	5·3	27·2
21	4·3	5·2	26·7
22	4·2	5·3	26·3
23	...	4·9	26·1
24	...	4·9	25·9
25	3·4	5·3	25·7
26	3·4	4·9	25·3
27	3·4	5·0	24·3
28	3·3	5·2	24·7
29	3·3	4·3	25·5
30	3·4	5·0	26·3
31	3·2	4·9	26·2

August 1977	Relative Phase Readings in Microseconds NPL—Station (Readings at 1500 UT)		
	MSF 60 kHz	GBR 16 kHz	Droitwich 200 kHz
1	3·3	5·0	25·5
2	3·3	5·0	26·2
3	3·5	4·7	26·7
4	3·8	4·8	27·7
5	3·3	5·1	28·0
6	3·2	4·7	28·3
7	3·2	4·8	28·6
8	3·2	5·3	28·8
9	3·2	5·0	28·7
10	3·2	5·0	28·9
11	3·1	5·0	29·4
12	3·1	4·9	30·2
13	3·1	4·7	31·3
14	3·1	4·6	32·5
15	3·1	4·6	33·4
16	3·0	4·6	34·6
17	2·9	4·6	35·7
18	3·0	4·7	36·8
19	3·1	4·6	35·5
20	3·1	4·6	34·5
21	3·2	4·6	33·2
22	3·2	4·6	32·0
23	3·1	4·5	30·5
24	3·2	4·7	29·4
25	3·3	4·7	28·5
26	3·3	4·7	27·7
27	3·3	4·6	26·7
28	3·5	4·5	25·7
29	3·3	4·7	24·4
30	3·3	4·7	23·6
31	3·1	4·7	22·9

Notes: (a) Relative to UTC scale (UTC_{NPL} Station) $= +10$ at 1500 UT, 1st January 1977.

(b) The convention followed is that a decrease in phase reading represents an increase in frequency.

(c) Phase differences may be converted to frequency differences by using the fact that 1 μs represents a frequency change of 1 part in 10^{11} per day.

Members' Appointments

CORPORATE MEMBERS

Mr. G. Angelou (Member 1967, Graduate 1963) has joined Crow of Reading Ltd as Overseas Marketing Manager. He was previously Sales Director for Cetec Systems Ltd. who he joined following engineering appointments with Granger Associates Ltd. and the Plessey Company.

Commander H. E. Cook, RN (Member 1965, Graduate 1959) has completed two years on the staff of the Royal Naval Engineering College, Manadon, and following promotion has been appointed Radio School Commander in HMS *Collingwood.*

Mr. G. D. Crawley (Member 1973, Graduate 1964) has moved to the Directorate of Air Radio at the Ministry of Defence Procurement Executive. He is a Professional and Technical Officer II and for the past eight years he was with the Technical Support Division of the Central Computer Agency of the Civil Service Department.

Mr. E. M. Glasscott (Member 1973, Associate 1961) formerly General Manager of the Jamaica Telephone Company has joined Continental Telephone Holdings Corporation as Deputy Director for a US/AID-financed study for improved telecommunications in the Arab Republic of Egypt.

Mr. M. F. Jollyman (Member 1971, Graduate 1967) who was with the Bell Punch Company, is now a project leader with Chubb Alarms Manufacturing Ltd.

Mr. M. A. Lane (Member 1973, Graduate 1971) is now Group Export Manager with Bowthorpe-Hellermann. He previously held the position of Export Marketing Manager with Jermyn Industries.

Mr. C. R. Wheeler (Member 1972, Graduate 1969) who went to the United States in 1970 after holding appointments with Westland Aircraft and Plessey Radar, has joined Smith Kline Instruments of Sunnyvale, California, as Senior Electronics Engineer Project Manager working on ultrasonic echocardiography instruments and systems.

Mr. Yuen Heng Seng, B.Sc. (Member 1976) has joined Shell Eastern Petroleum (Pte), Singapore, as an Instrument Engineer. For the past six years he served with the Singapore Armed Forces, latterly as a Staff Officer in the Applied Research Section of Headquarters Communications and Electronics.

