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 46 No. 10

October 1976

The Radio and Electronic Engineer

The Journal of the Institution of Electronic and Radio Engineers

Seventy Five Years of British Standards

IT may come as a surprise to some to learn that the British Standards Institution was founded as long ago as 1901. The oldest national standards organization in the world, it commenced life as the Engineering Standards Committee comprising members of the Institutions of Civil and Mechanical Engineers, and of Naval Architects, and the Iron and Steel Institute, and its earliest achievements, standardizing tramway rails and structural steel sections, are recorded as having led to spectacular reductions in the number of sizes. Perhaps even more striking and convincing, estimated savings in production costs for steel sections in one year were £1M.

Activity started on standardization in the radio (or wireless) industry in 1925, the year of the Institution's foundation, with the dimensions of plug-in coils, and in the next decade or so made rather slow progress. The Second World War gave a great impetus to standardization in radio components, just as the First World War had done in aircraft engineering, encouraged no doubt by the Minister of Munitions—Winston Churchill—who believed that the need for standardization of Allied aircraft materials was 'based on principles so obvious that they really do not at this time of day require to be emphasized.'

International standards work in the electrical field does in fact go back to 1904 when, under the presidency of Lord Kelvin, the International Electrotechnical Commission was formed, predating by well over thirty years the other international standardizing bodies. Today, the European Economic Community has introduced a portmanteau term 'harmonization', much of which is the standardization of engineering products and it is into this international work that the bulk of BSI's effort now goes.

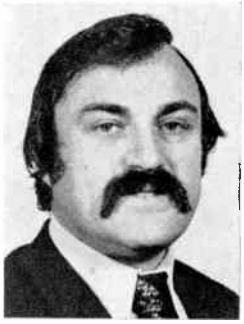
Ever since the War, the IERE has supported BSI, both as a financial subscriber and by appointing members to serve on relevant industry and technical committees, more especially those where an independent voice can contribute to the resolving of manufacturer and user problems. In addition, many more members represent trade or user associations on drafting committees. The monumental BS 9000 series—Electronic components of assessed quality—has dominated the work of many standards engineers both before and since the first parts began to appear in 1969, and this is forming an invaluable basis for the future 'harmonized' Common Market standards.

Closely linked with standardization, as the title of BS 9000 implies, is quality and it is interesting to see that the programme of the recent 75th Anniversary Conference 'British Standards for good design' included a lecture on 'British Standards and design for quality assurance' by Dr. Percy Allaway, President of the IERE, and a past chairman of the National Council for Quality and Reliability. Significant, too, was a lecture on 'British Standards and design for ease of maintenance' by the Director of Engineering of the British Steel Corporation, Mr. H. Darnell.

It is self-evident truth, as Winston Churchill implied, that good design and standards are complementary, although perhaps it is one of those truths found uncomfortable by many engineers. Let the last word therefore come from the introduction to the Conference programme by HRH The Duke of Edinburgh:

'Good design is the result of a mixture of almost equal parts of experience and originality. You have to be born with originality but experience can be acquired . . . People will always make mistakes, and anyone in a creative occupation such as design is more than usually liable to do so, but there is no excuse for anyone making mistakes which have been made before. Designers who stick to British Standards need have no fear of making old mistakes over again.'

Contributors to this issue*



Richard Jones (Member 1970, Graduate 1965) served a five-year apprenticeship with the Marconi Company during which time he took the Higher National Certificate and subsequently obtained endorsements to gain him Graduate membership of the Institution. At the end of his apprenticeship in 1961, Mr. Jones was employed as a development engineer working in the Aeronautical Division of the Marconi Company at Basildon.

In 1966 he transferred to a special projects team which eventually became the Electro-Optical Systems Division and is now part of Marconi-Elliott Avionic Systems Ltd. At that time he was controlling the development of equipment associated with the transmission of digital information between remote bodies. He has since headed a large design team working on sonar equipment.

Mohammad Aslam (Member 1974, Graduate 1967) served an Engineering Apprenticeship with the Royal Air Force from 1951 to 1954 and was with the Pakistan Air Force for the next nine years, latterly as an Instructor at the School of Electronics. In 1958 he obtained the Full Technological Certificate in telecommunication engineering from the City and Guilds of London Institute. After leaving the PAF Mr. Aslam worked as a microwave communication system engineer in Iran, and then joined Standard Telephones and Cables, Paignton, as an electronics engineer in 1966. Four years later he became a project engineer with Microwave Associates, Luton. In 1969 he gained the Post-Graduate Diploma in the physical basis of electronics from the North London Polytechnic and he was awarded an M.Sc. in electronics from Loughborough University of Technology in 1970.

From 1970 to 1976 Mr. Aslam was a Principal Engineer with Plessey Telecommunications, Beeston, leading a project on high-speed p.c.m. and digital line systems. He is now with a consultancy company in Riyadh, Saudi Arabia.

Raymond Steele completed his H.N.C. in electrical engineering and an apprenticeship in Radio Engineering at E.K. Cole Ltd. in 1955 and spent the next four years as a student at King's College, University of Durham where he obtained a degree in electrical engineering. He then returned to E.K. Cole Ltd., Southend-on-Sea, in 1959 for two years in a research capacity, and then joined Cossor Radar and Electronics Ltd., Harlow, Essex, as a development engineer on radar displays. In 1962 he became a research and development engineer with the Aeronautical Division of the Marconi Company Ltd., at Basildon, Essex, where he did research on magnetic logic and on navigational computers.

Dr. Steele entered the teaching profession in 1965 as a Lecturer at the Royal Naval College, Greenwich and lectured on telecommunications on the NATO and Commonwealth Officer's course and on the External London University Degree Course. Since 1968 he has been at Loughborough University of Technology where he is a Senior Lecturer in the Department of Electronic and Electrical Engineering and a member of this Department's digital communications research group. He was awarded his Ph.D. degree by Loughborough University two years ago. His main research activities are in delta modulation systems and he published the first book on this subject in January 1975.

David Ormerod served an apprenticeship with the Marconi Company and in 1966 obtained a Bachelor of Technology Degree in electrical engineering at Loughborough University of Technology. In 1967 after taking a post graduate course at Marconi College he joined the Baddow Research Laboratories and worked on digital signal processing techniques. He transferred to the Electro-Optical Systems Division in 1968 and was involved with high-integrity digital command systems. Two years later he was promoted to Principal Engineer and is now concerned with passive and active low-light television sensors.

Sidney Hewitt graduated in physics at Bristol University in 1952, and received a Ph.D. degree at Reading University in 1956 for research in electron diffraction. From then until 1959 he worked at EMI Research Laboratories, Hayes, participating in the design and development of high-power klystron amplifiers. Dr. Hewitt joined the Roke Manor research laboratories of the Plessey Company in 1960, where he has since been involved in a variety of microwave projects.

Raymond Pengelly graduated in 1969 from Southampton University with a B.Sc. (Hons) degree in physics. He then worked on r.f. front-end design and various low and high frequency ferrite circuits as a research engineer for the Plessey Company's research laboratories at Roke Manor, Romsey, Hampshire. In 1972 he obtained an M.Sc. in electronics at Southampton University, returning to Roke Manor to continue research into microstrip propagation. In the last three years, as a senior microwave engineer, he has been involved in microwave research into areas such as p-i-n diode attenuators, lumped-element circuits and GaAs f.e.t. amplifiers, and has written several papers on these subjects.

* See also pages 486 and 496.

UDC 621.3.049.774/6.029.64

Design techniques for integrated microwave amplifiers using gallium arsenide field-effect transistors

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Based on a paper given to a Joint IERE/IEE Southern Sections' meeting held at Newport, Isle of Wight, on 14th November 1975.

SUMMARY

This paper reviews the technology and methods used in the design of integrated microwave transistor amplifiers, and describes some examples in which gallium arsenide field effect transistors are used in both hybrid and monolithic forms of integrated circuit.

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1 Introduction

One of the most striking advances in microwave technology during the past decade has been in the design of microwave transistor amplifiers. This has been brought about by simultaneous developments in transistors, circuit technology and techniques for design and measurement—developments which have tended to reinforce one another. As a result, these amplifiers can now be used, with obvious advantages, to replace travelling-wave tubes in many applications and are clearly of great interest to the system designer.

This paper is intended to serve as a short introduction to the subject, including relevant aspects of microwave integrated circuit technology, and will be illustrated with examples of broad-band amplifiers using gallium arsenide field-effect transistors. The emphasis will be put on the especially challenging task of designing amplifiers to meet a closely-specified performance over a broad band at the higher microwave frequencies. For these, an integrated form of construction is needed, as indicated later.

It is convenient to start by pointing out some of the important differences between these circuits and conventional r.f. amplifiers at lower frequencies.

(a) The internal feedback within the transistors is *always* significant. This means that the stability of the circuit must be monitored carefully, and that stage-by-stage design is often inadequate if the best gain-bandwidth product is to be obtained in a multi-stage amplifier.

(b) Printed transmission lines or miniature lumped elements are used in the circuit. Any stray reactances become very important at the higher microwave frequencies, and dimensions become critical.

(c) The assembly techniques used have some similarity to those used in monolithic integrated circuits.

In the earlier part of this paper, the capabilities of present-day commercially-available transistors will be reviewed briefly, and the possible forms of circuit will be described. This will be followed by a summary of the design procedure used for broad-band amplifiers.

2 Microwave Transistors

The microwave transistors[1–7] currently in use can be divided into two main categories: silicon n-p-n bipolar transistors, with a wide range of power capabilities, and gallium arsenide field-effect transistors (GaAs f.e.t.), commercial versions of which at present are limited to small-signal applications.

The silicon transistors represent an extension upwards in frequency of the technology used to make conventional r.f. or fast logic devices, so the devices themselves will not be described in detail here.

The operation of the GaAs f.e.t.[8–10] differs from that of the bipolar transistor, and is conceptually fairly simple (Fig. 1). The electron current travels from the source to the drain electrode through a thin epitaxial

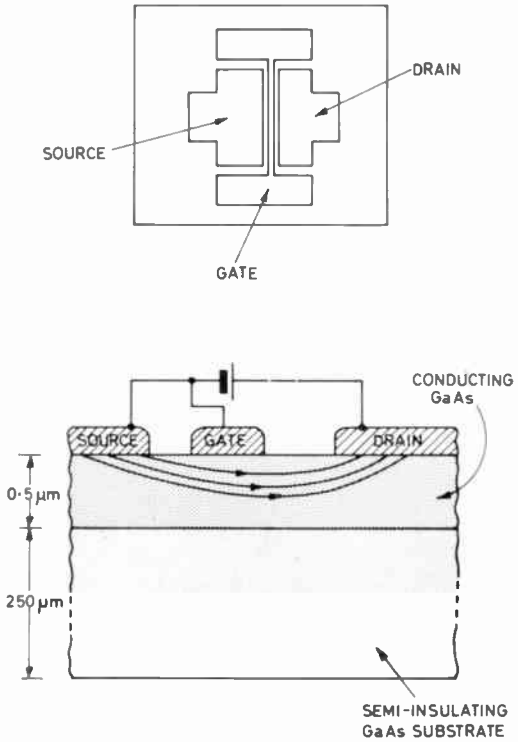


Fig. 1. Gallium arsenide field effect transistor.

conducting layer of gallium arsenide which is deposited on a semi-insulating gallium arsenide substrate. The current is controlled by the voltage on the gate electrode, which does not contribute any direct current. Satisfactory amplification is obtained with zero d.c. bias between source and gate—a feature which simplifies circuit design. However, for minimum noise figure, a small negative gate bias is required. In microwave f.e.t.s, the inter-electrode spacing and the gate length are typically 1–2 μm.

The gain and noise figure claimed for the best commercially available bipolar and field-effect transistors are plotted in Fig. 2. The noise figures are similar up to 6 GHz, but the GaAs f.e.t. has a greater available gain, giving it a marked superiority at higher frequencies; recent work[11] indicates that the difference will increase with further development. However, there are other factors to be considered which tend to restore the

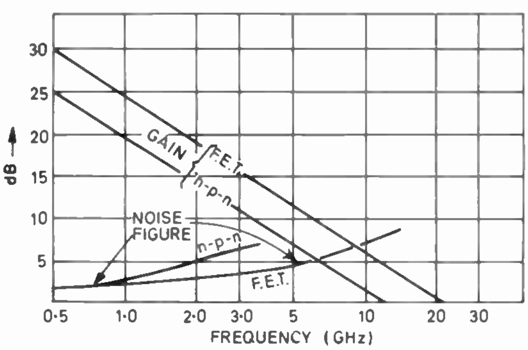
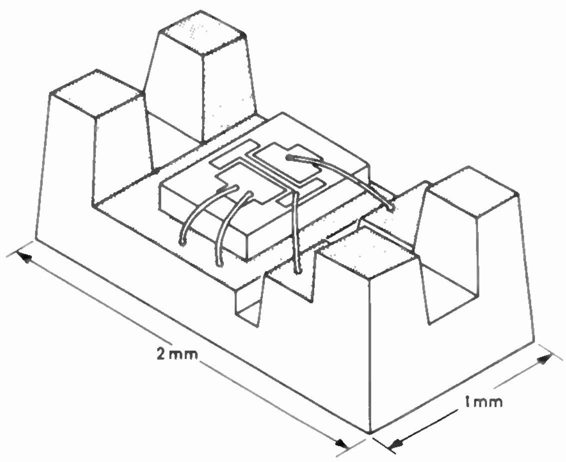


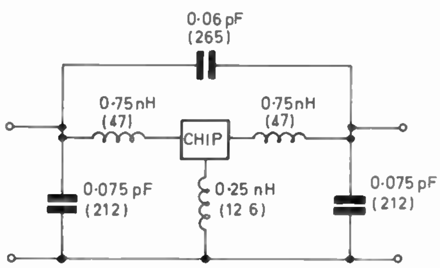
Fig. 2. Performance of microwave small-signal transistors.

balance; at the lower microwave frequencies, the f.e.t. has high impedance levels which make it difficult to match to transmission line impedances over a broad band. Silicon bipolar transistors, on the other hand, are very convenient for this purpose, so that the two types should be regarded as complementary.

Packages for microwave transistors are usually provided with ribbon leads for direct connection to printed transmission lines. The dielectric material used in these is generally alumina or beryllia to provide good heat conduction and low power loss. Mechanically and thermally these packages are very convenient, but the parasitic reactances associated with them make them unsuitable for broad-band circuits at the higher microwave frequencies. Of more use is the much smaller 'leadless inverted device' (l.i.d.) package, shown in Fig. 3(a). The package comprises a small ceramic carrier, which can be bonded in an inverted position directly on to a printed circuit.



(a) Transistor in l.i.d. package.



(b) Equivalent circuit of l.i.d. package.

Fig. 3.

Even this package has substantial reactances, as shown by the equivalent circuit of Fig. 3(b). The figures in brackets are reactance values in ohms, at 10 GHz. Design studies on the 8–12 GHz amplifiers described later in this paper showed that the l.i.d. package would be unacceptable, and that an integrated form of circuit was necessary, using unpackaged chips.

3 Circuit Construction

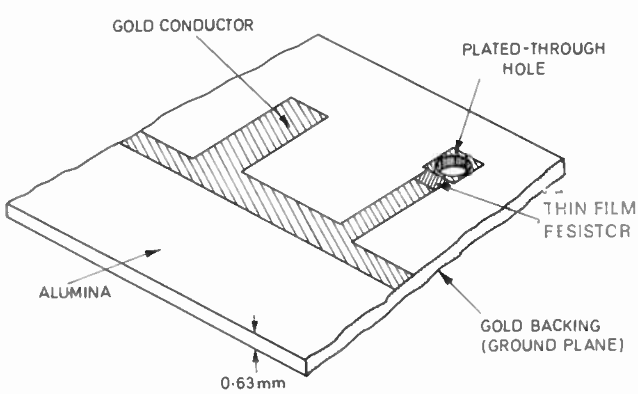
Ceramic substrates are generally used for microwave integrated circuits, since they are small, rugged and convenient for the attachment of semiconductor components. The circuit elements commonly used are printed transmission lines (in the form of open 'microstrip'), printed inductors and capacitors, chip capacitors and printed resistors, and typical circuit layouts incorporating these are shown in Fig. 4. The ceramic is usually alumina, either sintered or in single-crystal form as sapphire. The high value of relative permittivity (about 9·6) reduces the wave velocity in microstrip to well below the free-space value.

The lumped circuit elements shown in Fig. 4(b) have the advantages of compactness and wide impedance range compared with microstrip. However, they are more lossy, and each element will have at least two parasitic reactances associated with it.[12–14] The latter will usually need to be taken into account in circuit design, so that the full equivalent circuit of a layout such as that in Fig. 4(b) will be quite complex.

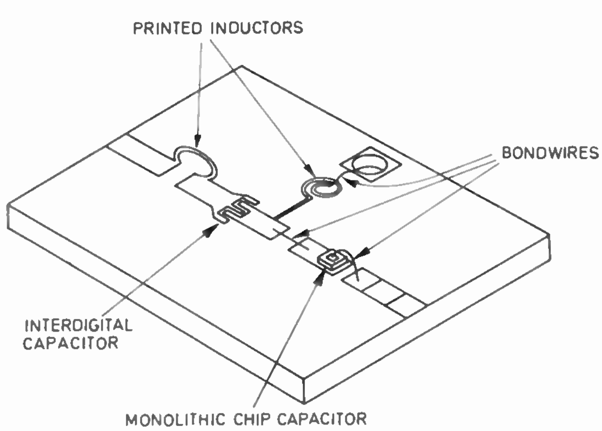
The choice of passive circuit layout is thus a matter for detailed numerical work combined with engineering judgement. The design of the passive circuits will normally be in two phases: preliminary investigations leading to a chosen circuit topology, followed by a calculation of the required numerical values.

4 Transistor Characterization

The design of a microwave integrated amplifier frequently begins with measurements of the characteristics of the transistors to be used. This is necessary because there will always be some stray reactances in the immediate vicinity of a transistor whose values will be determined by the exact method of mounting; it is therefore prudent to mount the transistor as in the final circuit while it is being characterized.

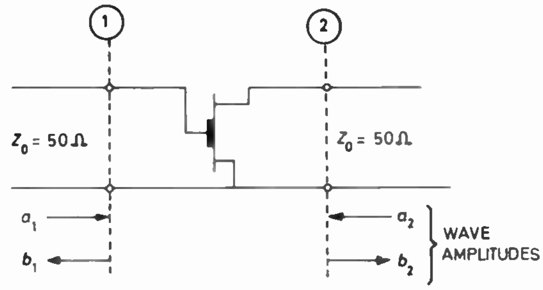


(a) Distributed microstrip circuit.



(b) Lumped ceramic circuits.

Fig. 4.



$$\begin{pmatrix} b_1 \\ b_2 \end{pmatrix} = \begin{pmatrix} S_{11} & S_{12} \\ S_{21} & S_{22} \end{pmatrix} \begin{pmatrix} a_1 \\ a_2 \end{pmatrix}$$

Fig. 5. Scattering matrix of transistor.

At microwave frequencies, the transistor behaviour is expressed as a set of scattering (S-) parameters,[15] used because they are the most convenient to measure. The measurement is carried out with a network analyser, in which the transistor is mounted in a 50 ohm transmission line between a signal source and a matched load. The network analyser has provision for interchanging these two with respect to the transistor. The scattering parameters relate the wave amplitudes b_1 and b_2 scattered from the transistor terminals to the amplitudes a_1 and a_2 incident at those terminals, as shown in Fig. 5. The electrical significance of the S-parameters can easily be seen: S_{11} is the input reflection coefficient when the output is terminated in 50 ohms; S_{22} is the corresponding output reflection coefficient; S_{21} is the amplitude gain for the the forward travelling wave, and in fact $|S_{21}|^2$ is the power gain under these conditions. S_{12} represents the corresponding reverse gain, and is a measure of the internal feedback in the transistor. In a microwave transistor, $|S_{21}|$ may not be much greater than 1, the usable power gain being obtained from the differing input and output impedance levels.

Figure 6 shows an example of a jig in which a transistor chip is mounted on microstrip for S-parameter measurements. In this case, the measured S-parameters will include the effects of stray capacitances and bond-wire inductances near the chip.

5 Power Gain and Stability

The scattering parameters provide a direct indication of transistor input and output impedances, and can also be used to derive other basic design information. The *stability factor* can be defined in a number of equivalent ways; for example, the Rollett[16] stability factor is:

$$K = \frac{2|S_{12}S_{21}|}{1+|S_{11}S_{22}-S_{12}S_{21}|^2-|S_{11}|^2-|S_{22}|^2}.$$

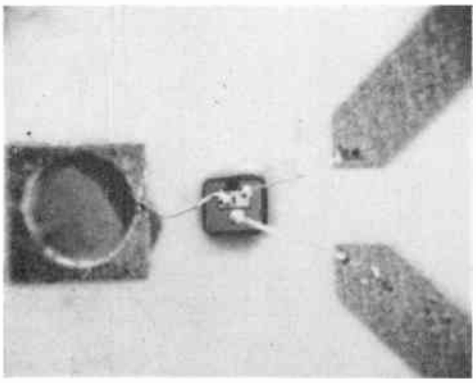


Fig. 6. Microstrip transistor characterization substrate.

If K is greater than 1 at a particular frequency, the transistor will not oscillate at that frequency for any passive loading, and the *maximum available gain* can be calculated:

$$\text{m.a.g.} = \frac{|S_{21}|}{|S_{12}|}(K-\sqrt{K^2-1}).$$

In the limiting case where $K = 1$ and the transistor is just stable, the maximum available gain becomes the *maximum stable gain*:

$$\text{m.s.g.} = \frac{|S_{21}|}{|S_{12}|}.$$

If K is less than 1, the transistor will oscillate under some loading conditions, and a maximum available gain cannot be defined. In these circumstances, the maximum stable gain can still be used as an index of transistor capability. (Reference 17 gives a detailed treatment of this subject.)