NON-CORPORATE MEMBERS

Sqn. Ldr. C. Davey (Graduate 1968) has been posted to RAF Benson as Senior Project Officer in the Radio Introduction Unit. He was formerly Engineering Support Officer, at RAF Gatow.

Mr. S. F. Lee, M.Sc. (Graduate 1976) who has been working as an Associate Engineer with Burroughs Machines Ltd., Glenrothes, since 1975 after completing his M.Sc. course in Manufacturing Technology at Cranfield Institute of Technology, has returned to Hong Kong. He has taken up an appointment as an Assistant Lecturer in the Department of Production and Industrial Engineering at Hong Kong Polytechnic.

Mr. S. Sitsabesan (Associate Member 1976) has joined Ceylon Business Appliances Ltd. as an Electronic Engineer; he was previously working as an Electronics Technician servicing office equipment with the Singer Sewing Machine Company, Colombo.

Obituary

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

Jean Baron (Associate Member 1973, Associate 1968) was killed in an accident at Nice on 28th February 1977. He was 56 years of age and leaves a widow and three children.

Born at Compiègne, Jean Baron received his early education in Morocco, but his university studies in physics at Rabat University were interrupted by mobilization into the French Air Force. After qualifying as a radio technician at Fez, he came to England and for the remainder of the War was a radar maintenance technician with RAF Bomber Command in Yorkshire.

In 1945 Jean Baron returned to France and became Assistant Head of the Electromagnetic Detection Section at the Centre National d'Etude des Telecommunications. From 1949 to 1959 he was initially Chief Engineer and then Assistant Technical Director of Laboratoires R. Derveaux. For the next six years he was with Société SEREL as Chief Engineer and then Technical Manager, and in 1965 he was appointed Technical Director of ROCHAR following its acquisition by Société d'Instrumentation Schlumberger. He rejoined Laboratoires R. Derveaux in 1968 as a senior executive.

During his professional life Jean Baron was concerned with a wide range of projects, from telemetry and radar to digital computers for process control in nuclear engineering and instrumentation generally. One of his principal achievements was the design in the early 'sixties of a computer using transfluxor storage in which all organization was recorded instead of being wired.

John Raymond Brinkley (Fellow 1952, Member 1948) died on 15th August 1977 after a short illness, aged 60 years. He leaves a widow and two sons.

John Brinkley began his career in telecommunications with the British Post Office at Dollis Hill Research Station. During the war he was at the Home Office where he was responsible for the primary development of police, fire and civil defence mobile services. In 1948 he joined Pye Telecommunications and was successively chief engineer, technical director and managing director between 1956 and 1966. During this period he was responsible for many innovations in the mobile radio field and he played a leading role in the introduction of 12·5 kHz channelling in the v.h.f. bands in the U.K.

In 1966 John Brinkley took up an appointment as an Executive Director with Standard Telephone & Cables and in 1969 he was appointed International Manager for mobile radio for ITT. Two years later he joined Redifon Ltd. as an Executive Director and formed the Communications and Marine Division into a subsidiary—Redifon Telecommunications Ltd., of which he became the first Managing Director. He later became also chairman of the company as well as chairman of Redifon Computers Ltd., and in 1976 he was appointed Managing Director of Redifon Ltd. and also Deputy Chairman.

Mr. Brinkley was member of the former Ministry of Posts and Telecommunications' Frequency Advisory Committee and of the Mobile Radio Advisory Committee. He served as a member of the Institution's Council from 1963–66, and was a representative on a B.S.I. Technical Committee for several years. He contributed several papers to the Institution on communications subjects and was the author of numerous articles in the technical press.

Ronald Gledhill (Associate Member 1973, Associate 1958) died on 1st July 1977 at the age of 60 after a few weeks of illness. He leaves a widow and a son.

He was born in Halifax and educated at Tyldesley Boys School in Blackpool. He started his technical education in Blackpool and continued it at Manchester Technical College. From 1939 to 1946 he served in the Royal Signals after which he joined Rediffusion Ltd. (St. Helens Area) as an Audio Equipment Engineer. He became Engineer-in-Charge in 1951 and then Manager of his Area, holding this post for many years.