6 Broad-band Amplifier Design

Many microwave systems require amplifiers which provide prescribed frequency-independent gain over a broad band and a good impedance match to 50-ohm circuits. The procedure for designing this type of amplifier will now be described briefly.

From the calculated power gain per transistor, the total number of stages needed to provide the specified gain can be estimated. If more than two or three stages are required, it is usual to split the circuit into a cascade of similar modules. The design is thereby simplified and manufacturing problems may also be eased.

The simplest method for choosing the circuit topology is to ignore the transistor internal feedback for a first approximation, and to design impedance-matching networks to suit the simplified transistor input and output impedances. In choosing the circuit, the designer must note a number of constraints, including:

(a) The r.f. circuit must obviously be consistent with the d.c. requirements of the transistor.

(b) The practical limitations on circuit elements, as described in Section 3, must be observed.

(c) The circuits should be kept as simple as possible to minimize unwanted resistive loss and tolerance problems.

(d) If there is a danger of instability, the circuit must be designed to quell the gain at critical frequencies, e.g. by the inclusion of frequency-dependent ohmic loss.

The outcome of these studies might perhaps be a circuit of which a typical stage is shown in Fig. 7. This example uses predominantly transmission-line matching elements.

It is at this point in the design process that computer optimization is most useful and often essential.[19,20] When applied to amplifier design, the procedure is to supply a coded description of the circuit to the computer, together with a set of numerical 'starting' values for the circuit elements, tables of the transistor S-parameters and a description of the desired performance. The computer analyses the circuit performance at a set of frequencies in the required band, and calculates the 'objective function', which measures the discrepancy between the desired performance and that obtained from the trial circuit. This function (F) will usually be a weighted combination involving gain and impedance matching, as illustrated in Fig. 8. The computer then systematically adjusts the circuit values, re-analysing the circuit at each stage, so as to reduce the objective function until an acceptable result is reached. The numerical values of the final circuit are then printed out.

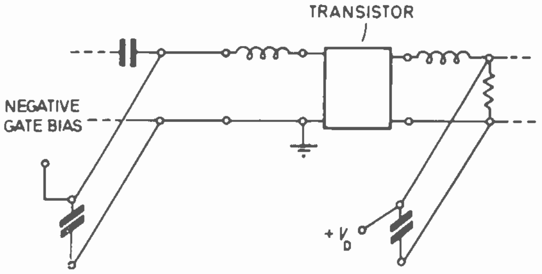


Fig. 7. Typical amplifier stage.

Experience has shown that if computer optimization of amplifier design is to be efficient and economical, the circuit designer must be able to monitor progress and to intervene where necessary. Moreover, the success of the optimization depends in a number of ways upon the judgement of the designer. For example, if the demanded performance is beyond the capability of the circuit, the optimization will obviously fail. Conversely,

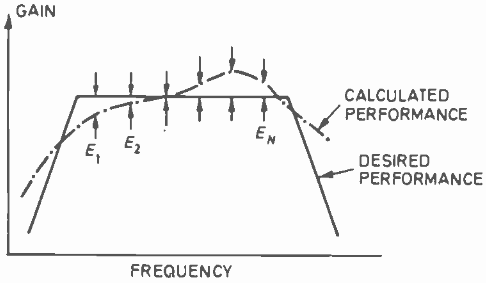


Fig. 8. Objective function.

$$F = W_1 \sum_1^N E_i^2 + W_2 \sum_1^N |\rho_i|^2$$

ρ_i = reflection coefficient
W_1, W_2 = weighting functions

if the demanded gain is too low, for example, the computer will tend to 'dump' excess gain by impedance mismatching, again leading to an unsatisfactory result.

The choice of starting values is also very important. The objective function is a non-linear function of many circuit values, and will exhibit many local minima with respect to changes in these elements. The best circuit design corresponds to the lowest minimum, of course, and it may be necessary to re-run the optimization several times with differing sets of starting values to ensure that this has been found.

Computer optimization of circuit design is founded on the assumption that the final practical circuit will faithfully represent that which has been analysed. Possible causes of discrepancy are:

(i) spread of transistor characteristics,

(ii) measurement error during transistor characterization, and

(iii) errors in the design of circuit layout (resulting from inadequate allowance for junction effects and stray reactances, for example).

The manufacture of microwave transistors is of necessity a closely controlled process, and this gives a high degree of consistency in performance. It has been found that, for frequencies up to X-band, a gain tolerance of $\pm 0{\cdot}1$ dB per transistor can be achieved without unduly rigorous selection. The factors (ii) and (iii) may contribute systematic errors of the same order, and to achieve this accuracy it will usually be necessary to check parts of the passive circuit by S-parameter measurement, applying corrections where necessary.

7 Some Examples of Broadband Amplifier Design

Some examples of amplifier design will now be described, to illustrate the design process and the practical solutions in more detail.

The examples chosen describe two approaches taken to produce X-band GaAs field-effect transistor amplifiers in the 6 to 12·4 GHz range. The hybrid amplifiers use lumped matching elements produced by the photolithographic process on alumina or sapphire, whilst the monolithic amplifiers use smaller lumped elements fabricated on semi-insulating GaAs together with the f.e.t. The design and construction of these amplifiers will now be described together with their relative performances.

The design requirements for these amplifiers were: gain constant to within ± 1 dB over the frequency range 8–12 GHz and input and output standing wave ratios (v.s.w.r.) less than 2·5 : 1. In each case, the Plessey GAT3 Schottky barrier field-effect transistor has been used. This f.e.t. has a 120 μm wide gate, 1 μm long with a buffer layer between the epilayer and the bulk gallium arsenide, thus giving near-bulk-mobility values.[21]

Bare-chip devices have been used in conjunction with microstrip and lumped inductors and capacitors on alumina ceramics for the hybrid amplifiers. Since in the monolithic amplifiers described later the S-parameters of the intrinsic transistor are required, the bare-chip 2-port small-signal scattering parameters need to be accurately measured. This has been done using an automatic network analyser together with off-line computer programs which correct the measured S-parameters for jig parasitics, wire bond inductances, end-effect capacitances, microstrip to coaxial transitions, etc. Three-port S-parameters are similarly corrected and used to evaluate the common lead reactances.

Such measurements allow evaluation of the maximum available gain, m.a.g., maximum stable gain, m.s.g. and stability factor, K of the intrinsic device. These parameters are given in Table 1. Separate measurements on m.a.g. and m.s.g. act as confirmation of successful S-parameter analysis.

The noise figure of the device is also measured under varying gate to source voltages and a minimum figure for the device at room temperature is found. The corresponding S-parameters of the device under such minimum noise conditions are then measured and computer-corrected. The minimum noise figure of the device is approximately 5 dB at 10 GHz.

Table 1. Gains and figures of merit of Plessey GAT3 f.e.t.

Frequency (GHz)	Max. stable gain (dB)	Max. available gain (dB)	U (dB)	K
8	12·26	8·84	11·46	1·46
9	11·25	6·99	10·37	1·50
10	10·03	6·92	8·42	1·48
11	8·9	5·92	8·67	1·25
12	8·6	5·59	7·06	1·19

Using a derived equivalent circuit for the f.e.t. similar to that recently published[22] a Chebyshev impedance matching network can be postulated at the f.e.t. input (Fig. 9). Starting values for the computer optimization routine can be found by assuming the transistor to be unilateral, i.e. all feedback elements are neglected. Once such values have been calculated the optimization routine evaluates component values for specified weighted functions, e.g. gain and input v.s.w.r. using the intrinsic device S-parameters.

The output matching network is evaluated in a similar manner and is designed to provide a frequency-dependent attentuation that compensates the f.e.t. gain slope for frequencies less than 12 GHz. Part of the network used for this purpose also introduces drain bias to the transistor (Fig. 9).

When realized in lumped form the inductors and capacitors used (examples of which are given in Fig.10) have their own parasitic reactances and losses. These must be measured separately and equivalent circuits for the components postulated such that theoretical and measured S-parameters agree closely.

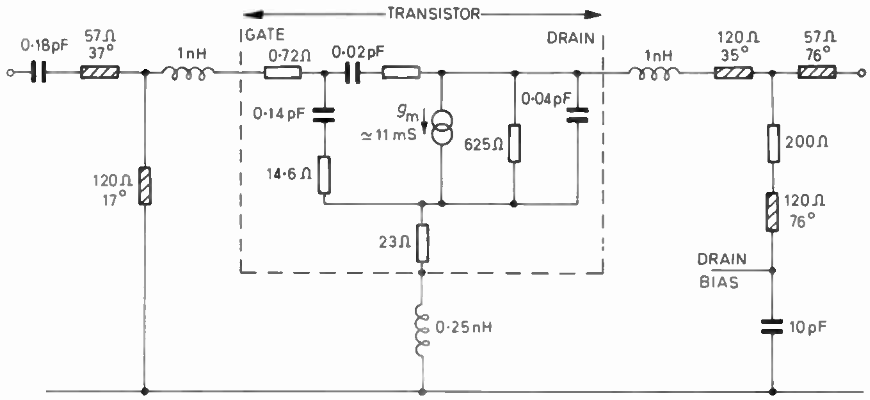


Fig. 9. Broadband matching of GAT3 device.

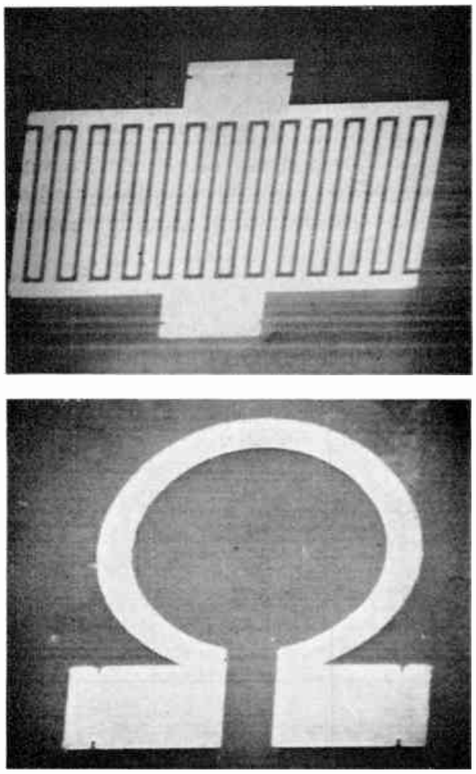


Fig. 10. Integrated capacitor and loop inductor. Approx. ×100 magnification.

Such equivalent circuits are then included in the optimization routine and a corrected circuit produced. In this way it has been found that certain circuit configurations are more useful than others in allowing the circuit designer to use the parasitics to his advantage (Fig. 11). Broadband gain can only be achieved reproducibly by optimizing the matching networks with all parasitics, discontinuities, junction effects and loss mechanisms included.

Stages other than the first, for cascading purposes, are designed for maximally flat gain with devices having high drain currents chosen for the last stage of any configuration, thus giving a maximum r.f. output power capability. The input and output matching networks are designed in a manner similar to that described but are optimized for gain and output v.s.w.r.

8 Hybrid Amplifier Performance

Figure 12 shows a two-stage amplifier constructed on 0·635 mm thick 99·7% pure alumina. The first stage is biased for minimum noise measure and the second stage for maximum gain and output power. The gain, input and output v.s.w.r.s of this amplifier are shown in Fig. 13. The power gain is 9·5 dB ± 1 dB over 6·2 to 12·4 GHz. The first stage yields a gain of 4·2 dB up to 12 GHz. This is approximately 1·3 dB less than the m.a.g. predicted under minimum noise conditions. The difference is due to loss in the lumped elements, microstrip lines,

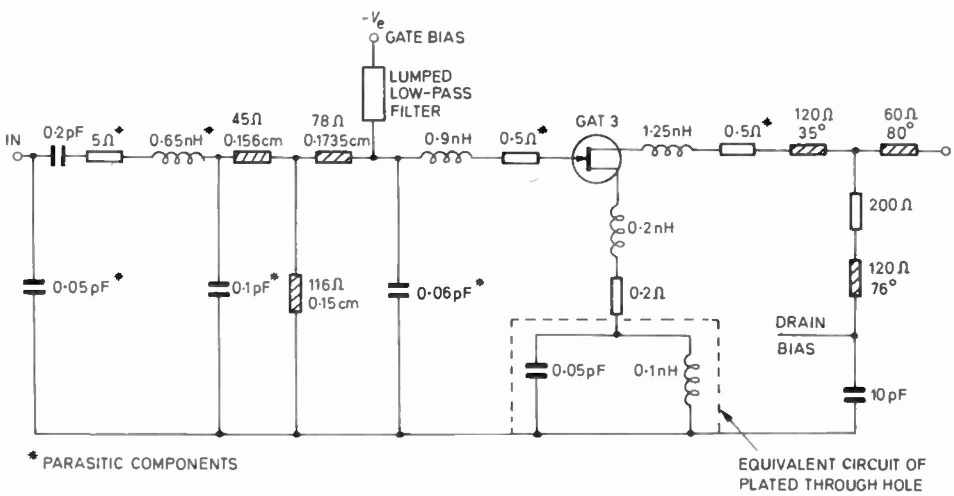


Fig. 11. A typical hybrid amplifier including parasitic elements.

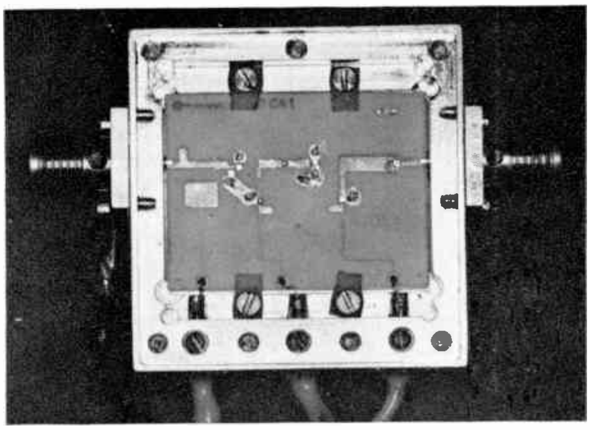


Fig. 12. Two-stage hybrid amplifier.

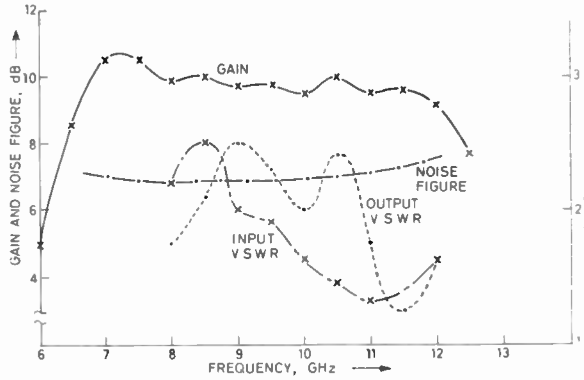


Fig. 13. Performance of 2-stage broadband amplifier.

transitions and radiation. The second stage gain is 5·5 dB up to 12 GHz. Gain variation is less than ±1 dB and is within the theoretical ±1·4 dB that can be achieved with the output matching circuits used. Further performance figures are given in Table 2.

A three-stage amplifier module with a gain in excess of 16 dB over the 6·5 to 12 GHz range has also been made. The hybrid amplifier design approach, i.e. optimizing on a modular basis, does not achieve the widest possible bandwidth but it does have the following practical advantages:

(1) Stages can be selected according to their performance (i.e. lowest noise unit selected for first stage);

(2) For maintenance purposes, in the case of failure, each stage can be easily changed; and

(3) the design and fabrication is limited to two circuits.

The assembled stages are built into a metallic enclosure which is so designed that any r.f. resonance or leakage is suppressed.

9 Monolithic Amplifier Design and Performance

Using matching concepts similar to those already described, a monolithic amplifier chip has been designed and fabricated on gallium arsenide which contains all input matching and limited output matching necessary for 8–12 GHz operation. The design has been computer optimized using techniques described later. Individual lumped inductors and capacitors, having dimensions approximately one quarter those used on the hybrid designs, have been separately measured and their equivalent circuits found.

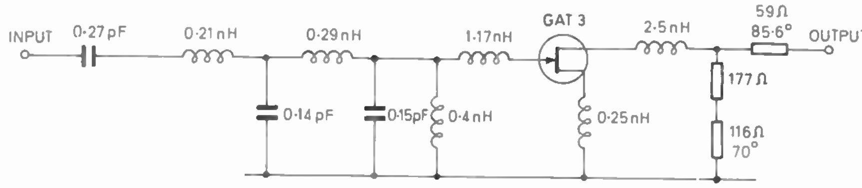
Figure 14(a) shows the basic matching circuits in lumped form neglecting the parasitic and loss components; inclusion of the latter gives the full circuit shown in Fig. 14(b). The latter network is of a complicated form and requires sophisticated computer optimization methods to realize maximum flat gain over the required bandwidth. Careful attention must be paid to the optimization routine to ensure that the computed values fall within specified limits. These limits include physical realizability in the chip sub-area allowed; lumped element modelling in discrete levels, e.g. a 0·5 pF interdigital capacitor will have different parasitic reactances from those of a 0·25 pF capacitor; and chip layout restrictions, e.g. distance between lumped elements and distance from any one lumped element to ground areas.

The first monolithic design produced on gallium arsenide is shown in Fig. 15. The fabrication uses a normal photolithographic process for the lumped matching circuits. In order to retain resolutions, particularly in the interdigital capacitors, a float-off procedure is used in fabrication. For this reason a typical metallization thickness is 1 μm. This thickness is close to one skin depth at 10 GHz and therefore a significant contribution to r.f. loss will be caused. Recent work on monolithic amplifiers, described later, has been directed to increasing the metallization thickness to approximately 3 μm. One method to achieve this is the use of d.c. sputter etching.[24]

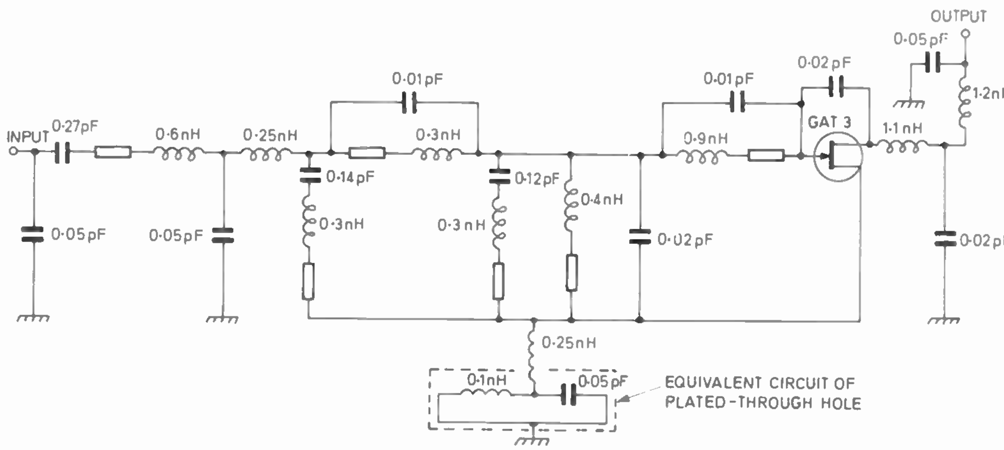
The chips were bonded onto alumina test substrates with closely controlled 25 μm diameter gold wire bonds —their inductance being part of the matching network. Gain compensation was obtained by using the output wire bond and biasing network. The amplifier produces broadband gain over 7·7–11·7 GHz. It has been found that good common grounding is needed and that layout of the components is important to minimize unwanted r.f. leakage from input to output. 10 : 1 scaled-up models (operating from 0·6 to 1·3 GHz) have been measured

Table 2. Performance of broadband 2-stage f.e.t. X-band amplifier

Frequency range, GHz	6·5–12·0
Gain, dB	9·5 ± 1·0
Reverse gain, dB	< −45
Input v.s.w.r.	< 2·5 : 1
Output v.s.w.r.	< 2·5 : 1
Noise figure, dB	< 7 at 10 GHz
1 dB gain compression, dBm	> 11 at 10 GHz
Third-order intermodulation intercept point, dBm	> 22 at 10 GHz
Dynamic range	⩾ 85 dB
Phase linearity	± 10°



(a) Monolithic amplifier equivalent circuit neglecting parasitic and loss components.



(b) Monolithic amplifier equivalent circuit including parasitic and loss components.

Fig. 14.

and analysed in order to reduce the feedback mechanisms of the first design substantially. A complete amplifier is produced on a 6·35 mm × 6·35 mm × 0·635 mm alumina substrate. Figure 16 shows the overall amplifier performance. The small lumped elements introduce r.f. losses consistent with average Q of 30–40. The repeatability with which the lumped element amplifiers can be produced has been demonstrated by performing a Monte Carlo analysis on the amplifier where the manufacturing tolerances due to variations in the photolithographic process, dielectric constant, transistor, etc., have been utilized. Such an analysis shows that a gain variation of ±0·6 dB can be expected in production.

The first monolithic integrated amplifier produced has shown that it is possible to produce broadband lumped circuits agreeing closely with theory.[23] In order to improve overall gain and bandwidth new designs are under way, at the time of writing, which also improve the cascadability of the chip. The design of a package with low parasitic reactances is also being studied. This is intended to contain one or more monolithic amplifier chips, to provide a convenient compact module, matched to 50 ohms impedance and easily assembled into microwave systems. Units of this type would be very suitable

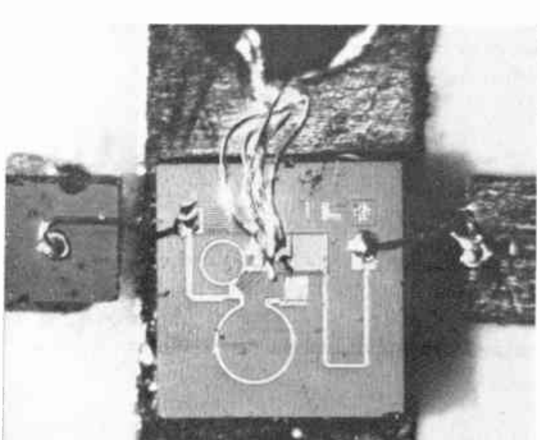


Fig. 15. Mark 1 lumped element GaAs f.e.t. amplifier.