Ron Gledhill joined the Merseyside Section Committee of the Institution a short time after his election as an Associate. For most of the next eighteen years up to his illness, he held the office of Treasurer. Those who served on the Section Committee during the last two decades, or who have otherwise involved themselves in the activities of the Section, will remember the gentleness, warmth, and unassuming nature of their friend and fellow member.

They will have valued his contributions in Committee and the spirit, good humour and efficiency with which he carried out his tasks for so long in the interests of his Institution. JAC

Philip Peter Reader (Graduate 1963) died on 18th January 1977, at the age of 42 years, leaving a widow and two children.

Following National Service in the RAF as a ground radar fitter, Peter Reader joined the Scientific Civil Service in 1956 and worked at the Radio Research Station, Slough, (now the Appleton Laboratory) as Senior Scientific Officer, initially on atmospheric noise measurements. For three years he was at the RRS's Singapore out-station during the period of the International Geophysical Year and on returning he continued part-time studies at Southall Technical College to complete his Higher National Certificate in Electronic Engineering with endorsements in Advanced Electronics. For the next fourteen years he worked on a wide range of projects, designing and developing and constructing measurement equipment for satellite observations and he was Assistant to the Project Manager for the 1973 Eclipse Project using *Concorde*. At the time of his final illness he was concerned with the International Ultraviolet Explorer satellite (I.U.E.).

Binalendra Prasad Singh (Member 1965) died on 28th March 1977 aged 42 years.

Born and educated at Begusarai, Inia, B. P. Singh studied at the T.N.J. College and obtained the B.Sc. degree of Bihar University in 1955. He then came to England and followed a sandwich course in electronic engineering at Woolwich Polytechnic with industrial experience in the Electrical Engineering Department of Tate and Lyle Ltd., being awarded the B.Sc.(Eng.) degree of London University in 1959. He then worked for a year as a junior design engineer in the Laboratories of EMI Ltd. on community television projects, returning to India in 1960 to take up an appointment as a lecturer in Electronics and Telecommunications at the Bihar College of Engineering, Patna University. Mr. Singh was appointed Technical Manager of the Gramophone Company of India Ltd., Calcutta, with whom he remained until his death.

Applicants for Election and Transfer

The Membership Committee at its meetings on 18th August and 13th September 1977 recommended to the Council the election and transfer of the following candidates. In accordance with Bye-law 23, 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 communication from Corporate Members concerning the proposed elections must be addressed by letter to the Secretary within twenty-eight days after publication of these details.

August Meeting (Membership Approval List No. 237)

GREAT BRITAIN AND IRELAND

CORPORATE MEMBERS

Transfer from Member to Fellow

ATKINSON, Eric James. *Malvern, Worcester and Hereford.*
DIGBY, Peter William. *Harrow, Middlesex.*
DIX, Dennis Lee. *Ruislip, Middlesex.*
WILCOX, David Antony. *Swindon, Wiltshire.*

Transfer from Graduate to Member

NWASOKWA, Joseph Chukwunwike. London.
SANDERS, Peter Philip. *London.*
STEVENSON, Ian James. *Donaghadee, County Down.*
THEAKER, Michael Edward. *Prestatyn, Clwyd.*
WALKER, John Richard. *Caversham, Berkshire.*

Direct Election to Member

CHEATER, William Edgar. *Irby, Wirral, Merseyside.*
KING, John Patrick. *Birkenhead, Cheshire.*
SULEMAN, Mohammad. *Swindon, Wiltshire.*
WOOD, David. *Winchester, Hampshire.*

NON CORPORATE MEMBERSHIPS

Transfer from Student to Graduate

BARKER, Michael Hunt. *Hayes, Middlesex.*
TANYI-TANG, Enoh. *London.*

Direct Election to Associate Member

COYNE, Terence Henry. *Kilcallen, County Kildare, Ireland.*
KARIITHI, Duncan Ngunjiri. *Chelmsford, Essex.*
LOIZIDIS, Nicholas. *Great Sutton, Wirral, Cheshire.*
LUCAS, Anthony James. *Shamley Green, Surrey.*