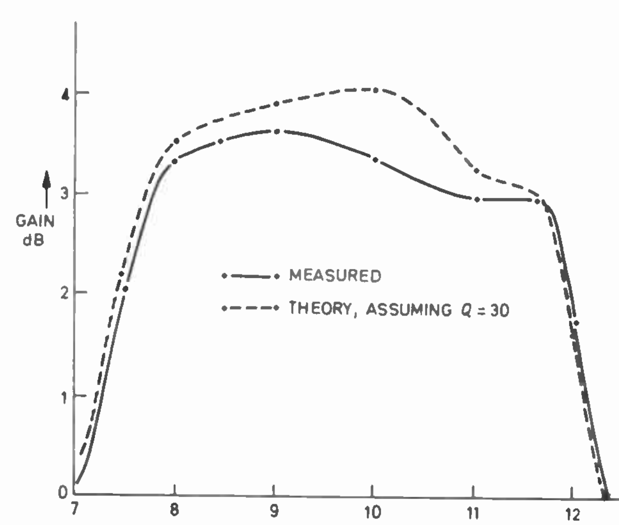


Fig. 16. Overall gain response of Mark 1 monolithic amplifier.

for applications such as panoramic receivers, phased array radars and radio relay systems.

10 Conclusions

This paper has reviewed briefly the technology and methods used in the design of small-signal microwave amplifiers. Highly consistent performance can be obtained from microwave transistors, justifying the use of rigorous methods in the characterization of devices and circuit elements and in the optimization of circuit design. For broadband applications, computer-aided design is almost essential, and designs for the higher microwave frequencies must take full account of the parasitic reactances associated with transmission line discontinuities and lumped circuit elements.

The procedures have been illustrated by examples of gallium arsenide f.e.t. amplifiers covering the range 6–12 GHz. In addition to the use of hybrid integrated circuit techniques, a unique monolithic GaAs amplifier chip is being designed which offers excellent prospects for very compact broadband amplifiers.

11 Acknowledgments

The circuits described in this paper have been designed as part of a research study sponsored by the Directorate of Components, Valves and Devices (DCVD) of the Ministry of Defence. The gallium arsenide devices have been fabricated by the Plessey Company's Allen Clark Research Centre, Caswell, Northants, and the authors are grateful to Mr. J. A. Turner of that laboratory for his assistance and advice.

This work has been carried out with the support of Procurement Executive, Ministry of Defence sponsored by DCVD and this paper is published by kind permission of DCVD and the Directors of The Plessey Company Limited.

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Manuscript first received by the Institution on 7th November 1975 and in final form on 20th January 1976. (Paper No. 1734/CC264.)

U.D.C. 621.396.969.34

Identification of complex geometrical shapes by means of low-frequency radar returns

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and

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SUMMARY

A study of the effectiveness of low-frequency electromagnetic responses for identifying objects of complex shape is presented. The linear separability of a large variety of objects such as cubes, cylinders and aircraft were examined. Two classification algorithms, a linear discriminant and a nearest neighbour rule, were used to classify a set of four aircraft models. The classification performance is presented in terms of the probability of misclassification versus noise level. The effectiveness of the various combinations of electromagnetic features was evaluated. The results indicate that amplitude, phase and polarization all contribute substantial amounts of target information. Making the assumption of *a priori* knowledge of the target's approximate aspect angle, a reliable classification can be attained utilizing a rather small number of frequencies.

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1 Introduction

Recent results in the study of the electromagnetic impulse response of finite objects[1] have indicated that the information required for the determination of the approximate shape of these objects is contained in the low frequency (Rayleigh) range, where the wavelength is longer than the overall dimensions of the object. The target's response may therefore be adequately represented by a rather small number of discrete samples in the frequency domain. Thus a set of radar returns at the specified frequencies should provide the required information for determining the approximate shape of a target. These returns can therefore constitute a set of features which may be used in an appropriate classification algorithm. Several algorithms were indeed tried to classify various simply shaped objects with considerable success.[2] In fact, it was demonstrated that of the 12 frequencies used to represent the target's response the contributions of the first few were dominant in reducing misclassification errors. The implications of this fact are very significant with respect to the practical implementability of the approach, since the complexity and cost of the system increase with the number of frequencies used. As will be shown below four frequencies are sufficient in most cases to provide reliable classification in the presence of substantial amounts of noise. A further reduction in the number of frequencies is possible by the introduction of polarization diversity.

The results of previous work[2] demonstrated that objects which differ substantially from each other in geometrical characteristics such as a cube, a sphere and a cylinder can be discriminated on the basis of their radar returns. The questions that may then be logically raised are, how large do these differences have to be in order that the objects be distinguishable and, on the other hand, will even small variations in shape cause the objects to be classified as different. The importance of the first question is obvious, the importance of the second question stems from the requirement that various objects of a similar approximate shape be recognized as having the same basic shape although they may differ in minor detail. For example, a cylindrical rocket must be recognized as such even though its nose may be of a different shape than expected, or the fins might be of different design. In more complex shapes, e.g., aircraft, they must be correctly identified by type regardless of whether they carry all their rockets or not.

The tests carried out have, indeed, shown that small object perturbations produce correspondingly small changes in the feature space, permitting the classification of objects in accordance with their approximate geometrical characteristics such as 'cylindrical shapes'. Also the types of object deformations were found which would indeed cause the object to be classified as, say a 'non-cylindrical object'. Thus a fairly good insight was gained into the sensitivity of the classifier to the geometrical variations of targets.

Since relatively simple geometrical shapes were successfully classified, the attempt was made to classify objects of substantially greater complexity and practical interest, namely, aircraft. A set of four aircraft was chosen, *F104*,

The four models, to normalized dimensions, used in the experiments.
KEY (with actual dimensions)

F104 (18·42 m fuselage, 16·60 m wingspan)
F4 (19·86 m fuselage, 11·70 m wingspan)
Mig 19 (13·49 m fuselage, 11·12 m wingspan)
Mig 21 (15·76 m fuselage, 7·15 m wingspan)

F4, *Mig 19* and *Mig 21*. Because of the complexity of these objects the backscattered returns had to be obtained for a large number of viewing, or aspect, angles. The classification was attempted initially for a limited range of aspect angles and was quite successful even in the presence of moderate amounts of noise. However as the range of viewing angles was expanded to cover a major portion of the totality of angles, classification failed. The classifier used to that point was a linear discriminant, as described in Section 2. It became apparent that the feature space occupancies of different aircraft were certainly not linearly separable but, encouragingly, actual overlap did not occur. A 'nearest neighbour' approach was then attempted and resulted in successful classification, as described in Section 3. The nearest neighbour rule is indeed superior to a linear classifier, however, it is somewhat more complicated to implement and considerably more difficult to evalu te. The performance, i.e., misclassification probability as a function of noise level, may be determined analytically for the linear classifier. Statistical techniques are required, however, to evaluate the errors for the nearest neighbour rule. Section 3 presents the performance of the nearest neighbour classifier and compares it to the linear classifier

The signal scattered by a target contains both phase and amplitude information representing the target's electromagnetic response. The results discussed above utilized amplitude information alone, the same was true for the work described in Reference 2. The reason is quite obvious, amplitudes are easier to obtain and process and are less sensitive to errors. However, one can obtain the phase information with a moderate amount of processing: and coherent radars providing the required information are well within the state of the art. It was thus of interest to study the improvement afforded by the phase information. The results, as described in Section 4, indicate a rather spectacular reduction in misclassification probabilities. The reason for that was the substantial increase in the distance between classes in the feature space. This increased separation produced in turn a far more than proportional decrease in error probabilities since the error probabilities are non-linearly related to the interclass distances.

In a further effort to utilize the full potential of the electromagnetic signal return the polarization parameter was introduced. The joint use of vertical and horizontal polarizations indeed resulted in substantial error reduction as described in Section 5.

2 Linear Discriminant Procedure

One of the simplest and yet most effective classification techniques is the linear discriminant procedure. It is relatively simple to apply, to implement and also to evaluate its performance. Because of these characteristics it was chosen to test the separability of the various classes of objects studied. Repjar *et al.*[2] modified the classical linear discriminant methods to provide for minimization of error probabilities in the presence of noise. This modified approach was used in the present study. The linear discriminant procedure for the utilization of the discrete frequency radar returns may be described as follows:

An n dimensional vector, $\bar{a}_i^k$, is formed whose jth component, a_{ij}^k, is defined as the amplitude return obtained at the jth frequency for the ith aspect angle of the kth object. As the aspect angle changes $\{\bar{a}_i^k\}$ forms a set of vectors, or a cluster of points, in Euclidean n-space representing a given object or class. Two differently shaped objects will be represented by two clusters in n-space. A straightforward method of classification is available when there exists a hyperplane separating these two classes, i.e. when the classes are linearly separable. The optimum treatment of this case is to project the two clusters orthogonally onto a line so that the mean distance between the two projected clusters is as large as possible relative to the spread within the clusters. Repjar[2] investigated the linear separability and classification of various simple geometrical shapes. All objects were successfully classified. This Section extends the use of the above linear discriminant method of classification to more complicated objects including aircraft. The effects of shape perturbations on the classification process are discussed and the separability and misclassification probability curves for different objects are presented.

2.1 Linear Separation of Simple Objects

A series of linear separability tests was made of cylindrical objects with various perturbations and contrasted with the separability from a basically different shape, that of a cube. The objects, c^i, which were chosen are—

- c^1: a 2 : 1 (3·81 cm × 1·905 cm) circular cylinder.
- c^2: above circular cylinder with a 2 mm dipole.
- c^3: above circular cylinder with an 8 mm dipole.
- c^4: above circular cylinder with 5 mm fins.
- c^5: above circular cylinder with a 60° conical cap.
- c^6: a stepped cylinder consisting of 2 cylinders, above 2 : 1 circular cylinder with a smaller 1·27 cm × 1·29 cm cylinder.
- c^7: above circular cylinder with a spherical cap.
- c^8: above circular cylinder with a 1·27 cm hole on one end.
- c^9: above circular cylinder with a 1·59 cm hole on one end.
- c^{10}: a cube of 2·1 cm on the side.

It was found that the cube whose shape is radically different from the cylindrical shapes was indeed linearly separable from the whole group of cylindrical objects (c^1 to c^9). The average probabilities of misclassification are plotted as a function of additive, zero mean Gaussian noise of variance σ^2 (σ in cm units) for the cube against each of the cylindrical objects and against the entire group of cylindrical objects and are shown in Fig. 1. The linear separation indicates that the low-frequency scattering data from different basic simple shapes occupy non-overlapping regions in feature space. The error probability is largest when the entire group of 9 objects is considered as a single class since its feature space occupancy is much larger and a hyperplane separating this group from the cube would be more constrained and thus closer to the cube feature points. To determine the distinguishability between basically similar objects with small geometrical variations, a series of linear separability tests was performed on a set of cylindrical shapes. The cylindrical shape was chosen since it could represent a prototype for both missiles and many satellites, and would relate to some extent to the fuselage of an aircraft. The main objective in examining these particular shapes was to determine the sensitivity of the classification process to varying amounts of object shape perturbation.

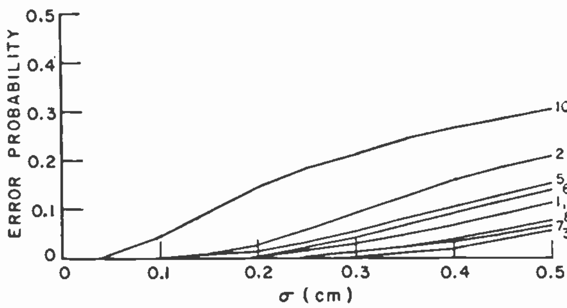

Fig. 1. Average probability of misclassification for a cube (c^{10}) vs. the cylindrical objects (c^1 to c^9).

1. c^{10} vs. c^1
2. c^{10} vs. c^2
3. c^{10} vs. c^3
4. c^{10} vs. c^4
5. c^{10} vs. c^5
6. c^{10} vs. c^6
7. c^{10} vs. c^7
8. c^{10} vs. c^8
9. c^{10} vs. c^9
10. c^{10} vs. $\bigcup_{i=1}^{9} c^i$

The perturbations were chosen to represent realistic modifications of the basic shape that would induce a substantial electromagnetic response change. The small deformation of the cylindrical shape such as, say, flat sections replacing the cylindrical curvature were not tested since they would not produce significant electromagnetic variations and consequently would not affect the scattered return. On the other hand the introduction of dipoles representing antennas or other substantial protrusions would evoke significant electrical response changes and would therefore indicate the level of interference or perturbation that the target signature may sustain without introducing substantial error in classification.

The objects tested were c^1 to c^{10}. The test results show that the objects are pairwise linearly separable for all pairs except for the cases of c^1 vs. c^9 and c^7 vs. c^9. Error probability curves as a function of noise variance σ^2 (σ in cm units) for a cylinder against each of the other objects are plotted in Fig. 2. We see that the probability of error for the cylinder vs. cylinder with large dipole is the lowest one; and the error for a cylinder vs. a cylinder with a small hole or a small dipole are the highest among all pairs. It is evident that the shape difference between a cylinder and a cylinder with a small hole or a small dipole are much smaller than the difference between a cylinder

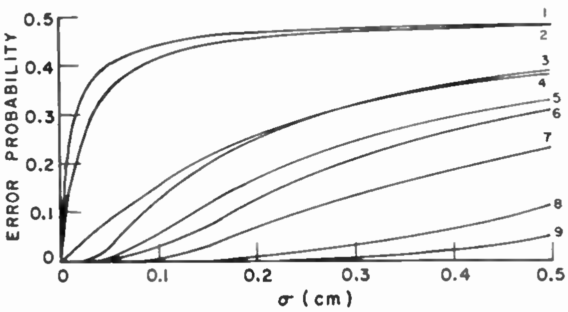

Fig. 2. Average probability of misclassification for Class 1 vs. Class 2 to Class 10.

1. c^1 vs. c^8
2. c^1 vs. c^2
3. c^1 vs. c^4
4. c^1 vs. c^9
5. c^1 vs. c^7
6. c^1 vs. c^6
7. c^1 vs. c^5
8. c^1 vs. c^{10}
9. c^1 vs. c^3

and a cylinder with a large dipole, which would be expected. The fact that some of the shapes are not separable at all indicates the strong electromagnetic similarity between these objects. It is thus apparent that the small object perturbations do not significantly affect the target response and thus do not permit separability, or if they do, only marginally so. These results indicate that the classifier will recognize basic shapes and will classify them as such although their precise shapes may vary somewhat.

To demonstrate this property the projection vector $\bar{\omega}$ derived for the best separation of a simple cylinder from a cylinder with a large dipole (representing a non-cylinder) was used to classify cylinders with various perturbations, classes c^2, c^8 and c^9. The error probability curves shown in Fig. 3 indicate clearly that the classifier recognizes the shapes as cylinders in spite of the perturbations. Indeed a comparison with the optimum separation of c^2 from c^3 is not significantly better than the one using the basic cylinder separating vector $\bar{\omega}$. Also, as already mentioned above the cube is readily separable from the whole cylindrical class.

2.2 Multi-class Separation

In this Section we will describe the use of pairwise classification for the identification of one out of a set of K possible classes. The method simply involves the combinations of $K-1$ classes into one class and separation from the single remaining class.

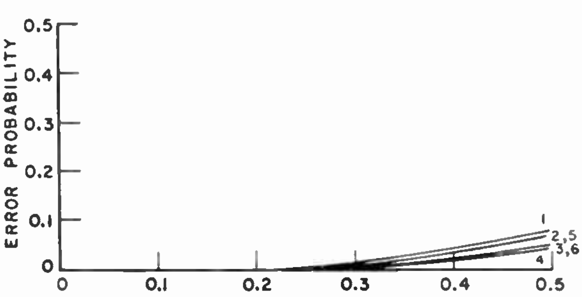

Fig. 3. Average probability of misclassification.

1. c^2 vs. c^3 using $\bar{\omega}$ of c^1 vs. c^3
2. c^8 vs. c^3 using $\bar{\omega}$ of c^1 vs. c^3
3. c^9 vs. c^3 using $\bar{\omega}$ of c^1 vs. c^3
4. c^2 vs. c^3
5. c^8 vs. c^3
6. c^9 vs. c^3

The following simply shaped objects were tested for multi-class separation.

c^{10}: a cube of 2·1 cm on the side.

c^{11}: a hemispherical boss of diameter 2·385 cm.

c^{12}: a 60° cone with a base of 4·3 cm in diameter.

c^{13}: a 1·905 cm × 1·905 cm circular cylinder.

c^{14}: a sphere of 3 cm diameter.

c^{15}: a sphere of 1·8 cm diameter.

c^{16}: a square plate of 2·1 cm on the side.

c^{17}: a dielectric sphere of 2 cm diameter with $\varepsilon_r = 2{\cdot}208$.

c^{18}: a thick wire of 0·015 cm in diameter and 3 cm in length.

For each class c^i, $i = 10, \ldots, 18$, a hyperplane was found by combining eight classes $U_{j \neq i}\, c^j$ into one class and separating from c^i. It was found that these 9 classes were linearly separable using this technique. The average probability of misclassification for each case is plotted in Fig. 4. The following conclusions may be drawn from the relative error levels of the various object groupings. The highest misclassification error was experienced by c^{14}, namely the 3 cm conducting sphere. This object has the closest electromagnetic similarity to the majority of the objects, most of which have major dimensions of approximately 3 cm. Thus the location of the n-tuple representing this sphere must be located near the centre of the n space occupied by those corresponding to other objects and very close to at least some of the n-tuples belonging to other classes. On the other hand c^{12}, a 60° cone, has dimensions and shape sufficiently different from the rest of the group so that it is more easily separable and leads to a low misclassification level.

The error curves corresponding to other groupings portray the level of commonality or similarity between a given object and the rest of the group. It is thus evident that the pairwise classification approach would be useful for multi-class classification if the objects were sufficiently

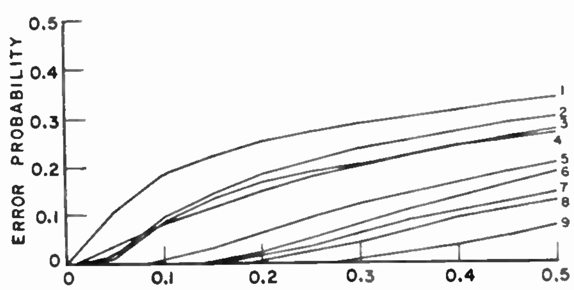

Fig. 4. Average probability of misclassification by the linear discriminant method where 8 classes are combined together into one class and separated from remaining class.

1. c^{14} vs. all others
2. c^{16} vs. all others
3. c^{10} vs. all others
4. c^{18} vs. all others
5. c^{13} vs. all others
6. c^{11} vs. all others
7. c^{17} vs. all others
8. c^{15} vs. all others
9. c^{12} vs. all others

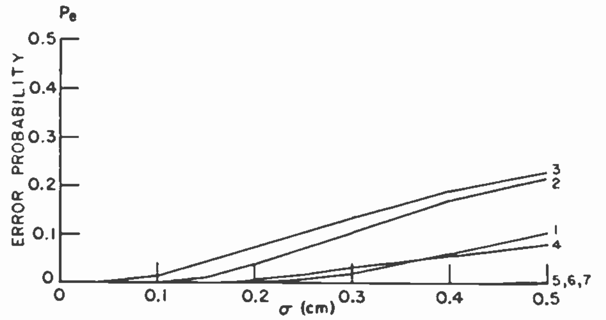

Fig. 5. Average probability of misclassification of various roll angles for *F104* vs. *Mig 19*. Vertical polarization. Ten frequencies.

1. level flight 0–360° azimuth
2. 15° tilt 0–360° azimuth
3. 27° tilt 0–360° azimuth
4. 45° tilt 0–360° azimuth
5. 60° tilt 0–360° azimuth
6. 75° tilt 0–360° azimuth
7. 90° tilt 0–360° azimuth

dissimilar. As a further test a prolate spheroid 4·77 cm by 2·38 cm was introduced and the same multi-class separation tested. Note that its overall dimensions are similar to those of the sphere and its volume is almost identical to that of the sphere. It would thus be located in the same region of n-space as the sphere. Since it is not, however, spherically symmetric, its representative n-tuples could easily be interspersed with most of the other objects. Indeed the results showed the spheroid is linearly inseparable from the rest of the group although when combined with other objects it is separable from one object at a time. Under these conditions the error level was not substantially worse than that of the sphere.

2.3 Aircraft Classification Utilizing Linear Discriminants

In view of the successful classification of relatively simply shaped objects, an attempt was made to classify substantially more complex objects such as aircraft. Because of the basic similarity of the objects to be classified, they all have the same structure—fuselage, wings, tail, etc.—it was expected that the feature distributions representing different aircraft will overlap at least in some portions of the n-dimensional space. It was hoped however that differences in shape would be sufficient to permit separability in some regions. A natural subdivision of regions for which linear separation could be attempted was provided by different observation regions, or aspect angle ranges. The aspect angle ranges chosen were 0–360° in azimuth, and a set of discrete tilt or roll angles. Specifically the roll angles chosen were 0°, 15°, 27°, 45°, 60°, 75°, and 90°. Two aircraft, an *F104* and a *Mig 19*, were tested. The measurements were carried out on scaled models, with the frequencies chosen to maintain the correct ratio of aircraft dimensions to wavelength. The frequencies used were 1·08 GHz and its multiples. The airplane model maximum dimensions were approximately 8 cm for both models. The true size of the aircraft differed, however, substantially.† Thus the classification utilizing measurements of the equal size model provide an upper bound on the error probabilities. Scattering data representing the true dimension of the objects can be expected to lead to significantly lower classification errors since the overall size differences will provide added feature contrast.