STUDENTS REGISTERED

CABRERA, Oscar. *London*
INYIAMA, Hyacinth Chibueze. *Swansea.*

OVERSEAS

CORPORATE MEMBERS

Transfer from Member to Fellow

TAN, Hong Siang. *Kuala Lumpur, Malaysia.*

Transfer from Graduate to Member

BRIDGER, Richard Dennis. *Tehran, Iran.*

Direct Election to Member

DOMAN, Uri. *Tel-Aviv, Israel.*

NON-CORPORATE MEMBERS

Direct Election to Graduate

APPADOO, Vroudhaya. *Mauritius.*
TSE, Kai Ming. *Hong Kong.*

Direct Election to Associate Member

MOK, Sng Yam. *Kuala Langat, Selangor, Malaysia.*

STUDENTS REGISTERED

CHAN, Sau Lin. *Singapore.*
CHAN, Yew Jin. *Singapore.*
CHUI, Ying Yeung. *Hong Kong.*
KHOO, Chew Yam. *Singapore.*
KOH, Kay Hoe. *Malacca, Malaysia.*
LEE, Lai Kiong. *Singapore.*
LEONG, Tian Yam. *Singapore.*
LIM, Beng Soon. *Singapore.*
LO, Kam Chuen. Hong Kong.
LOW, Wah Chai. *Singapore.*
NG, Liang Seng. *Singapore.*
NG, Wing. *Hong Kong.*
SIT, Lee-Wai. *Hong Kong.*
TAN, Yiok Teck. *Singapore.*
TSANG, Chung Wai. *Hong Kong.*
WAY, Kok Chay. *Singapore.*
WITHARAMA, G. W. H. P. *Ganegoda, Getahetta, Sri Lanka.*
WONG, Sau Nam. *Hong Kong.*
YEO, Yong Yan. *Singapore.*
YUEN, Yiu Kam. *Hong Kong.*

September Meeting (Membership Approval List No. 238)

GREAT BRITAIN AND IRELAND

CORPORATE MEMBERS

Transfer from Member to Fellow

ROWE, Roy Edward. *Helensburgh, Dunbartonshire*

Transfer from Graduate to Member

HUSBAND, Edward. *Downfield, Dundee.*
SIMPSON, George Niven. *Leighton Buzzard, Bedfordshire.*

Transfer from Associate Member to Member

OBERSBY Derek. *Great Haywood, Stafford.*

Direct Election to Member

DASZKIEWICZ, Anthony. *Newbury, Berkshire.*
HOBBS, Leonard Alan. *Dagenham, Essex.*

NON CORPORATE MEMBERS

Direct Election to Graduate

AMES, Anthony Michael. *Cobham, Surrey.*
OGBOGOH, Fidelis Iweanya. *London.*

Direct Election to Associate Member

BRITTAIN, Nicholas William. *Malvern Link, Worcester and Hereford.*
SACKVILLE, John Edward. *Tividale, Warley, West Midlands.*

OVERSEAS

CORPORATE MEMBERS

Transfer from Graduate to Member

LEE, Yim-Shu. *Hong Kong.*

Direct Election to Member

SHORT, Christopher David. *BFPO 26, Belgium*

NON CORPORATE MEMBERS

Transfer from Student to Graduate

HO, Yu Sum. *Hong Kong.*

Direct Election to Graduate

GEORGAKAKIS, Emmanuel. *Thessalonica, Greece.*
WONG, Kwok-Wing. *Hong Kong.*

Direct Election to Associate Member

CHONG, Yin Ken. *Malacca, Malaysia.*

STUDENTS REGISTERED

CHAM, Yoke Oi. Singapore.
CHENG, Ping-Chin. *Hong Kong.*
JASTHI, Ravi Shankar. *Rajahmundry, India.*
LAU, Kwong Kuen. *Hong Kong.*
LEE, Kum Fun. *Singapore.*
LEUNG, King Fai. *Hong Kong.*
LIM, Eng Hian. *Singapore.*
ONG, Hui Lim. *Singapore.*
PANG, Kwok Wai. *Hong Kong.*
TAN, Gek Eng. *Singapore.*
YU, Him Chi (Joseph). *Hong Kong.*

Forthcoming Institution Meetings

Tuesday, 1st November

AUTOMATION AND CONTROL SYSTEMS GROUP

Colloquium on LABORATORY AUTOMATION

Royal Institution, Albemarle Street, London W1, 2 p.m.