† The need for experimental data necessitated the use of models, and the above models were the only ones available at the required scale. (See photograph on p. 473.)

The classification tests carried out on these planes have shown that they are indeed linearly separable for every roll angle tested as mentioned above. The average probability of misclassification for each range of observation angles is presented in Fig. 5.

Unfortunately, classification by means of linear discriminants runs into increasing difficulties as more look-angles are considered at a time. A thorough investigation of the distribution of the n-tuples in the feature space[3] revealed that the classes are distributed on fairly convolved non-planar surfaces which cannot be separated by hyperplanes except for local regions. A more effective classification method for such situations is the 'nearest neighbour' rule described in the next Section.

3 Nearest Neighbour Decision Rule

The nearest neighbour (n.n.) decision rule can be described as follows. Given training samples, $\{\bar{a}_i^k\}$, from each of several (k) classes, the rule is to classify an unknown sample, $\bar{y}$, as a member of the class, C^r, to which its n.n. belongs, i.e.,

$$\bar{y} \in C^r \quad \text{if} \quad d(\bar{y}, \bar{a}^r) = \min_{i,k} \|\bar{y} - \bar{a}_i^k\| \tag{1}$$

where

$$d(\bar{y}, \bar{a}^k) = \|\bar{y} - \bar{a}^k\| = [(\bar{y} - \bar{a}^k)^T(\bar{y} - \bar{a}^k)]^{1/2}$$
$$k = 1, 2, \ldots, M$$
$$i = 1, 2, \ldots, N_k$$
$$M = \text{number of classes}$$
$$N_k = \text{number of samples in class } k.$$

Cover and Hart[4] showed that in the large sample case the probability of error of the n.n. rule is bounded from below by the Bayes probability of error and from above by twice the Bayes error. These bounds are sufficient for indicating the merits of the n.n. rule as a classifier. An analytical calculation of the exact probability of error for the n.n. rule is rather involved because it requires the integration of a multivariate density function over extremely complicated boundaries. For this reason, statistical computation was resorted to for obtaining the probability of misclassification.

In the estimation of the probability of misclassification, the Monte Carlo method was used with training data generated by computation.[5] As in the linear discriminant procedure, we assumed all scattering data as equally probable. The test data are obtained by adding noise to the training data. Each training data point is contaminated additively by multivariate Gaussian noise with zero mean and equal variance, σ^2, (normalized to the signal level), and classified according to the n.n. decision rule. The normalization of the noise standard deviation to signal return level is aimed at providing a meaningful measure of the effect of both system noise and measurement errors on classification performance. Thus $\sigma = 0{\cdot}1$ corresponds to a noise level of approximately 10% of the signal level. The repetition of this process through all the data gives the cumulative count of hit and miss. The estimated probability of misclassification is simply the ratio of the number of test samples misclassified to the total number of test samples in the experiment. For each given σ, two thousand test samples were tested to obtain an estimate of average probability of misclassification with an accuracy of two significant digits. The parameter σ was varied between 0·1 to 0·5.

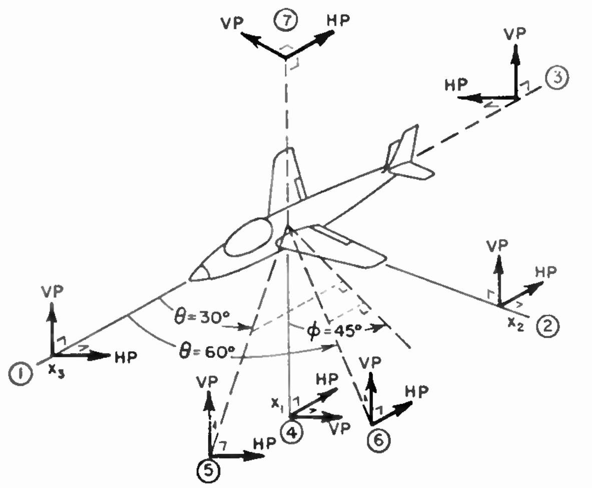

Fig. 6. The coordinate system of the aspect angles chosen for experiments.

1.	$\theta = 0°$	$\phi = 90°$	5.	$\theta = 30°$	$\phi = 45°$
2.	$\theta = 90°$	$\phi = 90°$	6.	$\theta = 60°$	$\phi = 45°$
3.	$\theta = 180°$	$\phi = 90°$	7.	$\theta = 90°$	$\phi = 180°$
4.	$\theta = 90°$	$\phi = 0°$			

3.1 Classification of Aircraft by Nearest Neighbour Decision Rule

Computationally-obtained scattering data[5] for the *F104*, *Mig 19*, *F4* and *Mig 21* aircraft provided the data base from which the numerous results described below were obtained. The observation angles referred to in the results presented are described in Fig. 6. The bulk of the results present the probabilities of misclassification occurring around specified observation angles. The angles were chosen to represent the various aspect angles at which an aircraft would be observed either by a ground station or from the air. Since the approximate observation angle would usually be obtained from radar tracking data, it is reasonable to assume that the aspect angle is known to within a few degrees. We can therefore consider only the sample vectors from a region around the given aspect angle as the training set. The classification performance using the n.n. decision rule for seven different look angles as indicated in Fig. 6 is described below. The error probabilities are polarization dependent and consequently the performance corresponding to the two polarizations are presented separately.

The vertically polarized scattering data used involved the target illumination by an electric vector oriented along the unit vector x_1. Since the aircraft orientation with respect to the illuminating radar beam was varied, the polarization of the wave incident on the aircraft also

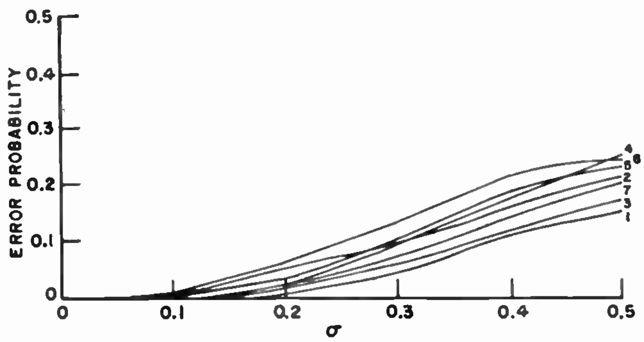

Fig. 7. Average probability of misclassification for *F104* vs. *Mig 19* by n.n. classifier.

Computed 4 frequencies, v.p. amplitude returns

1.	$\theta = 0°$	$\phi = 90°$	nose on	12 pts	$\bar{A} = 0{\cdot}9$
2.	$\theta = 90°$	$\phi = 90°$	right wing	18 pts	$\bar{A} = 1{\cdot}78$
3.	$\theta = 180°$	$\phi = 90°$	tail on	12 pts	$\bar{A} = 1{\cdot}0$
4.	$\theta = 90°$	$\phi = 90°$	bottom	12 pts	$\bar{A} = 6{\cdot}85$
5.	$\theta = 30°$	$\phi = 45°$		18 pts	$\bar{A} = 2.67$
6.	$\theta = 60°$	$\phi = 45°$		18 pts	$\bar{A} = 2{\cdot}88$
7.	$\theta = 90°$	$\phi = 180°$	top	12 pts	$\bar{A} = 5{\cdot}85$

varied. But in the majority of the situations considered in this section the incident polarization had a dominant component normal to the major aircraft components, i.e. the fuselage and the wings. This tended to produce lower return signals than in the horizontal illumination case but it was found that the vertically polarized signals had significantly better discriminating properties than the horizontal ones.

The error probability was computed for the pairwise classification between all possible pairs formed from the four aircraft, *F104*, *Mig 19*, *F4* and *Mig 21*. Typical error probability curves for different aspect angles are shown in Fig. 7 for *F104* vs. *Mig 19*. A grid of eight points separated by 10° in (θ, ϕ) space was formed around the point in (θ, ϕ) which was obtained as the estimate of the orientation of the unknown aircraft from the radar tracking data. Thus there were 18 distinct points representing the two alternative classes. In the case of nose, tail, bottom and top observation angles only 12 independent scattering sets of data were obtained for the 18 observation points because of the object symmetry. The average (r.m.s.) value of all signal return data points in the given region over all frequencies and aircraft is denoted by A, it is used to normalize the standard deviation of the noise added to the scattering return to compute misclassification probabilities. The normalization to average signal level is used to provide a realistic measure of performance which takes into account the level of transmitter power, range to target, antenna gains and clutter. In fact, it takes into account multiplicative errors as well since the injected noise is proportional to the signal level.

Figure 8 shows the typical error curves for all six pairwise classifications at $(\theta, \phi) = (30°, 45°)$. It appears that the *Mig 19* is most dissimilar from the *Mig 21* and thus is least susceptible to misclassification due to noise. The dissimilarity may be due to the difference in the wing shapes of the two aircraft—the small delta wing of the *Mig 21* compared to the big swept wing of the *Mig 19*.

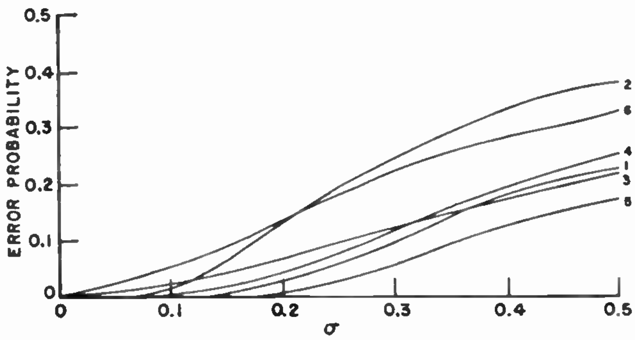

Fig. 8. Average probability of misclassification by n.n. classifier. Computed 4 frequencies, v.p. amplitude returns,

$\theta = 30°$, $\phi = 45°$, 18 pts.

1. *F104* vs. *Mig 19* $\bar{A} = 2{\cdot}67$
2. *F104* vs. *F4* $\bar{A} = 2{\cdot}73$
3. *F104* vs. *Mig 21* $\bar{A} = 2{\cdot}31$
4. *Mig 19* vs. *F4* $\bar{A} = 2{\cdot}89$
5. *Mig 19* vs. *Mig 21* $\bar{A} = 2{\cdot}5$
6. *F4* vs. *Mig. 21* $\bar{A} = 2{\cdot}56$

The *F104* and *F4* on the other hand are the most proximate pair. It should be recalled that all four aircraft models are scaled to about the same maximum size. Thus we are actually studying the difficult problem of classifying complex objects based on minor shape variations. In real situations different aircraft will be of different sizes and produce a larger interclass spread, leading to better classification performance.

The next problem considered is multiclass identification. One could utilize the pairwise classification results and compute the misclassification error for a sequence of six pairwise tests, in which the most likely class is selected. It is preferable to consider, however, the simultaneous classification of one of the four classes, with all four classes being simultaneously present in the feature space. Some typical results of these tests are shown in Figs. 9 and 10. The number of points participating in each test is twice as large as in the previous test since four classes rather than two are represented. As can be seen from Fig. 10, the highest error rate occurs for *F4* vs. *Mig 21*, *F104* and *Mig 19*. The apparent reason is due to *F4*'s larger surface area and shape complexity which cause its scattering parameters to range more widely than those of the other aircraft. If its relative size with respect to the other aircraft were taken into consideration, as would be the case in an actual situation, the *F4* group of points would have been removed from those of the other aircraft with a substantial reduction of error probabilities.

Horizontal polarization implies that for the majority of cases considered the illumination is either completely parallel to the fuselage or wings or that its dominant component is so oriented. It is apparent, then, that the signal returns will be significantly larger than for the vertical polarization case, but since the overall dimensions of the aircraft have been normalized, these returns could be expected not to differ substantially for different aircraft. This indeed is the case, a study of the distribution of the points in the frequency or feature space having shown that they tend to cluster quite closely to each other, which is reflected in the error probabilities described by Figs. 11 and 12 for pairwise classification and Figs. 13 and 14 for multiclass classification. Here again, if the true relative dimensions of those classes were included the errors would have been substantially reduced, but if aircraft of truly equal overall dimensions were considered the above figures do represent a realistic estimate of performance.

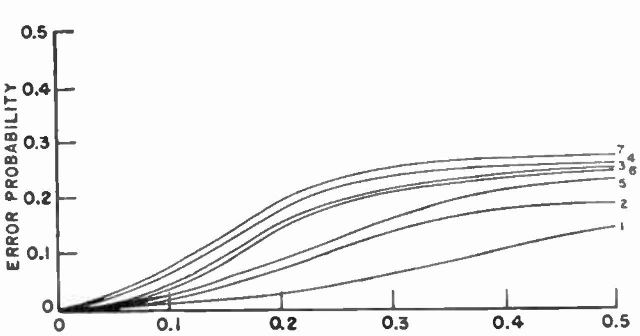

Fig. 9. Average probability of misclassification for *F104* vs. *Mig 19*, *F4* and *Mig 21* by n.n. classifier.
Computed 4 frequencies, v.p. amplitude returns.

1.	$\theta = 0°$	$\phi = 90°$	nose on	24 pts	$\bar{A} = 1{\cdot}37$
2.	$\theta = 90°$	$\phi = 90°$	left wing	36 pts	$\bar{A} = 2{\cdot}3$
3.	$\theta = 180°$	$\phi = 90°$	tail on	24 pts	$\bar{A} = 1{\cdot}38$
4.	$\theta = 90°$	$\phi = 0°$	bottom	24 pts	$\bar{A} = 6{\cdot}55$
5.	$\theta = 30°$	$\phi = 45°$		36 pts	$\bar{A} = 2{\cdot}62$
6.	$\theta = 60°$	$\phi = 45°$		36 pts	$\bar{A} = 3{\cdot}1$
7.	$\theta = 90°$	$\phi = 180°$	top	24 pts	$\bar{A} = 5{\cdot}79$

3.2 Performance Comparison of Linear Discriminants vs. Nearest Neighbour Procedure

To compare the performance between the linear discriminant to that of the n.n. procedures, the same vertical polarization data used in the previous section are used to test a linear discriminant procedure. The probability of misclassification for all pairwise and multiclass cases at $(\theta, \phi) = (30°, 45°)$ are presented in Figs. 15 and 16. The results indicate that linear separation is not possible in a few cases, especially in the multiclass situation. The comparison of Figs. 15 and 16 to Figs. 8 and 10 clearly demonstrates that the n.n. rule is superior to the linear discriminant procedure. This is

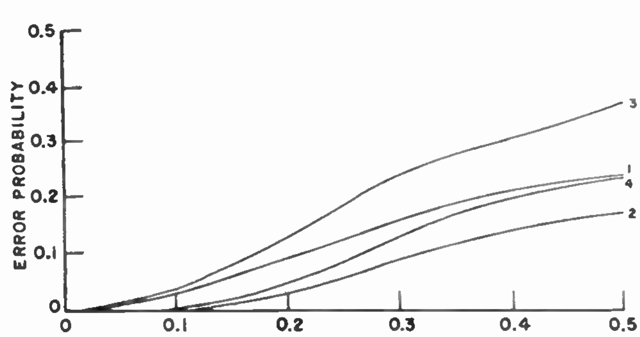

Fig. 10. Average probability of misclassification by n.n. classifier. Computed 4 frequencies, v.p. amplitude returns,
$\theta = 30°$, $\phi = 45°$, 36 pts, $\bar{A} = 2{\cdot}62$.

1. *F104* vs. *Mig 19*, *F4* and *Mig 21*
2. *Mig 19* vs. *F4*, *Mig 21* and *F104*
3. *F4* vs. *Mig 21*, *F104* and *Mig 19*
4. *Mig 21* vs. *F104*, *Mig 19* and *F4*

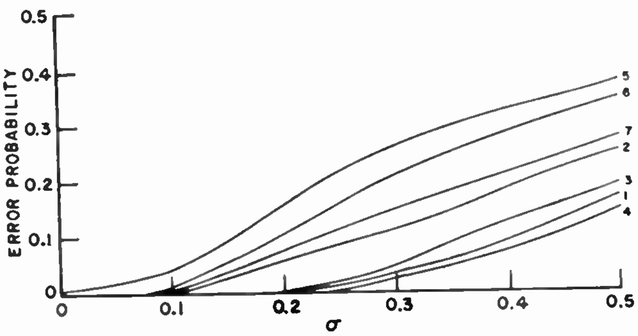

Fig. 11. Average probability of misclassification for *F104* vs. *Mig 19* by n.n. classifier.
Computed 4 frequencies, h.p. amplitude returns.

1.	$\theta = 0°$	$\phi = 90°$	nose	12 pts	$\bar{A} = 1{\cdot}37$
2.	$\theta = 90°$	$\phi = 90°$	left wing	18 pts	$\bar{A} = 5{\cdot}79$
3.	$\theta = 180°$	$\phi = 90°$	tail	12 pts	$\bar{A} = 6{\cdot}18$
4.	$\theta = 90°$	$\phi = 0°$	bottom	12 pts	$\bar{A} = 5{\cdot}89$
5.	$\theta = 30°$	$\phi = 45°$		18 pts	$\bar{A} = 2{\cdot}74$
6.	$\theta = 60°$	$\phi = 45°$		18 pts	$\bar{A} = 4{\cdot}4$
7.	$\theta = 90°$	$\phi = 180°$	top	12 pts	$\bar{A} = 6{\cdot}25$

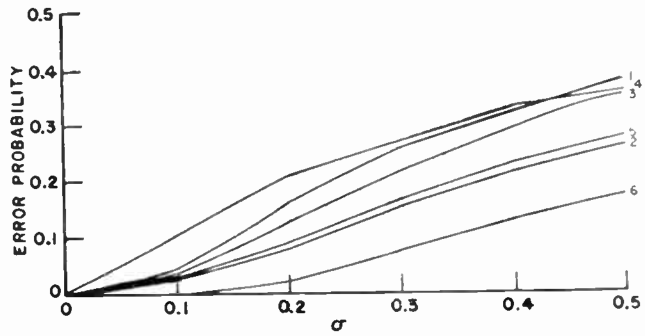

Fig. 12. Average probability of misclassification by n.n. classifier. Computed 4 frequencies, h.p. amplitude returns.
$\theta = 30°$, $\phi = 45°$, 18 pts.

1. *F104* vs. *Mig 19* $\bar{A} = 2{\cdot}74$
2. *F104* vs. *F4* $\bar{A} = 3{\cdot}21$
3. *F104* vs. *Mig 21* $\bar{A} = 2{\cdot}6$
4. *Mig 19* vs. *F4* $\bar{A} = 3{\cdot}14$
5. *Mig 19* vs. *Mig 21* $\bar{A} = 2{\cdot}52$
6. *F4* vs. *Mig 21* $\bar{A} = 3{\cdot}17$

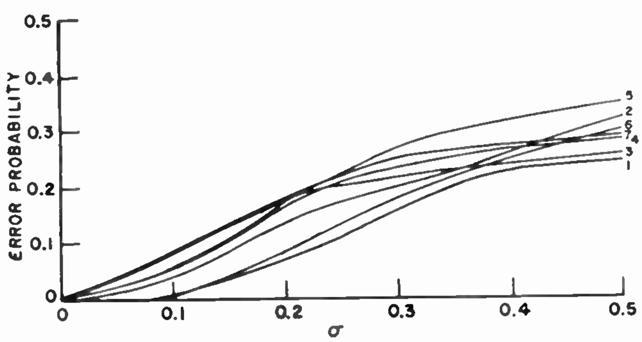

Fig. 13. Average probability of misclassification for *F104* vs. *Mig 19*, *F4* and *Mig 21* by n.n. classifier.
Computed 4 frequencies, h.p. amplitude returns.

1.	$\theta = 0°$	$\phi = 0°$	nose on	12 pts	$\bar{A} = 4{\cdot}95$
2.	$\theta = 90°$	$\phi = 90°$	left wing	18 pts	$\bar{A} = 5{\cdot}82$
3.	$\theta = 180°$	$\phi = 0°$	tail on	12 pts	$\bar{A} = 6{\cdot}01$
4.	$\theta = 90°$	$\phi = 0°$	bottom	12 pts	$\bar{A} = 6{\cdot}36$
5.	$\theta = 30°$	$\phi = 45°$		18 pts	$\bar{A} = 3{\cdot}14$
6.	$\theta = 60°$	$\phi = 45°$		18 pts	$\bar{A} = 4{\cdot}13$
7.	$\theta = 90°$	$\phi = 180°$	top	12 pts	$\bar{A} = 7{\cdot}1$

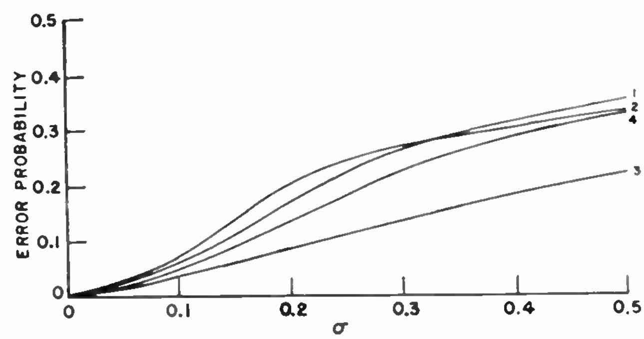

Fig. 14. Average probability of misclassification by n.n. classifier Computed 4 frequencies, h.p. amplitude returns,
$\theta = 30°$, $\phi = 45°$, 36 pts, $\bar{A} = 4{\cdot}31$.

1. *F104* vs. *Mig 19*, *F4* and *Mig 21*
2. *Mig 19* vs. *F4*, *Mig 21* and *F104*
3. *F4* vs. *Mig 21*, *F104* and *Mig 19*
4. *Mig 21* vs. *F104*, *Mig 19* and *F4*

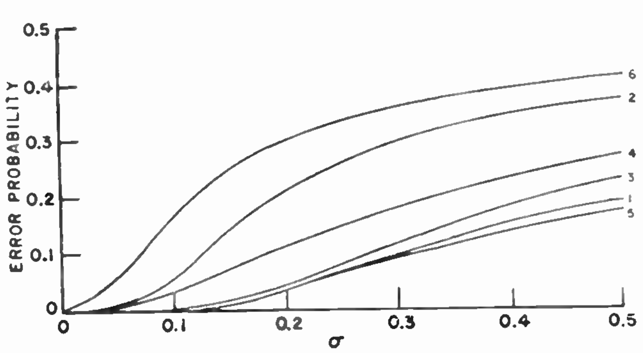

Fig. 15. Average probability of misclassification by linear classifier. Computed 4 frequencies, v.p. amplitude returns,
$\theta = 30°$, $\phi = 45°$, 18 pts.

1. *F104* vs. *Mig 19* $\bar{A} = 2{\cdot}67$
2. *F104* vs. *F4* $\bar{A} = 2{\cdot}73$
3. *F104* vs. *Mig 21* $\bar{A} = 2{\cdot}31$
4. *Mig 19* vs. *F4* $\bar{A} = 3{\cdot}09$
5. *Mig 19* vs. *Mig 21* $\bar{A} = 2{\cdot}5$
6. *F4* vs. *Mig 21* $\bar{A} = 2{\cdot}56$

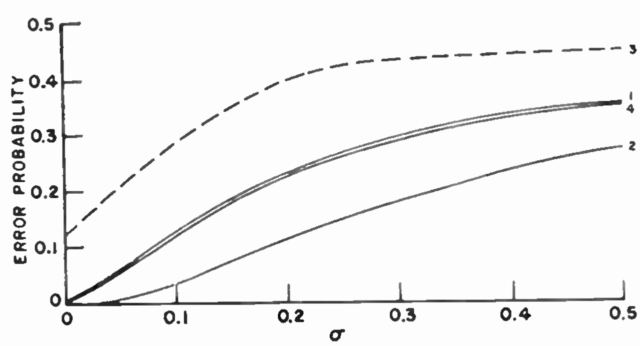

Fig. 16. Average probability of misclassification by linear classifier. Computed 4 frequencies, v.p. amplitude returns,
$\theta = 30°$, $\phi = 45°$, 36 pts, $\bar{A} = 2{\cdot}62$.

1. *F104* vs. *Mig 19*, *F4* and *Mig 21*
2. *Mig 19* vs. *F4*, *Mig 21* and *F104*
3. *F4* vs. *Mig 21*, *F104* and *Mig 21*
4. *Mig 21* vs. *F104*, *Mig 19* and *F4*

Note: Dashed line indicates not linearly separable.

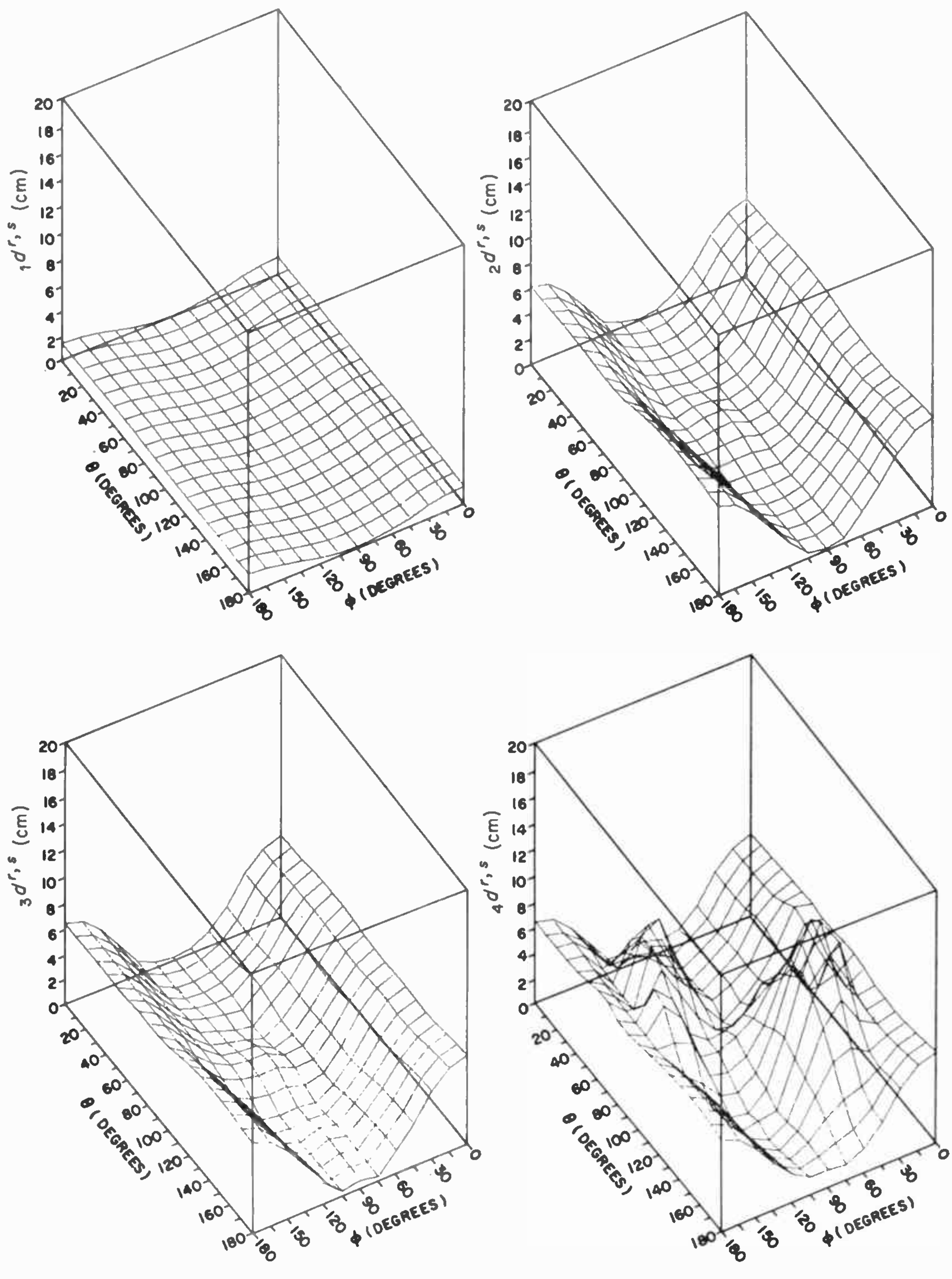

Fig. 17. Interclass distance between *F104* and *Mig 19* using v.p. amplitude returns.

because the n.n. decision rule is a more powerful classifier in the high proximity circumstance.

The n.n. decision rule is capable of generating quite complex class boundary to fit the high proximity cases and involved data structure. Such a complicated discriminant boundary can only be accomplished by piecewise linear decision functions. It is known that the n.n. classifier is very simple and its performance is very good. The only disadvantages of n.n. decision is that it requires a considerable amount of storage as well as computation for a decision to be reached in the classification processes.

4. Object Classification using Both Amplitude and Phase of Radar Returns

In the previous discussion the classification of various objects was based on the amplitude of the radar returns. The returns obviously carry both phase and amplitude information, but in the interest of measurement and

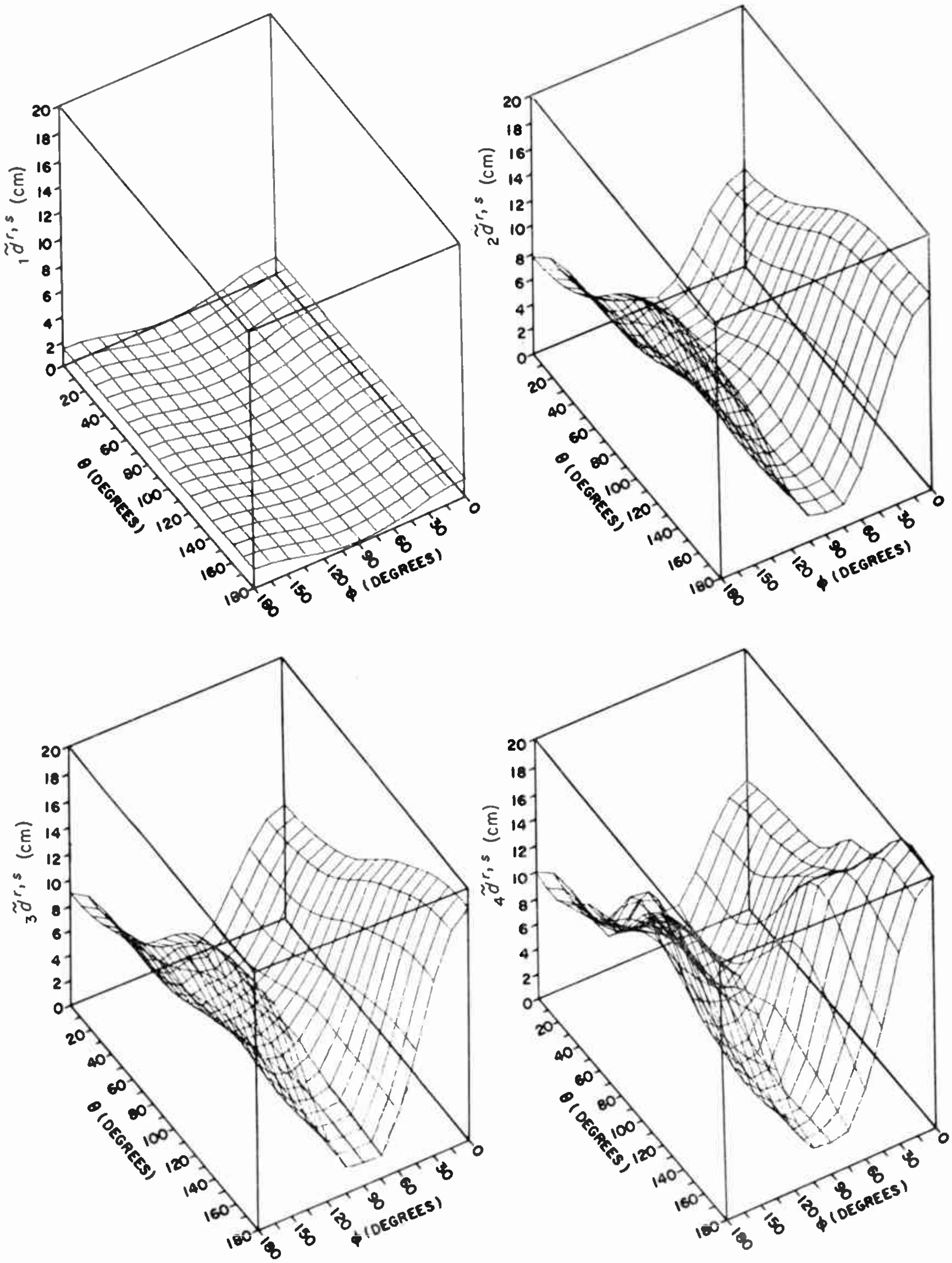

Fig. 18. Interclass distance between *F104* and *Mig 19* using v.p. complex returns.

equipment simplicity amplitude alone was used. After a thorough investigation of classification performance using amplitude alone it was appropriate to study the added benefits of utilizing phase as well. In principle it appeared that phase should contain just as much information as the amplitude, resulting in substantially reduced error probabilities. This indeed occurred, as will be described below.

From previous discussions we know that it is the distance between classes that is the dominant factor affecting misclassification probabilities. We define a complex radar return,

$$z_{ij}^k = a_{ij}^k(\cos \psi_{ij}^k + i \sin \psi_{ij}^k)$$

as the combined amplitude, a_{ij}^k, and phase ψ_{ij}^k, obtained at the jth frequency at the ith aspect of kth object.

Thus the Euclidean distance between the points corresponding to the ith aspect of class r and class s for L frequencies is defined as

$$_L\tilde{d}_i^{r,s} = [(\bar{z}_i^r - \bar{z}_i^s)^T(\bar{z}_i^r - \bar{z}_i^s)^*]^{1/2} \qquad (2)$$
$$= \left[\sum_{j=1}^{L} |z_{ij}^r - z_{ij}^s|^2\right]^{1/2}$$

where superscript * denotes conjugate, and the tilde over d signifies distance derived from complex returns. Comparing to the distance derived from only the amplitude returns ($a_{ij}^k = |a_{ij}^k|$)

$$_L d_i^{r,s} = \left[\sum_{i=1}^{L} (|z_{ij}^r| - |z_{ij}^s|)^2\right]^{1/2}, \qquad (3)$$

we can see that

$$_L\tilde{d}_i^{r,s} \geq {_L d_i^{r,s}}. \qquad (4)$$

Using the complex returns, i.e., phase and amplitude data, the quantities $_L d^{r,s}$ and $_L\tilde{d}^{r,s}$ for *F104* and *Mig 19* are plotted in Figs. 17 and 18. From these figures it is apparent that the interclass distance is significantly increased by the introduction of phase, especially for the second and third frequencies. The introduction of phase to the first frequency return does not increase the interclass distance since the scattering phase of targets in the low Rayleigh region is negligibly small. The fourth frequency does not contribute significantly to the average interclass distance, it accentuates however local distance variations since the scattering return at the higher frequencies are characterized by the detail of the target. A computation of the interclass distance for all the aspects shows that the average $_L\tilde{d}^{r,s}$ is about twice as large as the average $_L d^{r,s}$. Therefore, using the complete complex returns, we can expect that the error probability will be substantially reduced as compared to the use of amplitude alone.

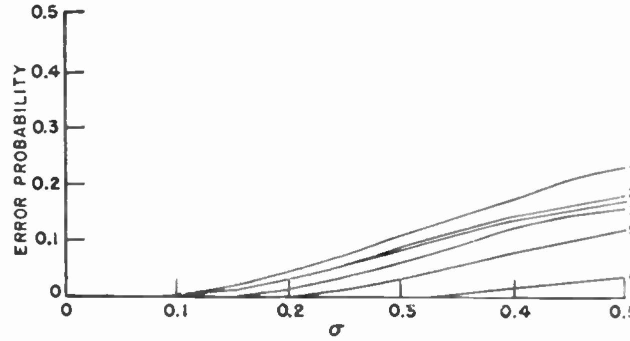

Fig. 21. Average probability of misclassification by linear classifier. Computed 4 frequencies, v.p. complex returns, $\theta = 30°$, $\phi = 45°$, 18 pts.

1. *F104* vs. *Mig 19* $\bar{A} = 2{\cdot}50$
2. *F104* vs. *F4* $\bar{A} = 2{\cdot}73$
3. *F104* vs. *Mig 21* $\bar{A} = 2{\cdot}31$
4. *Mig 19* vs. *F4* $\bar{A} = 2{\cdot}89$
5. *Mig 19* vs. *Mig 21* $\bar{A} = 2{\cdot}50$
6. *F4* vs. *Mig 21* $\bar{A} = 2{\cdot}66$

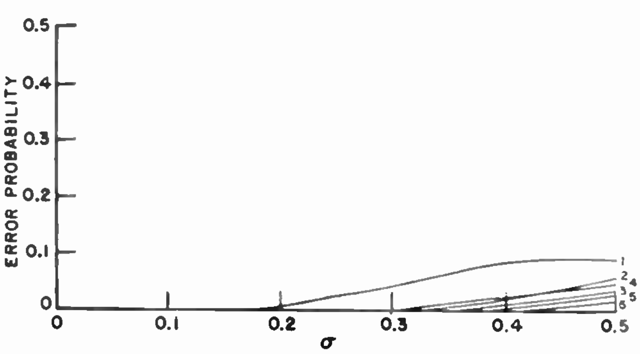

Fig. 19. Average probability of misclassification by n.n. classifier. Computed 4 frequencies, v.p. complex returns, $\theta = 30°$, $\phi = 45°$, 18 pts.

1. *F104* vs. *Mig 19* $\bar{A} = 2{\cdot}67$
2. *F104* vs. *F4* $\bar{A} = 2{\cdot}73$
3. *F104* vs. *Mig 21* $\bar{A} = 2{\cdot}31$
4. *Mig 19* vs. *F4* $\bar{A} = 2{\cdot}89$
5. *Mig 19* vs. *Mig 21* $\bar{A} = 2{\cdot}5$
6. *F4* vs. *Mig 21* $\bar{A} = 2{\cdot}56$

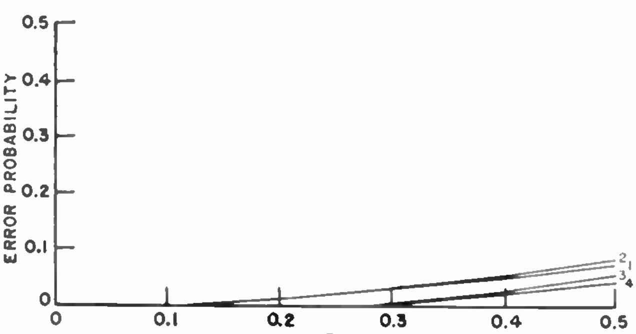

Fig. 20. Average probability of misclassification by n.n. classifier. Computed 4 frequencies, v.p. complex returns, $\theta = 30°$, $\phi = 45°$, 36 pts, $\bar{A} = 2{\cdot}62$.

1. *F104* vs. *Mig 19*, *F4* and *Mig 21*
2. *Mig 19* vs. *F4*, *Mig 21* and *F104*
3. *F4* vs. *Mig 21*, *F104* and *Mig 19*
4. *Mig 21* vs. *F104*, *Mig 19* and *F4*

Using L frequency complex returns, the minimum distance from an unknown sample $\bar{y}$ to any point $\bar{z}^k$ of the kth sample cluster can be computed according to eqn. (2), i.e.,

$$\tilde{d}_i^k = \left[\sum_{j=1}^{L} |z_{ij}^k - y_{ij}|^2\right]^{1/2} \qquad (5)$$

The n.n. rule decides that

$$\bar{y} \in C^r \quad \text{if} \quad d(\bar{y}, \bar{z}^r) = \min_{i,k} \tilde{d}_i^k \qquad (6)$$

where $\bar{z}^r$ is any member of the $\{\bar{z}_i^r\}$. The estimate of the average probability of misclassification for the n.n. rule using complex returns were carried out using the same technique as in the amplitude case. The experimental results for pairwise and multiclass cases are presented in Figs. 19 and 20 for the n.n. classifier and Figs. 21 and 22 for the linear classifier. Comparing the results obtained by using amplitude returns only, it is clear that a very impressive improvement in probability of misclassification resulted. The reduction of error probabilities is substantially better than 50%, this might be expected due to doubling the distance between classes. The reason for the drastic improvement can be seen from the illustration of Fig. 23. Assuming that the original distance between classes is d with class 1 centred at α_1 and class 2 at α_2, the curves represent the probability densities arising out of noise and error contamination which cause the points of each class to be displaced from their original location

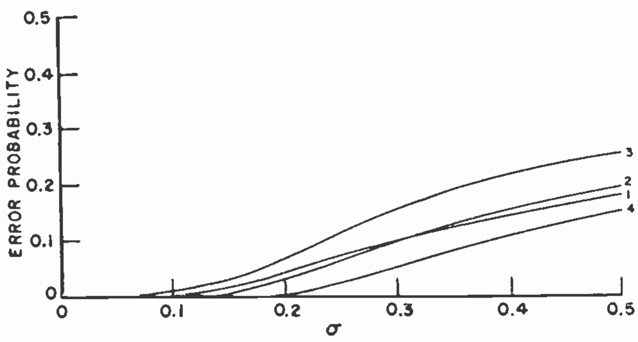

Fig. 22. Average probability of misclassification by linear classifier. Computed 4 frequencies, v.p. complex returns,

$\theta = 30°$, $\phi = 45°$, 36 pts, $\bar{A} = 2{\cdot}62$.

1. *F104* vs. *Mig 19*, *F4* and *Mig 21*
2. *Mig 19* vs. *F4*, *Mig 21* and *F104*
3. *F4* vs. *Mig 21*, *F104* and *Mig 19*
4. *Mig 21* vs. *F104*, *Mig 19* and *F4*

α_1 and α_2, respectively. If the threshold is chosen as $\alpha = (\alpha_1+\alpha_2)/2$, the error probability is the cross-hatched area shown in the figure. If the distance between the classes is now doubled to $2d$, the distribution will be represented by the solid curves with one centred at α_1 and the other at α'_2. The threshold would now be at $\alpha = (\alpha_1+\alpha'_2)/2$ and the resulting error probability would be negligibly small compared to the previous one.

5 Object Classification using Amplitude, Phase and Polarization of Radar Returns

The identification process so far utilized a linearly polarized return to classify the samples into classes. To further minimize the number of frequencies required for reliable classification, other possible parameters are examined. If we assume for a given observation angle the horizontal polarization returns as a descriptor of the length and width of an aircraft and the vertical polarization returns as a descriptor of its height (see Fig. 6), it is reasonable to expect that using both vertical and horizontal polarization returns would give a better description of an aircraft than is obtained with a single polarization return. The availability of a pair of orthogonal polarizations enables one to double the feature space. Thus the polarization parameter may provide an effective substitute for bandwith requirements.