Advance registration necessary. For further details and registration forms apply to Meetings Officer, IERE.

Thursday, 3rd November

JOINT IERE/IEE MEDICAL AND BIOLOGICAL ELECTRONICS GROUP IN ASSOCIATION WITH SOUTH WALES SECTION

Colloquium on MODERN TRENDS IN THE ASSESSMENT OF CARDIO-PULMONARY FUNCTION

Welsh National Medical School, Cardiff, 10 a.m.

Advance registration necessary. For further details and registration forms apply to Meetings Officer, IERE.

Tuesday, 15th November

COMPONENTS AND CIRCUITS GROUP

Colloquium on ANALOGUE FILTERS

Royal Institution, Albemarle Street, London W1, 10.30 a.m.

Advance registration necessary. For further details and registration forms apply to Meetings Officer, IERE.

Tuesday, 22nd November

AEROSPACE, MARITIME AND MILITARY SYSTEMS AND COMMUNICATIONS GROUPS

Colloquium on PORTABLE COMMUNICATIONS SYSTEMS

Royal Institution, Albemarle Street, London W1, 2 p.m.

Advance registration necessary. For further details and registration forms apply to Meetings Officer, IERE.

Tuesday, 29th November

JOINT IERE/IEE COMPUTER GROUP

Colloquium on ELECTRONIC SECURITY AND PERSONAL ACCESS SYSTEMS

Royal Institution, Albemarle Street, London W1, 10.30 a.m.

Advance registration necessary. For further details and registration forms apply to Meetings Officer, IERE.

Beds and Herts Section

Thursday, 27th October

Automobile electronics—some aspects of research

By Dr. R. D. Codd (*Joseph Lucas*)

Synopsis: Electronics for the automotive industry has been the subject of discussion for many years. However, electronics on the motor car, excluding in-car entertainment, is still very limited. The aim of this talk is to give an indication of the amount of research work that has to be undertaken before a potential product in its first prototype form can be fitted to a car. Examples of various aspects of electronic circuit development will be discussed with comments relating to cost and reliability requirements of the various technologies currently available to the circuit designer.

Luton College, Luton, Beds., 7.45 p.m. (Tea 7.15 p.m.).

Thursday, 24th November

Electronic calculators—past and current technology

By C. N. Peart (*Commodore Business Machines*)

Hatfield Polytechnic, Hatfield, Herts. 7.45 p.m. (Tea 7.15 p.m.).

Thames Valley Section

Thursday, 17th November

Viewdata

By K. E. Clark (*Post Office*)

Caversham Bridge Hotel, Reading, 7.30 p.m.

Kent Section

Thursday, 17th November

JOINT MEETING WITH IEETE, IPOEE AND MEDWAY AND MAIDSTONE COLLEGE OF TECHNOLOGY

London's telecommunications—switchboards to satellites

By K. Ford (*P.O. Telecommunications*)

Lecture accompanied by an exhibition of subscribers' apparatus (old and new) and also other telecommunications working systems including microwave equipment, cables and Post Office Tower model etc.

Medway and Maidstone College of Technology, Horstead, Chatham, Kent, 7 p.m. (Tea 6.30 p.m.).

Southern Section

Tuesday, 25th October

JOINT MEETING WITH IMA

Error correctiona— simple Golay decoder

By R. A. Croft (*Plessey Co.*)

Mountbatten Theatre, Southampton College of Technology, 6.15 p.m.

Thursday, 27th October

Quadraphony

C. Daubney and R. Collins (IBA)

Synopsis: The paper outlines the development of sound reproduction from the first gramophone systems through stereo to the various current proposals for quadraphony. Both matrix and so-called discrete systems are considered and suggestions made for the requirements which a broadcast system would need to meet if it were to obtain widespread acceptance in Europe. Following from these suggestions some remarks are made concerning the systems which are preferred for broadcasting purposes.