Arbitrarily polarized incident and scattered waves may be resolved into orthogonal linearly polarized components. The directions of the orthogonal reference axes are arbitrary but without loss of generality we can choose those shown in Fig. 24. The coordinates for incident and scattered waves have a common x axis perpendicular to the plane of scattering, target directed z axis, and y axis in the plane of scattering. In the special case of backscattering, the two coordinate frames coincide. The rectangular components of the vector field intensity are expressed in the conventional complex form to denote amplitude and relative phase of each component. The relationship between the components of a plane wave $\mathbf{E}^i$ incident on a target and the scattered plane wave $\mathbf{E}^s$ is given by a matrix equation:

$$\mathbf{E}^s = \begin{bmatrix} E^s_x \\ E^s_y \end{bmatrix} = \begin{bmatrix} S_{11} & S_{12} \\ S_{21} & S_{22} \end{bmatrix} \begin{bmatrix} E^i_x \\ F^i_y \end{bmatrix}. \tag{7}$$

The square matrix $[S]$ is referred to as the scattering matrix of the target. The values of the four complex components of the scattering matrix depend on the aspect of the target, the scattering properties of the target and the frequency.

It can be shown that the scattering matrix can be transformed to represent scattering in terms of an arbitrarily selected pair of orthogonal reference polarizations. Lowenschuss[6] demonstrated that target rotation about the radar line of sight is equivalent in effect to rotation of the transmitting and receiving antennas about the line of sight. Thus the object rotation about the line of sight or the rotation of the plane of polarization contribute to additional information regarding the scattering properties of the object. The monostatic scattering matrix $[S]$ considered can be arbitrarily defined by two orthogonal linear polarizations, say, vertical and horizontal. Then $[S]$ is a symmetric matrix and can be defined by the six real parameters, three amplitudes and three phases, as

$$(S) = \begin{bmatrix} {}^{\mathrm{VV}}a \exp(\mathrm{j}\,{}^{\mathrm{VV}}\psi) & {}^{\mathrm{VH}}a \exp(\mathrm{j}\,{}^{\mathrm{VH}}\psi) \\ {}^{\mathrm{VH}}a \exp(\mathrm{j}\,{}^{\mathrm{VH}}\psi) & {}^{\mathrm{HH}}a \exp(\mathrm{j}\,{}^{\mathrm{HH}}\psi) \end{bmatrix} \tag{8}$$

where ${}^{\mathrm{VV}}a$ corresponds to the amplitude of the vertically polarized return from a target illuminated by a vertically polarized wave, etc. The parameters ${}^{\mathrm{VV}}a$, ${}^{\mathrm{VH}}a$, ${}^{\mathrm{HH}}a$, ${}^{\mathrm{VV}}\psi$, ${}^{\mathrm{VH}}\psi$ and ${}^{\mathrm{HH}}\psi$ can be interpreted as forming a six-dimensional orthogonal vector space. A single scattering matrix at a given observation aspect corresponds to a single point in this vector space. We can define a 6-tuple feature vector

$$\begin{aligned} S^k_{ij} = (&{}^{\mathrm{VV}}a^k_{ij} \cos {}^{\mathrm{VV}}\psi^k_{ij}, \quad {}^{\mathrm{VV}}a^k_{ij} \sin {}^{\mathrm{VV}}\psi^k_{ij}, \\ &{}^{\mathrm{VH}}a^k_{ij} \cos {}^{\mathrm{VH}}\psi^k_{ij}, \quad {}^{\mathrm{VH}}a^k_{ij} \sin {}^{\mathrm{VH}}\psi^k_{ij}, \\ &{}^{\mathrm{HH}}a^k_{ij} \cos {}^{\mathrm{HH}}\psi^k_{ij}, \quad {}^{\mathrm{HH}}a^k_{ij} \sin {}^{\mathrm{HH}}\psi^k_{ij}). \end{aligned}$$

Accordingly, the distance between two sample points corresponding to the ith aspect of class r and class s for L frequencies is defined as the Euclidean distance in $6L$-tuple space:

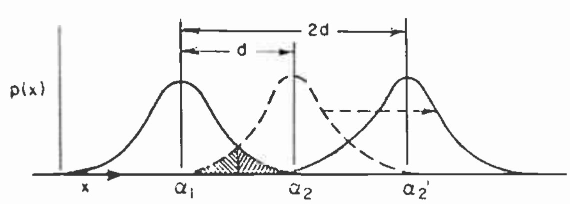

Fig. 23. Error probability as a function of interclass distance.

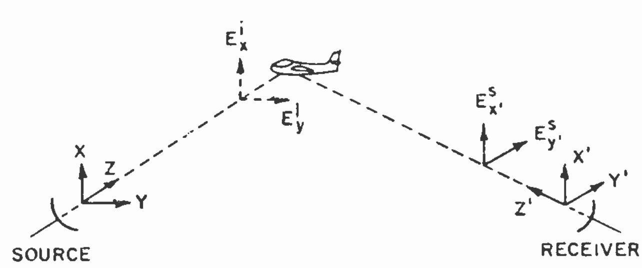

Fig. 24. Radar range geometry.

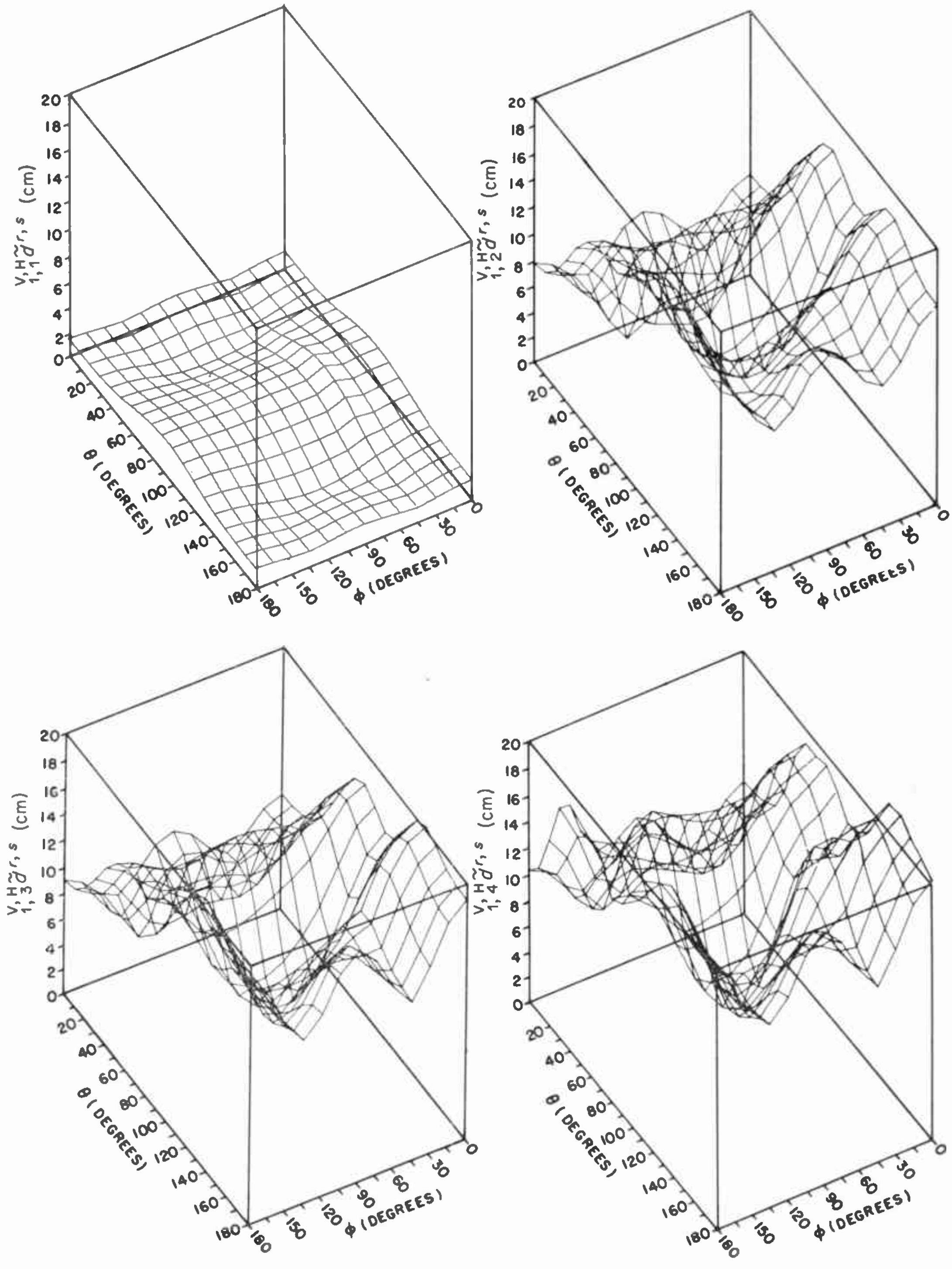

Fig. 25. Interclass distance for *F104* vs. *Mig 19* using both v.p. and h.p. complex returns.

$$ {}^{V,H}_{L}d_i^{r,s} = \left[\sum_{i=1}^{L} \left|s^r_{i,j} - s^s_{ij}\right|^2\right]^{1/2}. \tag{9}$$

Comparing this with the distance derived in eqns. (2) and (3), we note the following relationship

$$ {}_L d_i^{r,s} \le {}_L\tilde{d}_i^{r,s} \le {}^{V,H}_{L}\tilde{d}_i^{r,s}. \tag{10}$$

Using the computed vertical and horizontal polarization complex returns to form a $4L$-dimensional feature space, omitting the cross polarized components, the interclass distance ${}^{V,H}\tilde{d}^{r,s}$ for *F104* and *Mig 19* are plotted in Fig. 25. In general the cross-polarization terms are smaller than either the vertical or horizontal polarization returns. It is a good measure of the symmetry of the target around the line of sight. The cross-polarization terms were not used in the examples because they were not available. Comparing Fig. 25 with Figs. 17 and 18, it can be seen that the interclass distance is substantially increased by using both vertical and horizontal returns.

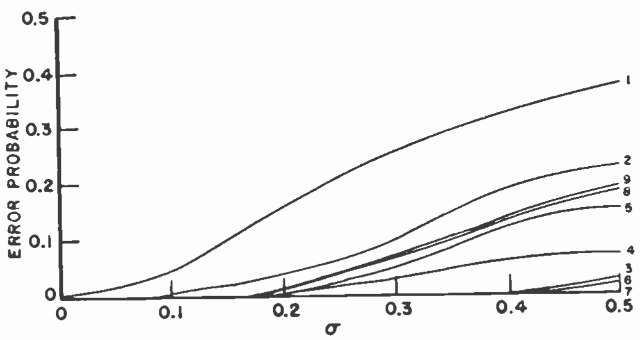

Fig. 26. Average probability of misclassification for *F104* vs. *Mig 19* by n.n. classifier at $\theta = 30°$, $\phi = 45°$.

1.	4 freq.	h.p.	amplitude returns	$\bar{A} = 2{\cdot}74$
2.	4 freq.	v.p.	amplitude returns	$\bar{A} = 2{\cdot}67$
3.	4 freq.	h.p.	complex returns	$\bar{A} = 2{\cdot}74$
4.	4 freq.	v.p.	complex returns	$\bar{A} = 2{\cdot}67$
5.	first 2 freq.	v.p. & h.p.	complex returns	$\bar{A} = 2{\cdot}54$
6.	first 3 freq.	v.p. & h.p.	complex returns	$\bar{A} = 2{\cdot}72$
7.	4 freq.	v.p. & h.p.	complex returns	$\bar{A} = 2{\cdot}62$
8.	4 freq.	v.p. & h.p.	amplitude returns	$\bar{A} = 2{\cdot}62$
9.	first 3 freq.	v.p. & h.p.	amplitude returns	$\bar{A} = 2{\cdot}72$

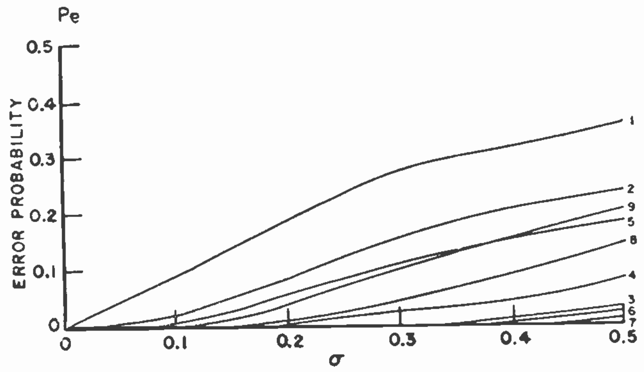

Fig. 27. Average probability of misclassification for *F104* vs. *Mig 19, F4* and *Mig 21* by n.n. classifier at $\theta = 30°$, $\phi = 45°$.

1.	4 freq.	h.p.	amplitude returns	$\bar{A} = 2{\cdot}88$
2.	4 freq.	v.p.	amplitude returns	$\bar{A} = 2{\cdot}62$
3.	4 freq.	h.p.	complex returns	$\bar{A} = 2{\cdot}88$
4.	4 freq.	v.p.	complex returns	$\bar{A} = 2{\cdot}62$
5.	first 2 freq.	v.p. & h.p.	complex returns	$\bar{A} = 2{\cdot}55$
6.	first 3 freq.	v.p. & h.p.	complex returns	$\bar{A} = 2{\cdot}76$
7.	4 freq.	v.p. & h.p.	complex returns	$\bar{A} = 2{\cdot}75$
8.	4 freq.	v.p. & h.p.	amplitude returns	$\bar{A} = 2{\cdot}75$
9.	first 3 freq.	v.p. & h.p.	amplitude returns	$\bar{A} = 2{\cdot}76$

Figures 26 and 27 demonstrate the effectiveness of the various feature sets for pairwise and multiclass cases at observation angle $(\theta, \phi) = (30°, 45°)$. As expected, the best performance is obtained by using the features with the highest dimensionality, that is, the four frequency v.p. and h.p. complex returns. Using the second highest number of features, including the first 3 frequencies, v.p. and h.p. complex returns, yields the next lowest error probability. The use of only the first two frequencies, v.p. and h.p. complex returns does not yield as good results as the use of four-frequency single-polarization complex returns. The reason for this is that the 2nd and 3rd frequencies provide most discrimination. It is also interesting to note that the use of 4-frequency v.p. or h.p. amplitude returns alone is not as effective as the use of 4-frequency single polarization complex returns, even though they have the same total number of dimensions in feature space. It can be concluded that the most effective use of a radar return would include both the phase and polarization parameters. Thus two-frequency samples appear to be quite adequate for identifying an object if complex returns at both vertical and horizontal polarizations are available.

6 Conclusion

The objective of this research was to examine the effectiveness of low-frequency electromagnetic features for identifying objects of complex shapes. The results have shown that reliable classification can be obtained utilizing relatively straightforward techniques. The separability of various objects with simple shapes was first investigated using a linear discriminant method. A study of perturbation effects on separability was then carried out to determine the sensitivity of the classification process to deviations of the test sets from the learning set. The results indicated that the classifier was indeed tolerant to substantial changes.

With regard to more complicated shapes, linear separability of aircraft was attempted. Two models, an *F104* and a *Mig 19* were examined and found separable for all aspect angles tested. However, as the number of alternative aspect angles or aircraft classes was increased, a linear discriminant was found lacking. The n.n. rule was then chosen as the more appropriate classifier to handle these more convolved data structures. The n.n. classifier is approximately equivalent to a piecewise linear classifier where the decision boundary is determined by the local distribution of data, or learning points, surrounding the unknown or test point. Thus the object can be reliably classified as long as the classes are not overlapping in the region of interest. The results have shown that using the n.n. rule the various aircraft were indeed classified reliably in spite of the increased number of aspect angles used. In fact, two additional aircraft, an *F4* and a *Mig 21* were introduced and reliable classification was obtained for all aspect angles.

The scattering data obtained by measurement or computation carry both amplitude and phase information. For reasons of simplicity only amplitude information was initially used. It was, however, expected that the use of both amplitude and phase information would approximately double the information content of the radar returns. Phase information was therefore introduced and the performance using amplitude and phase was compared to that of using amplitude alone. The results have shown that phase information plays just as important a role as amplitude and drastically reduces the probability of misclassification when introduced. Although the original classification tests were carried out utilizing ten frequency samples, it has been shown that as few as four frequencies can provide adequate classification performance. The identification process to this point utilized amplitude and phase returns at a set of frequencies with a single linear polarization to form the feature space. It was found that the interclass distance could be increased substantially by using a pair of orthogonal polarization returns, such as vertical and horizontal polarization. Indeed it has been demonstrated that by taking advantage of electromagnetic parameters such as amplitude, phase, and polarization as few as two to three well-chosen frequency samples can provide adequate classification performance even in the presence of substantial noise levels.

The classification techniques considered here used a single observation vector to make a decision via linear discriminants or n.n. rule. Since it may be assumed a sequence of return pulses would ordinarily be available, it is suggested that averaging the sequence of observation vectors prior to classification would have the effect of filtering the noise and reducing the probability of misclassification.

It has been shown that a set of four aircraft models scaled to approximately the same size can be reliably identified by means of the technique considered in this research effort. The question arises as to the classifiability of aircraft of actual size. Preliminary test results indicate that the scattering data from actual size aircraft would provide better separation in feature space than those scaled to the same length. Further study in this line seems very promising.

The classification algorithms used in this study were rather simple linear discriminants and the n.n. rule. More complicated and effective classification techniques are available. This study was aimed to determine a set of radar measurable electromagnetic features to which pattern recognition could be applied. The fact that good class separation was achieved should lead to the application of more sophisticated classification techniques to further improve reliability of identification.

7 Acknowledgment

The work reported in this paper was supported in part by Grant No. AFOSR-74-2611 between Air Force Office of Scientific Research, and The Ohio State University Research Foundation.

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Manuscript first received by the Institution on 11th August 1975 and in final form on 31st October 1975. (Paper No. 1735/AMMS 72.)

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USC 681.327.2

The comparative cost of associative memory

R. M. LEA, M.Sc., A.R.C.S., C.Eng., M.I.E.E.*

SUMMARY

The production cost-per-bit of high-speed associative memory, implemented with m.o.s. l.s.i. content addressable memory (c.a.m.) devices (selected from a survey of 20 designs), is compared with that of conventional memory, implemented with currently available m.o.s. l.s.i. random access memory (r.a.m.) devices.

For a development cost of about £20k, the cost-per-bit of the m.o.s. c.a.m.s is about 5× that of a 1024-bit p-channel m.o.s. dynamic r.a.m. and about 9× that of a 4096-bit n-channel m.o.s. dynamic r.a.m., if plastic dual-in-line packaging is employed. However, for open-chip (beam-lead or flip-chip) assembly the cost-per-bit of the m.o.s. c.a.m.s is less than 2× that of the 1024-bit r.a.m. and less than 3× that of the 4096-bit r.a.m.

It is shown that for most associative processing applications the associative memory can provide a cheaper alternative to conventional r.a.m.-based computer systems.

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List of Symbols

C_A	chip assembly and packaging cost
C_P	chip preparation cost
C_{PW}	average cost of processing a 2-in (51-mm) wafer
C_T	final device testing cost
C_W	cost of a 2-in wafer
n	average number of usable chips per wafer
N	number of complete chips per 2-in wafer
WCB	works-cost-per-bit
WCP	works-cost-price
Y	percentage yield of usable chips

The subscripts '23', '24' and 'S' associate the symbol with the m.o.s. c.a.m. of reference 23, m.o.s. c.a.m. of reference 24 and the standard 1024-bit p-channel m.o.s. dynamic r.a.m. respectively.

Encapsulation in plastic dual-in-line packaging is assumed unless the symbol is subscripted with 'O' which refers to open-chip (beam-lead or flip-chip) assembly.

1 Introduction

The potential processing power and flexibility of the associative memory have stimulated a very large number of research investigations over a period of nearly twenty years.[1–3] The combined features of content-addressing and parallel processing endow the associative memory with a high hardware operational efficiency. Moreover, this efficiency can be maintained over a wide range of information processing applications without incurring software support.[4] Consequently the current growth of interest in non-numerical computer applications will ensure a new revival of interest in the associative memory. Despite the high degree of innovatory effort already expended, there is still no well-established implementation technology for cost-effective associative memory hardware. First there was the woven cryotron, which led to the development of the cryogenic memory. The apparent availability of cheap associative memory caused a proliferation of research proposals for associative processor development.[1,2] Next came non-cryogenic magnetic techniques which stimulated still further interest in the associative memory.[1,2] However, the exacting technological requirements of the associative memory prevented its cost-effective implementation by these techniques. By the mid-1960s the basic architecture of an associative processor and its processing advantages were well appreciated but hopes for its availability were falling fast. More recently, the rapid advances in l.s.i. fabrication techniques have brought about a renewed interest in associative memory implementation.[3,5] Many designs for l.s.i. content addressable memories (c.a.m.s) have been proposed. However, by this time, the momentum of the demand for associative processing had been lost and the various different claims of the l.s.i. manufacturers only increase its inertia. In fact, there is a widespread belief that associative memory hardware is inherently expensive and will always remain cost-ineffective when in comparison with conventional computing hardware. The purpose of this paper is to prove that this contention is no longer valid.

2 C.A.M. Design Requirements

Low-cost content addressable memory (c.a.m.) devices are required as cheap building blocks for cost-effective associative memory implementation. The regular array-like structure of a c.a.m. device, as shown in Fig. 1, makes it particularly well suited to the batch-fabrication techniques of current l.s.i. technology. To achieve low-cost associative memory it is necessary to:

1. Maximize the number of bits which can be supported in a cheaply packaged c.a.m. device. This entails minimizing the number of transistors, 'bit-lines', 'word' lines and voltage supply lines per 1-bit cell, reducing the power dissipation per cell and the total number of chip terminals.
2. Maximize the number of cheap c.a.m. devices which can be controlled by cheap interface logic. This entails minimizing both the static and dynamic power dissipation of the c.a.m. and maximizing both the input noise immunity and the output signal/noise ratio.

From these design requirements it is clear that m.o.s. l.s.i. technology is well suited to c.a.m. fabrication.[4] Unfortunately the recent advances in m.o.s. l.s.i. random access memory (r.a.m.) device manufacture do not apply to the c.a.m., for the following reasons:

1. On-chip decoding cannot be utilized because of the parallel input/output requirement of the c.a.m. Hence the c.a.m. suffers a high pin-count as shown in Fig. 1.
2. Dynamic charge storage can be applied to c.a.m.s, but not to the same degree as with r.a.m.s because of the extra comparison logic of the c.a.m. cell.
3. C.a.m. devices require a higher operating speed than r.a.m.s. A simple analysis[5] has shown that, for cost-effective associative memory, c.a.m. speeds should be as follows:

   Match output delay < 10 ns

   Read output delay < 35 ns

   Write 'toggle' time < 45 ns

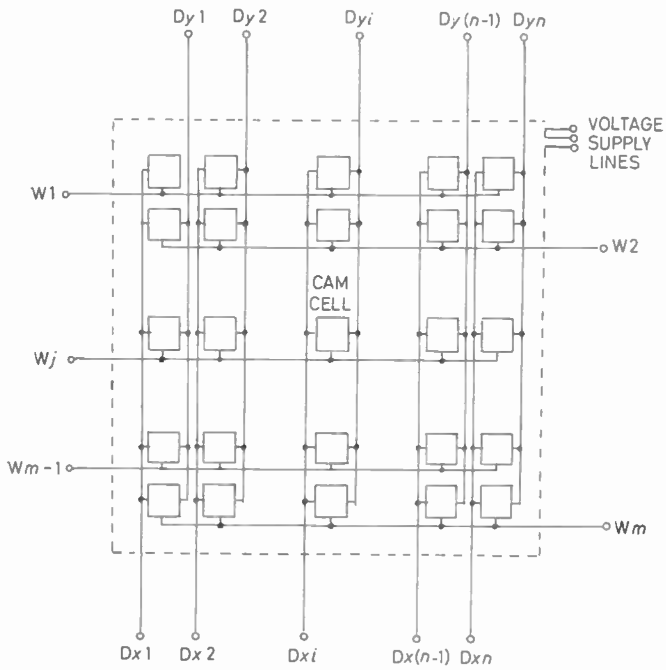

Fig. 1. Schematic content addressable memory (c.a.m.) array.

M.o.s. r.a.m. speeds are an order of magnitude higher than this requirement. Nevertheless, many m.o.s. l.s.i. c.a.m.s have been produced. Moreover, it has been shown that the m.o.s. l.s.i. fabrication process is far more suitable than the equivalent bipolar technology for c.a.m. production.[5] In fact, it is only the ease of interfacing bipolar c.a.m.s with t.t.l. system logic that justifies their existence. Since the cost of special interface circuitry is minimized by the large number of m.o.s. c.a.m. devices it can support, this requirement is inconvenient rather than expensive. Furthermore, the current interest in low-threshold n-channel m.o.s. l.s.i. devices could well remove the need for interface circuitry.

3 M.O.S. L.S.I. C.A.M. Development

At the time of writing, twenty designs for m.o.s. content addressable memories have been published. The first was reported by Igarashi *et al.*[6] of the Nippon Electric Company, Japan. The cell comprised six m.o.s. transistors and two thin-film resistors. A single cell was implemented on a 36×40 mil† chip and a 150 ns cycle time was recorded. The cell design was later improved by replacing the thin-film resistors with m.o.s. transistors and a 4-bit array was constructed.[7] The new chip allowed a cycle time of 100 ns to be measured. Two years later Herlein and Thompson, of American Microsystems Incorporated, reported a 64-bit m.o.s. c.a.m.,[8] constructed for the NASA Marshall Space Flight Center. The device was organized as 8 words of 8 bits each, and a cycle time of 1 μs was measured. In the same year Burns and Scott, of RCA Laboratories, described a silicon-on-sapphire complementary m.o.s. c.a.m.[9] A 16-bit test vehicle was built and access times of the order of 10 ns were measured. Another complementary m.o.s. cell design was proposed in 1970 by Koo, of Bell Telephone Laboratories.[10] This cell comprised 6 p-channel and 2 n-channel m.o.s. transistors. No performance data were given. During the same year four cell designs for m.o.s. c.a.m. were disclosed by IBM.[11,12] Again, no performance data were recorded.

The first widely reported m.o.s. c.a.m. was announced by Wald, of Honeywell Incorporated in 1970. A 128-bit array, organized as 16 words of 8 bits each, was fabricated on a 130×110 mil chip.[13] The cell comprised 10 m.o.s. transistors and occupied about 75 sq. mils of silicon substrate. Cycle times of the order of 300 ns were recorded. The device was manufactured by Texas Instruments and for a period it was made commercially available. Another m.o.s. transistor cell design was proposed by Koo in the same year.[14] This cell comprised 9 m.o.s. transistors. Although a prototype was not constructed a search time of 400 ns was reported. In 1971, Mundy *et al.*, of the General Electric Company

† 1 mil = 0·001 in = 0·0254 mm.

(USA), described a five-transistor m.o.s. c.a.m. cell,[15,16] The cell is 'dynamic' in operation and, consequently, requires regular refreshing by external circuitry. A 72-bit test vehicle was fabricated in molybdenum-gate m.o.s. technology, and search times of the order of 200 ns were measured. The following year another dynamic cell design incorporating a special m.o.s.v.a.c. element was published by the same team.[16,17] A 512-bit array was integrated in a 100 × 150 mil chip and became the largest m.o.s. c.a.m. chip yet constructed. A search time of 300 ns was recorded for this device. Kaiser and Collins, of RCA and the US Naval Research Laboratory respectively, reported an associative processor microelectronic element (a.p.m.e.) in 1972.[18] The device comprised 63 associative cells, fabricated in c.o.s.m.o.s. technology, and occupied 155 × 161 mil of silicon substrate. A search time of 250 ns was recorded for the a.p.m.e.

In 1972 the author described a 64-bit m.o.s. static c.a.m.[19] This device was produced as part of an ACTP contract at GEC–Marconi Ltd. to investigate the suitability of l.s.i. microelectronic techniques for low-cost high-speed associative memory implementation. The device was integrated on a 120 × 120 mil chip, as shown in Fig. 2, and its cell design comprised 14 m.o.s. transistors, as shown in Fig. 3. Cycle times of less than 15 ns were measured for this device. The chip was encapsulated in a 28-pin ceramic dual-in-line package and samples were distributed to gauge the commercial viability of the device. In an attempt to improve on the cell packing density and interface compatibility, a 4-bit test vehicle fabricated by the silicon-gate m.o.s. process with double-level metallization was also produced.[20] The chip, as shown in Fig. 4, incorporated a slightly modified version of the cell design shown in Fig. 3. To investigate the applicability of open-chip assembly techniques to low-cost associative memory implementation, a thick-film substrate was produced to support 16 of the 64-bit c.a.m. chips by the beam-lead technique, as shown in Figs. 5 and 6. The substrate, measuring only 2 in by 3 in, provided a very convenient building block (64 words of 16 bits each) for low-cost associative memory implementation.

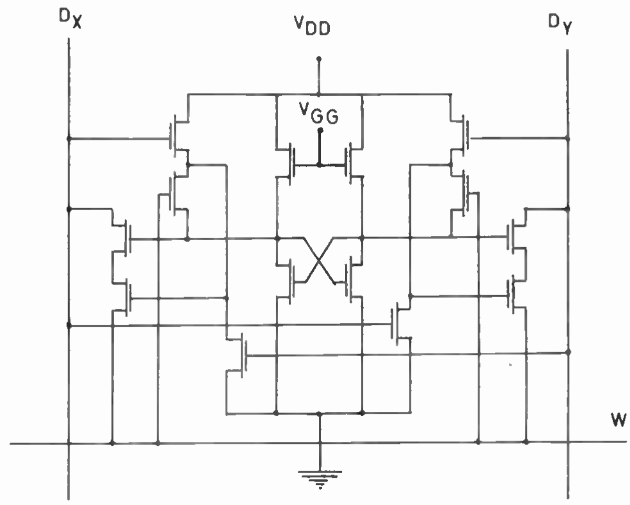

Fig. 3. Cell design for the 64-bit m.o.s. c.a.m. shown in Fig. 2.

In 1973, Carlstedt of Chalmers University (Sweden) described a 2-transistor m.n.o.s. c.a.m. cell.[21] This cell could provide non-volatile associative storage and occupy less than 8 sq. mils of silicon substrate. A 4-bit test vehicle, incorporating 4 m.n.o.s. transistors per cell was fabricated, but no performance data were reported. Another attempt to open up a market for m.o.s. c.a.m.s was made, by Solid State Scientific Incorporated, in February of 1975. The 64-bit array, organized as 8 words of 8 bits each and fabricated using c.-m.o.s. technology, is encapsulated in a 48-pin ceramic dual-in-line pack. The manufacturer claimed a 110 ns interrogate time, a 150 ns read access time and a quiescent power dissipation of only 25 μW for this static c.a.m.[22]

In April 1975 the auther described in this Journal the design of a 128-bit m.o.s. c.a.m.[23] The memory could be integrated on a 63 × 114 mil chip using the silicon-gate

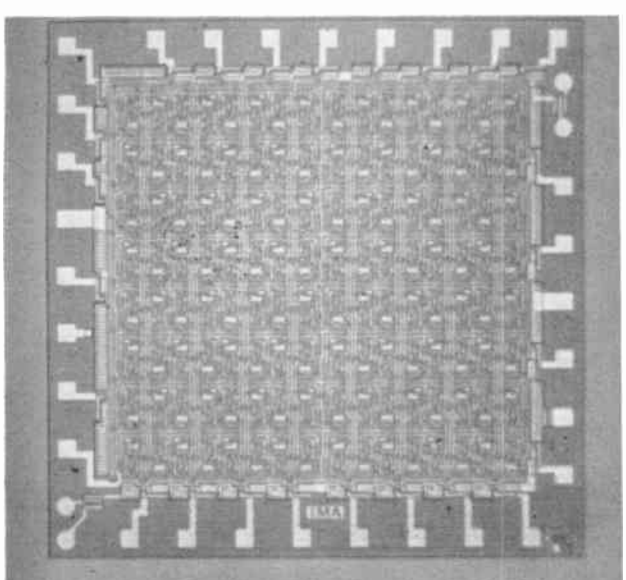

Fig. 2. 64-bit m.o.s. c.a.m.[19]

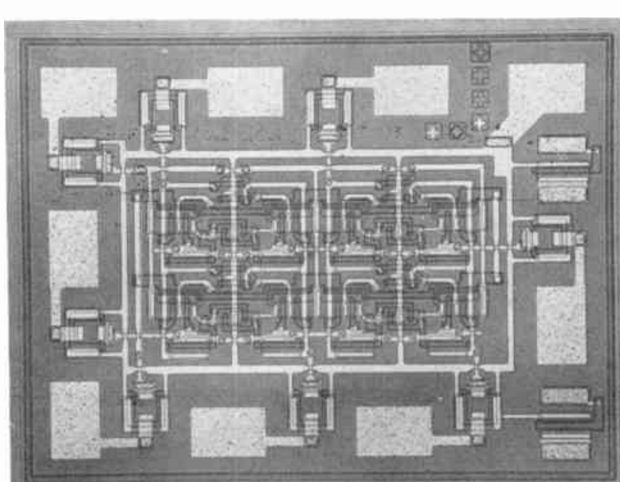

Fig. 4. 4-bit silicon-gate m.o.s. c.a.m. test vehicle.[20]

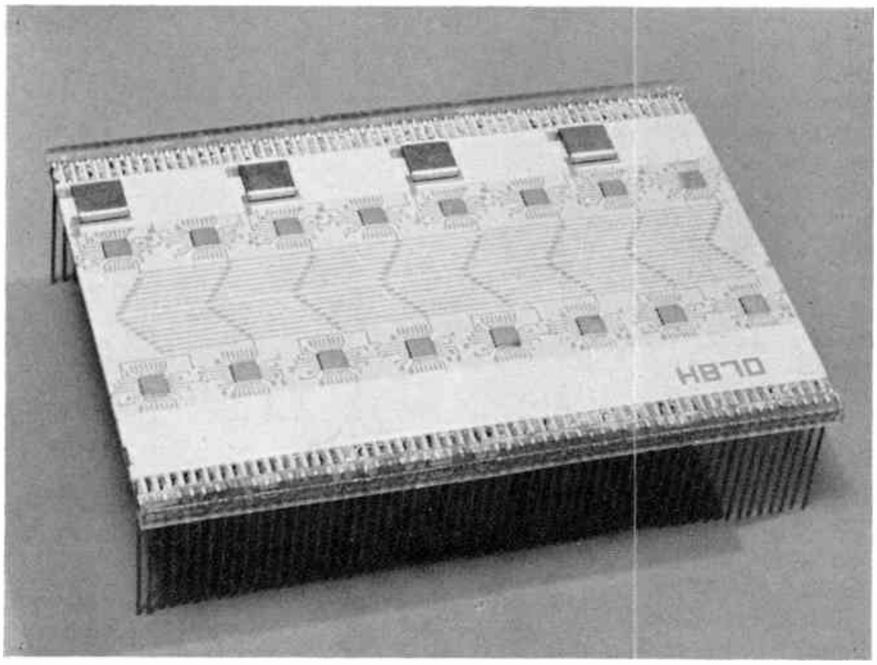

Fig. 5. Thick-film substrate supporting 16 of the 64-bit m.o.s. c.a.m.s shown in Fig. 2.

m.o.s. fabrication technology. The memory cell, comprising 6 m.o.s. transistors as shown in Fig. 7, was dynamic in operation, but external refresh circuitry was not required. Computer simulation studies predicted match and read access times of 10 ns and a write 'toggle' of 25 ns. An open-chip assembly method, such as that shown in Figs. 5 and 6, would allow a 256-bit version of the design, with similar performance, to be integrated on a 100×114 mil chip.[23] Another design for a 128-bit m.o.s. c.a.m. was described in June 1975.[24] This design could be operated as a 128-bit c.a.m. (16 words of 8 bits each) or a 256-bit c.a.m. (16 words of 16 bits each) with reduced functional capability. The memory cell, comprising four m.o.s. transistors as shown in Fig. 8, was dynamic in operation and external circuitry was required for refreshing purposes. The memory could be integrated on a 60×80 mil chip using silicon-gate m.o.s. fabrication technology. Computer simulation studies predicted match, read and write access times of less than 10 ns. Open-chip assembly was also considered for this device. A 512-bit version of the design, integrated on a 100×140 mil chip could be supported in this way. This device would be organized as 32 words of 16-bits each or as 32 words of 32-bits each, with reduced functional capability.

It is evident, from the above survey of m.o.s. c.a.m.s that some considerable experience has been gained in this area. The main characteristics of these devices are summarized in Table 1.

It can be seen from Table 1 that references 23 and 24 describe the most attractive c.a.m. implementations yet reported. All of the designs could benefit in speed, chip size and interface compatibility by a conversion from p-channel to n-channel m.o.s. transistors. However, even in this event the cell circuit designs of references 23 (shown in Fig. 7) and 24 (shown in Fig. 8) still remain the most attractive. In fact, this author doubts that a simpler c.a.m. implementation can be achieved with m.o.s. fabrication technology. Consequently if semiconductor manufacturers decide to venture into commercial c.a.m.

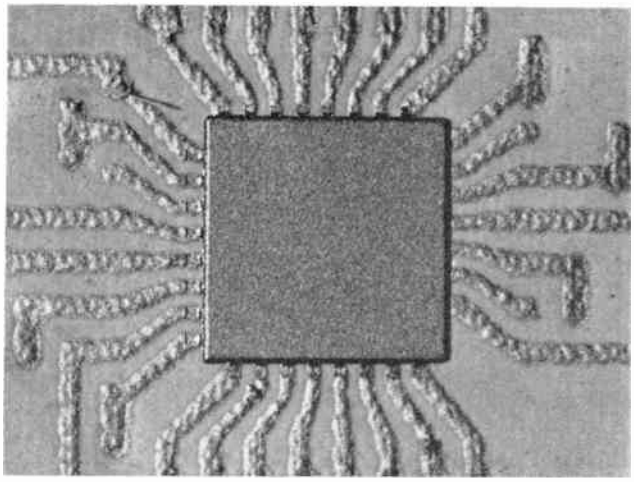

Fig. 6. Detail of a beam-leaded chip from Fig. 5.

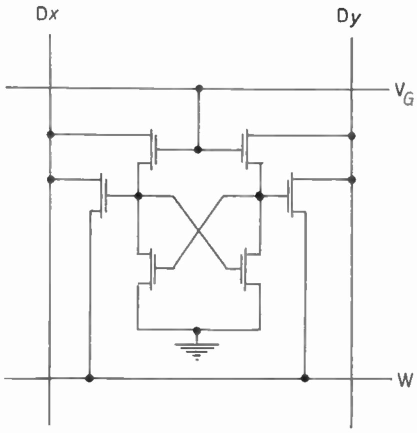

Fig. 7. Dynamic m.o.s. c.a.m. cell design.[23]

Table 1.
Summary of published implementations of m.o.s. c.a.m.s

Ref.	A	B	C	D	E	F	G	H	I
6	1	—	36 × 40	8	150	8	4	2	2
7	4	2 × 2	—	14	100	8	4	2	2
8	64	8 × 8	156 × 144	40	1000	—	—	—	—
9	16	4 × 4	77 × 53	14	10	14	2	1	2
13	128	16 × 8	130 × 110	40	300	10	2	2	4
15	72	8 × 9	77 × 50	34	200	5	2	2	0
17	512	32 × 16	150 × 100	68	300	7	2	2	0
19	64	8 × 8	120 × 120	27	15	14	2	1	3
20	4	2 × 2	49 × 37*	9	—	12	2	1	3
21	4	2 × 2	8·3 × 3·9*	4	—	2	2	2	0
22	64	8 × 8	—	48	150	—	2	2	—
23	128	16 × 8	114 × 63	36	25	6	2	2	1
	256†	16 × 16	114 × 100	52	25	6	2	2	1
24	128	16 × 8	80 × 60	35	10	4	2	2	0
	256‡	16 × 16	80 × 60	35	10	2	1	2	0
	512†	32 × 16	140 × 100	67	10	4	2	2	0
	1024†‡	32 × 32	140 × 100	67	10	2	1	2	0

* Active memory area only.
† Open-chip assembly (beam-lead or flip-chip technique).
‡ Match 0 not allowed.
A = no. of cells. F = no. of m.o.s.t.s in cell.
B = no. of words × bits. G = no. of 'bit' lines.
C = chip dimensions (mil). H = no. of 'word' lines.
D = no. of chip terminals. I = no. of voltage supply lines.
E = worst access time (ns).

production, as has been considered,[13,19,20,22] then it is indeed possible that two standard devices could emerge:

(1) A c.a.m. similar to that described in reference 23 and Fig. 7 for small special-purpose associative memory implementations.

(2) A c.a.m. similar to that described in reference 24 and Fig. 8 for larger associative memory arrays where external refresh circuitry can be economically utilized.

4 Cost Estimate for M.O.S. C.A.M. Production

Although a cost analysis of a well-established l.s.i. memory device can be attempted with reasonable accuracy, estimation of the cost of a new l.s.i. memory is highly speculative. In fact the prices quoted by semiconductor manufacturers for new l.s.i. products are notoriously irrational. Consequently quotations of cost-per-bit for m.o.s. c.a.m. devices must be treated with suspicion. To confuse the matter further, the cost of l.s.i. memory has fallen dramatically over the last five years.

The main problem in cost estimation is to predict the recovery of the high development costs associated with new l.s.i. products. For standard l.s.i. memories, which can be produced in high-volume, the development cost can be recovered fairly quickly. Moreover, if high

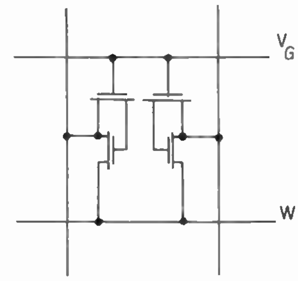

Fig. 8. Dynamic m.o.s. c.a.m. cell design requiring external refresh circuitry.[24]

volume production can be maintained the fabrication yield tends to increase, as shown in Fig. 9, due to successive improvements in the process and optimization of the chip circuit and layout designs. Hence the rapid fall in l.s.i. m.o.s. r.a.m. prices. However, for m.o.s. c.a.m. devices the production volume is an unknown quantity. Therefore, until this parameter is known, cost estimates for m.o.s. c.a.m.s are likely to be high and, consequently, discouraging. Therefore until there is an accepted market volume for l.s.i. c.a.m. devices it is essential that c.a.m. designs are compatible with existing well-established fabrication technologies. In such cases the development cost is substantially reduced and, perhaps more importantly, the c.a.m. device can be compared with a standard device, for which a cost analysis is reasonably accurate. Such a comparison is given in the following sub-sections for the two m.o.s. c.a.m.s described in references 23 and 24. The c.a.m.s are compared with a typical standard 1024-bit p-channel

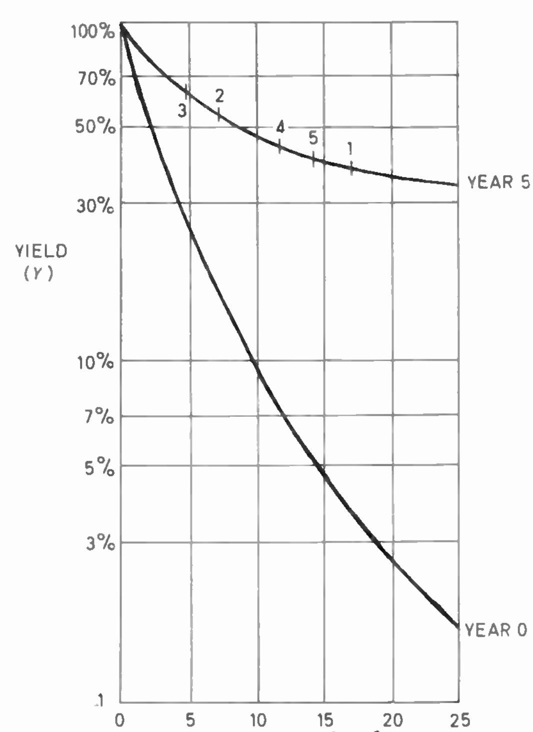

Fig. 9. Typical wafer processing yield curves (over a 5-year production period) for the p-channel silicon-gate m.o.s. fabrication process.