Crawley College of Technology, 7 p.m.

Wednesday, 2nd November

JOINT MEETING WITH IEE

Symposium on The Planning and Design of the IBA's Radio and Television Transmitter Network

Three papers by R. Byrne, D. S. Chambers and R. Wellbeloved

IBA, Crawley Court, Winchester, 3 p.m.

Thursday, 3rd November

Charge coupled devices for analogue signal processing

By C. P. Traynar (*University of Southampton*)

Farnborough College of Technology, 7 p.m.

Tuesday, 8th November

Microcomputers and their applications

By H. Kornstein (*Intel Corporation* (*UK*))

Synopsis: This paper will provide a brief review of the motivation in using microcomputers and some of the past history of the state of these devices. Some of the major trends in the microcomputer industry will also be presented, particularly significant trends in circuit integration level, prices and performance. Examples of some modern third-generation microcomputer components will be given including a discussion of the single-chip microcomputer card set. The paper will then deal with the various applications of microcomputers and will include a specific example of a microcomputer based system utilising the process industry. Some of the hardware and software considerations pertinent to this system will be reviewed.

School of Signals, Blandford Camp, 6.30 p.m.

Wednesday, 9th November

Holographic memories

By P. Waterworth (*Plessey Microsystems*)

Synopsis: A brief description of holographic memories will be given. The major technological problems in any holographic system will be described, and it will be shown that the available storage media and input devices are the main limitations. Holographic techniques will be compared with both conventional imaging and magnetic recording systems. The main advantages of holographic techniques over these systems will be described with reference to work at Plessey. Finally a description of a holographic mass storage device being developed by Plessey will be given, together with some of the applications for this system.

Lanchester Theatre, University of Southampton, 7 p.m.

Friday, 18th November

Charge coupled devices for analogue signal processing

By C. P. Traynar (*University of Southampton*)

Isle of Wight College of Arts and Technology, 7 p.m.

Wednesday, 23rd November

Switched mode power supplies

By P. Chapman (*Marconi-Elliott*)

Room ABO11, Portsmouth Polytechnic, King Henry I Street, Portsmouth, 7.30 p.m.

Wednesday, 30th November

JOINT MEETING WITN IEE

The use of microprocessors in numerical control

By Dr. V. Latham (*University of Southampton*)

Lanchester Theatre, University of Southampton, 6.30 p.m.

East Anglian Section

Thursday, 27th October

JOINT MEETING WITH IEE

C.T. scanning of the brain and body

By P. McAtamney (*EMI*)

Engineering Laboratories, Trumpington Street, Cambridge, 6 p.m. (Tea 5.30 p.m.).

Wednesday, 9th November

CEI LECTURE—

The engineer's contribution to a better world

By Prof. M. W. Thring (*Queen Mary College*)

Engineering Laboratories, Trumpington Street, Cambridge, 7 p.m. (Tea 6.30 p.m.).

Wednesday, 16th November

Electronic ignition

By Dr. M. J. Werson (*University of Southampton*)

Cossor Electronics, Elizabeth Way, Harlow Essex, 6.30 p.m. (Tea 6 p.m.).

Thursday, 24th November

JOINT MEETING WITH IEE

Development of miniature TV receivers

By Clive Sinclair (*Sinclair Radionics*)

Engineering Laboratories, Trumpington Street, Cambridge, 6 p.m. (Tea 5.30 p.m.).

Wednesday, 30th November

JOINT MEETING WITH IEE

Switched mode power supplies

By P. Chapman (*Marconi-Elliott*)

Civic Centre, Chelmsford, 6.30 p.m. (Tea 6 p.m.).

South Western Section

Wednesday, 26th October

CEEFAX—A new form of broadcasting

By J. P. Chambers (*BBC*)

Chemistry Lecture Theatre No. 4, University of Bristol, 7 p.m. (Tea 6.30 p.m.)

Tuesday, 1st November

JOINT MEETING WITH IEE

Special effects on TV

By A. B. Palmer (*BBC*)

The College, Swindon, 7 p.m. (Tea 6.30 p.m.)

Tuesday, 1st November

JOINT MEETING WITH IEE

Software engineering applied to stored program control (SPC)

By G. Owens (*GEC*)

Main Lecture Theatre, Plymouth Polytechnic, 7 p.m. (Tea 6.30 p.m.).

Wednesday, 16th November

Page facsimile transmission

By D. F. Banks (*Muirhead*)

Room 5W2.4, University of Bath, 7 p.m. (Tea 6.30 p.m.).

East Midlands Section

Tuesday, 8th November

JOINT MEETING WITH IEE

Future telephone switching

By J. R. Pollard (*Plessey Telecommunications*)

Department of Electronic and Electrical Engineering, Loughborough University of Technology, 7 p.m. (Tea 6.30 p.m.).

South Midlands Section

Thursday, 10th November

The voltage to current transactor (VCT)

By Professor W. Gosling (*University of Bath*)

Majestic Hotel, Cheltenham, 7.30 p.m.

West Midlands Section

Thursday, 17th November

JOINT MEETING WITH CEI/IME

Energy resources and the mining engineer

By L. J. Mills (*NCB*)

Synopsis: The paper describes the evolution of the engineer with particular reference to the coal mining engineer. The contribution to coal mining by other specialists are illustrated and an account is given of technical achievements to date together with a look at the future. The present and future place of the industry in meeting the country's energy needs are also described.

Haworth Lecture Theatre, University of Birmingham, 6.30 p.m.

South Wales Section

Wednesday, 16th November

JOINT MEETING WITH IOP

Charge coupled devices and their applications

By Prof. J. Beynon (*UWIST*)

Room 112, Applied Physics Department, UWIST, Cathays Park, Cardiff, 6.30 p.m. (Tea in Refectory 5.30 p.m.).

North Eastern Section

Tuesday, 8th November

Automated intelligence for an interstellar probe

By Dr. W. F. Hilton (*Consultant*)

Synopsis: Interplanetary probes have almost become routine engineering in the USA. Construction of the next class of space craft, the Shuttle, proceeds apace, both in the USA and in Europe, to the virtual exclusion of the UK. Can we prepare expertise for the next great forward step in space? Contracts for an interstellar probe will be let in the next decade. The object of this lecture is to stimulate discussion by defining the new problem areas, in the belief that solutions can be found, leading to hardware and software.

YMCA, Ellison Place, Newcastle upon Tyne, 6 p.m. (Tea 5.30 p.m.).

North Western Section

Thursday, 20th October

Quadraphonics

By Dr. K. Barker (*Sheffield University*)

Synopsis: The lecture will introduce the subject of quadraphonics through the historical development of sound recording and will illustrate the subject extensively with visual and audio demonstrations. All presently available quadraphonic systems will be covered including some new aspects of multi-channel broadcasting.

Bolton Institute of Technology, Deane Road, Bolton, Lancs, 6.15 p.m. (Tea 5.45 p.m.).

Thursday, 17th November

Philosophy of maintenance

By Col. W. Barker (*Ministry of Defence*)

Renold Building, UMIST, Sackville Street, Manchester, 6.15 p.m. (Tea 5.45 p.m.).

Yorkshire Section

Wednesday, 26th October

Optical fibre communication systems

By Dr. D. H. Newman (*Post Office Research Dept.*)

Synopsis: The state of development of fibre systems for civil telecommunications networks and the performance requirements of optical sources for such systems are reviewed. The design and technology of suitable semi-conductor sources, current problem areas and future trends are discussed.

Leeds Polytechnic, 5 p.m.

Thursday, 27th October

CEI LECTURE—

Nuclear power

By E. S. Booth (*Yorkshire Electricity Board*)

St. Georges Hall, Bradford, 7 p.m.

Merseyside Section

Wednesday, 9th November

Marine gas turbine control

By W. S. Brown (*Hawker Siddeley Dynamics*)

Department of Electrical Engineering and Electronics, University of Liverpool, 7 p.m. (Tea 6.30 p.m.).

Northern Ireland Section

Wednesday, 2nd November

Energy conservation

By A. Kane (*Post Office*)

Cregagh Technical College, 7 p.m.