(silicon-gate) m.o.s. dynamic r.a.m. A cost analysis for such a device is given in the Appendix.

4.1 M.O.S. C.A.M. Development Costs

A rough estimate for the development cost of a m.o.s. c.a.m. (similar to the devices described in references 23 and 24) is of the order of £20k. This figure is based on a cost per man/month of £1k, as follows:

Design (chip circuit and layout)	£5k–£10k
Mask preparation and fabrication	£3k
Pilot production and device evaluation	£2k
Design and mask modifications	£3k–£6k
Prototype production and performance specification	£3k

These figures are rough estimates of the cost to a typical manufacturer and do not allow for commercial profit.

4.2 M.O.S. C.A.M. Production Costs

It can be seen from the Appendix that the 'works cost price' of an l.s.i. device will depend on:

1. Chip preparation cost (C_P)
2. Chip assembly cost (C_A)
3. Final device testing cost (C_T)

4.2.1 *M.o.s. c.a.m. chip preparation cost (C_P)*

The chip areas of the two m.o.s. c.a.m.s are 7182 mil^2 (114×63 mil) (Ref. 23) and 4800 mil^2 (80×60 mil) (Ref. 24). Therefore, from points 2 and 3 on the curves of Figs. 9 and 10:

Number of complete chips per 2-in wafer $\begin{cases} N_{23} = 320 \\ N_{24} = 430 \end{cases}$

Wafer processing yield (assuming 1975 curve) $\begin{cases} Y_{23} = 54\% \\ Y_{24} = 63\% \end{cases}$

Hence from equation (12) of the Appendix:

Average number of usable chips per 2-in wafer $\begin{cases} n_{23} = 173 \\ n_{24} = 271 \end{cases}$

and from equation (11) of the Appendix:

Preparation costs of the m.o.s. c.a.m. chips $\begin{cases} (C_P)_{23} = £0{\cdot}09 \\ (C_P)_{24} = £0{\cdot}06 \end{cases}$ (1)

For open-chip assembly, such as the method shown in Fig. 5, the chip areas of the two m.o.s. c.a.m.s are 11 400 mil^2 (114×100 mil) (Ref. 23) and 14 000 mil^2 (140×100 mil) (Ref. 24). Hence, from points 4 and 5 on the curves of Figs. 9 and 10:

Number of complete chips per 2-in wafer $\begin{cases} (N_O)_{23} = 200 \\ (N_O)_{24} = 156 \end{cases}$

Wafer processing yield (assuming 1975 curve) $\begin{cases} (Y_O)_{23} = 44\% \\ (Y_O)_{24} = 40\% \end{cases}$

Hence from equation (12) of the Appendix:

Average number of usable chips per 2-in wafer $\begin{cases} (n_O)_{23} = 88 \\ (n_O)_{24} = 62 \end{cases}$

and from equation (11) of the Appendix:

Preparation costs of the m.o.s. c.a.m. chips for open-chip assembly $\begin{cases} (C_{PO})_{23} = \ 0{\cdot}18 \\ (C_{PO})_{24} = \ 0{\cdot}26 \end{cases}$ (2)

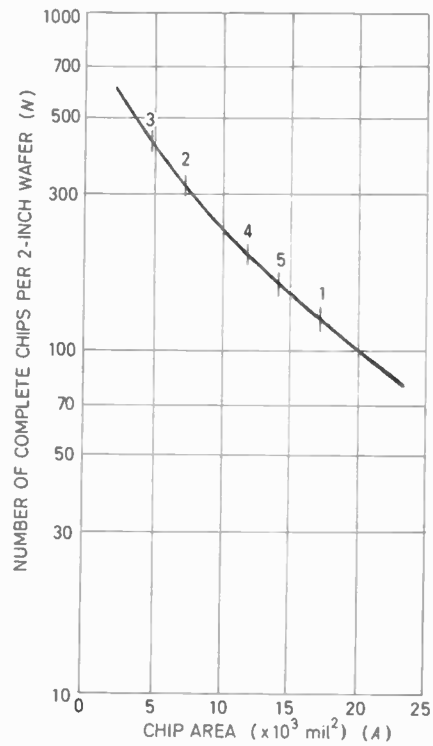

Fig. 10. Typical curve for estimating the number of complete chips per 2-in wafer.

4.2.2 *M.o.s. c.a.m. chip assembly cost (C_A)*

The pin-count for the two m.o.s. c.a.m. designs is 36, if encapsulation in cheap plastic dual-in-line packs is to be employed. Hence from the Appendix, the chip assembly cost (C_A), to a typical manufacturer, would be:

$$(C_A)_{23} = (C_A)_{24} = \ 0{\cdot}30 \tag{3}$$

By further applying the arguments of the Appendix the open-chip assembly cost $(C_{AO})_{23}$ can be estimated to be:

$$(C_{AO})_{23} = (C_{AO})_{24} = \ 0{\cdot}03 \tag{4}$$

4.2.3 *M.o.s. c.a.m. final device testing cost (C_T)*

From equations (7) and (8) of the Appendix it can be seen that the estimated final device testing cost (C_T) for the standard 1024-bit m.o.s. r.a.m. is £0·12. The m.o.s. c.a.m. designs are much simpler to test, the c.a.m. arrays are smaller and are not decoded on-chip. Assuming a *pro-rata* cost reduction the final device testing costs (C_T) for the m.o.s. c.a.m.s are estimated to be:

$$(C_T)_{23} = (C_T)_{24} = \ 0{\cdot}02 \tag{5}$$

and for open chip assembly

$$\begin{aligned} (C_{TO})_{23} &= £0{\cdot}03 \\ (C_{TO})_{24} &= £0{\cdot}06 \end{aligned} \tag{6}$$

4.2.4 *Summary of results*

The 'works-cost-price' for the production of the two m.o.s. c.a.m.s (encapsulated in 36-pin plastic dual-in-line packs) can be estimated, from equations (1), (3), (5) and equations (10), (13), (14) and (17) of the Appendix. These results are shown in Table 2.

It can be seen from Table 2 that the 'works-cost-prices' of the 128-bit m.o.s. c.a.m.s are 65% and 60%

Table 2. Comparison of estimated costs (assuming plastic dual-in-line packaging)

M.o.s. l.s.i. memory	Chip size (mil)	Pin-count	C_P	C_A	C_T	WCP
1024-bit m.o.s. r.a.m.	140 × 120	18	£0·36	£0·15	£0·12	£0·63
128-bit m.o.s. c.a.m.[23]	114 × 63	36	£0·09	£0·30	£0·02	£0·41
128-bit m.o.s. c.a.m.[24]	80 × 60	35	£0·06	£0·30	£0·02	£0·38

(for references 23 and 24 respectively) of that of a typical standard 1024-bit m.o.s. dynamic r.a.m., if plastic dual-in-line packing is employed. Put another way the 'works-costs-per-bit' (WCB) of the m.o.s. c.a.m.s are given by the expressions:

$$\text{Works-cost-per-bit} \quad \begin{matrix} (WCB)_{23} = 5{\cdot}2\,(WCB)_S \\ (WCB)_{24} = 4{\cdot}8\,(WCB)_S \end{matrix} \qquad (8)$$

It can be clearly seen from Table 2 that the chip assembly cost (C_A) dominates the 'works-cost-price' (WCP) for both m.o.s. c.a.m. devices. Consequently the use of open-chip assembly techniques, as shown in Fig. 5 would lead to a much reduced cost-per-bit. The 'works-cost-price' (WCP) for the production of the two m.o.s. c.a.m.s for open-chip assembly can be estimated from equations (2), (4), (6) and equations (10), (13), (15), (16) and (17) of the Appendix. These results are shown in Table 3.

It can be seen from Table 3 that the 'works-cost-prices' of the m.o.s. c.a.m.s are 48% and 70% (for references 23 and 24 respectively) of that of the typical standard 1024-bit m.o.s. dynamic r.a.m., if open-chip assembly is employed. Put another way, the 'works-costs-per-bit' (WCB) of the m.o.s. c.a.m.s are given by the expressions:

$$\text{Works-cost-per-bit} \quad \begin{matrix} (WCB_O)_{23} = 1{\cdot}9\,(WCB_O)_S \\ (WCB_O)_{24} = 1{\cdot}4\,(WCB_O)_S \end{matrix} \qquad (9)$$

In summary, equations (8) and (9) suggest that the costs-per-bit of the m.o.s. c.a.m. designs proposed in references 23 (see Fig. 7) and 24 (see Fig. 8) are 5·2× and 4·8× higher than that of a standard 1024-bit m.o.s. dynamic r.a.m., if the c.a.m.s are encapsulated in a plastic dual-in-line pack. However, if an open-chip or beam-lead (see Fig. 5) technique, is used, the costs-per-bit of the m.o.s. c.a.m.s fall to 1·9× and 1·4× that of the m.o.s. r.a.m.

Of course these figures are for the production costs of the m.o.s. c.a.m. devices *after* amortization of their development costs, which are estimated to be about £20k.

To provide a different perspective on the costing of m.o.s. c.a.m.s the cost-estimation model was applied again, using a standard 4096-bit n-channel m.o.s. dynamic r.a.m. as reference and yield curves similar to those of Fig. 9. The results predict costs-per-bit of 9·0× and 8·3× (for the m.o.s. c.a.m.s of references 23 and 24 respectively) higher than that of the n-channel m.o.s. r.a.m., if plastic dual-in-line packaging is employed. For open-chip assembly these costs-per-bit

Table 3. Comparison of estimated costs (assuming open-chip assembly)

M.o.s. l.s.i. memory	Chip size (mil)	Pin-count	C_{PO}	C_{AO}	C_{TO}	WCP_O
1024-bit m.o.s. r.a.m.	140 × 120	18	£0·36	£0·02	£0·12	£0·50
256- bit m.o.s. c.a.m.[23]	114 × 100	52	£0·18	£0·03	£0·03	£0·24
512-bit m.o.s. c.a.m.[24]	140 × 100	67	£0·26	£0·03	£0·06	£0·35

would fall to 3·0× and 2·2× that of the m.o.s. r.a.m.

It has been pointed out that the cell circuit designs of the two m.o.s. c.a.m.s[23,24] are already near optimum. Conversion to n-channel m.o.s. technology would cause a small improvement in speed and a significant improvement in interface compatibility. However, this change could not match the near 50% reduction in the cost of m.o.s. r.a.m.s caused by the change to n-channel m.o.s. technology. Since this was mainly due to a reduction in cell design complexity. Applying the cost-estimation model to n-channel versions of the m.o.s. c.a.m. devices predicted costs-per-bit of 10·4× and 8·9× (for references 23 and 24 respectively) higher than that of a standard 4096-bit n-channel m.o.s. dynamic r.a.m., if plastic dual-in-line packaging is employed. For open-chip assembly these costs-per-bit would fall to 5·1× and 3·9× that of the m.o.s. r.a.m.

The above results for the different costings of the m.o.s. c.a.m. devices are summarized in Table 4.

5 Conclusions

This paper set out to examine the comparative cost of high-speed associative memory when implemented with m.o.s. l.s.i. c.a.m. devices. A comprehensive survey of m.o.s. l.s.i. c.a.m. development has shown that some considerable experience has already been gained in this field. Two m.o.s. c.a.m. designs, suitable for application as associative memory building-blocks, were selected for cost estimation purposes. The results of this exercise, shown in Table 4, indicate that an associative memory implemented with m.o.s. c.a.m.s would be an expensive alternative to a conventional memory, built from currently available m.o.s. r.a.m. devices.

Table 4. Summary of the estimated costs-per-bit of the two m.o.s. c.a.m. devices[23,24] with reference to those of two standard m.o.s. r.a.m. devices

M.o.s. c.a.m. fabrication technology	Assembly technique	1024-bit m.o.s. r.a.m. (p-channel) Ref. 23	1024-bit m.o.s. r.a.m. (p-channel) Ref. 24	4096-bit m.o.s. r.a.m. (n-channel) Ref. 23	4096-bit m.o.s. r.a.m. (n-channel) Ref. 24
p-channel	Plastic d.i.l. pack	5·2×	4·8×	9·0×	8·3×
	Open-chip assembly	1·9×	1·4×	3·0×	2·2×
n-channel	Plastic d.i.l. pack		—	10·4×	8·9×
	Open-chip assembly	—	—	5·1×	3·9×

Of course, the benefits of higher processing speeds and simpler software implications provided by the associative memory must be taken into account when assessing its cost. Current research into associative processing[4] indicates that the associative memory provides advantages in:

(i) speed: 10–1000 times faster

(ii) storage requirement: 1/10–1/2 times less

when compared with a conventional computer (based on location-addressed r.a.m. storage) over a wide range of applications.

Hence, referring to Table 4, it can be seen that, with plastic d.i.l. packaging, an associative memory would be cheaper than r.a.m. storage for some applications. However, with open-chip assembly, an associative memory would be cheaper than conventional storage implemented with 1024-bit m.o.s. dynamic r.a.m.s for *all* applications and conventional storage implemented with 4096-bit m.o.s. dynamic r.a.m.s for nearly all applications. Consequently, m.o.s. c.a.m.s cannot be regarded as expensive building-blocks for associative memory implementation, especially in view of the already low, and falling, costs of the above-mentioned r.a.m. devices.

It can also be inferred from Table 4 that unless n-channel m.o.s. technology is required for ease of interfacing that p-channel m.o.s. c.a.m. devices are eminently preferable.

It could be argued from the curves of Fig. 9 that at least five years of high-volume production would be required to achieve this order of production cost. However, the fabrication technology required for the m.o.s. c.a.m.s is identical to that used for the 1024-bit m.o.s. dynamic r.a.m. (namely, the p-channel, silicon-gate, m.o.s. fabrication process) and so the m.o.s. c.a.m. devices would benefit from an already optimized fabrication process. Moreover, the estimated development cost of £20k includes circuit and layout improvements. Consequently the m.o.s. c.a.m. designs[23,24] can achieve the reported costs-per-bit without a protracted period of high-volume production.

6 Acknowledgment

The author wishes to acknowledge the many helpful conversations held with colleagues in the semiconductor manufacturing industry concerning the costing of l.s.i. devices. In addition, the author wishes to express his gratitude to GEC–Marconi Ltd. for permission to publish the photographs shown in Figs. 2, 4, 5 and 6.

The two m.o.s. l.s.i. c.a.m. designs[23,24] have been patented in collaboration with the National Research Development Corporation.

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8 Appendix: Cost analysis for a standard 1024-bit p-channel (silicon-gate) m.o.s. dynamic r.a.m.

The 'works-cost-price' of a standard l.s.i. device $(WCP)_S$ can be estimated from the following expression:

$$(WCP)_S = (C_P)_S + (C_A)_S + (C_T)_S \qquad (10)$$

where C_P = chip preparation cost,
C_A = chip assembly and packaging cost,
C_T = final device testing cost.

8.1 Chip Preparation Cost (C_P)

The cost (C_W) of a 2-in silicon wafer is approximately £2·00, at the time of writing. The average cost-per-processed wafer (C_{PW}) is of the order of £14 for a 2-in silicon water. Hence the cost ($(C_P)_S$) of the preparation of a standard l.s.i. chip can be estimated from the expression

Preparation cost of a standard l.s.i. chip

$$= (C_P)_S = \frac{1}{n}(C_W + C_{PW}) = \frac{£16}{n} \qquad (11)$$

where n = average number of usable chips per wafer.

The value for n for a particular l.s.i. chip can be calculated from the expression

$$n = NY \qquad (12)$$

where N = number of complete chips per 2-in wafer,
Y = percentage yield of usable chips.

The number (N) of complete chips per wafer depends on the chip area (A), as shown in Fig. 10. This curve is an estimate, for a more accurate value the exact dimensions of the chip, spacing between chips and the orientation of processing masks relative to the wafer must be known. Much research has been devoted to the problem of low yields for l.s.i. chips and various different mathematical models for yield prediction exist.[25–28] The curves shown in Fig. 9 for the silicon-gate p-channel m.o.s. fabrication process are derived from a simple consensus of these models, tempered with some practical experience of semiconductor industry. The curves indicate the extent to which yield decreases with increasing chip area (due to process deficiencies and operator errors) and increases with time (due to successive improvements in the fabrication process and optimization of the chip circuit and layout designs). Although some of the reasons for the loss of yield are understood, it remains a fickle parameter which can be contained but not controlled. Consequently the curves of Fig. 9 must be regarded as an estimate for a typical manufacturer rather than a precise indication.

The chip area of the typical 1024-bit p-channel (silicon-gate) m.o.s. dynamic r.a.m. is 16 800 mil^2 (120×140 mil). Therefore, from the curves (point 1) of Figs. 9 and 10:

Number of complete chips per 2-in wafer = N = 119

Wafer processing yield (assuming 5 years production) = Y = 37%

Hence, from equation (12)

Average number of usable chips per 2-in wafer = n = 44

and from equation (11)

Preparation cost of the 1024-bit m.o.s. r.a.m. = $(C_P)_S$ = £0·36 (13)

8.2 Chip Assembly Cost (C_A)

The cost of l.s.i. chip assembly (C_A) is mainly dependent on the number of chip terminals. M.o.s. l.s.i. memory chips are currently being encapsulated in cheap plastic dual-in-line packs. Many different 'standard' packs are available; common pin-counts are 14, 16, 18, 22, 24, 28, 36 and 40. The chip assembly cost (C_A) for a particular pin-count will vary somewhat between manufacturers, depending on the skill and efficiency of its l.s.i. assembly and packaging line. Typical chip assembly costs for the low pin-counts can be much less than 10p per pack. However, for the higher pin-counts terminal bonding becomes intricate and therefore more expensive. Hence for l.s.i. memory devices it is more realistic to estimate the chip assembly cost (C_A) from the range £0·12–£0·35 per pack, depending on the pin-count.

The pin-count of the standard 1024-bit m.o.s. r.a.m. is 18. Hence the chip assembly cost (C_A), to a typical manufacturer, for this device would be:

$$(C_A)_S = £0{\cdot}15 \qquad (14)$$

More encouraging is the current development of semi-automatic open-chip assembly methods, such as the flip-chip and beam-lead techniques. An example of the latter method is shown in Fig. 5. With these techniques the chip terminals are bonded simultaneously or in rapid sequence by a machine tool. Reductions in chip assembly cost (C_A) of an order of magnitude can be achieved in this way. Hence for the standard 1024-bit m.o.s. r.a.m. the open-chip assembly cost $(C_{AO})_S$ could be estimated to be:

$$(C_{AO})_S = 0{\cdot}02 \qquad (15)$$

8.3 Final Device Testing (C_T)

The cost of final device testing (C_T) can be expensive in cases of high circuit complexity. This cost is normally depreciated by making extensive use of computerized testing schedules for very large numbers of the same i.c. devices. For regular two-dimensional arrays, such as most memory circuits, the final testing cost (C_T) is likely to be in the range of 10%–30% of the total device cost (WCP). Hence for the standard 1024-bit m.o.s. r.a.m. a feasible assumption for final device testing cost (C_T) would be:

$$(C_T)_S = £0{\cdot}20(WCP)_S \qquad (16)$$

8.4 Conclusion

The 'works-cost-price' of the standard 1024-bit p-channel (silicon-gate) m.o.s. dynamic r.a.m. can be estimated from equations (10), (13), (15) and (16) as follows:

$$(WCP)_S = £0{\cdot}36 + £0{\cdot}15 + £0{\cdot}20\,(WCP)_S$$

Therefore, for an 18-pin plastic dual-in-line pack

$$(WCP)_S = £0{\cdot}63 \qquad (17)$$

Hence, the cost-per-bit of this memory, to a typical l.s.i. device manufacturer, would be of the order of:

$$\text{Works-cost-per-bit} = (WCB)_S = \frac{£0{\cdot}63}{1024}$$

$$= 6{\cdot}15 \times 10^{-2}\ \text{p} \quad (18)$$

If an open-chip assembly technique were employed the 'works-cost-price' could be estimated from equations (1), (4), (6), (7) and (8) as follows:

$$(WCP_O)_S = £0{\cdot}36 + £0{\cdot}02 + £0{\cdot}20\,(0{\cdot}63)$$

Therefore, for open-chip assembly

$$(WCP_O)_S = £0{\cdot}50 \quad (19)$$

In this case the cost-per-bit of the memory to a typical l.s.i. device manufacturer would be of the order of:

$$(WCB_O)_S = \frac{£0{\cdot}50}{1024} 4{\cdot}88 \times 10^{-2}\ \text{p} \quad (20)$$

It must be stressed that all the above-mentioned costs are *estimates* for a typical semiconductor manufacturer, at the time of writing.

Manuscript first received by the Institution on 4th December 1975 and in final form on 1st March 1976. (Paper No. 1736/Comp. 176.)

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UDC 621.398:621.391.883.2

Error-rate in a digital simplex telemetry system

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SUMMARY

This paper describes the digital part of a simplex binary telemetry system for conveying control signals to some remote body such as a missile. The decoder in the remote body is small, light, has a low power consumption, and is designed to operate above a certain noise threshold. When the signal-to-noise ratio of the signal at the input to the decoder is low the latter only accepts information which has a high probability of being correct thereby causing the rejection of most of the transmitted information. When information is rejected the remote body functions on the data which were last accepted by the decoder.

Both experimental and theoretical results are presented for error rate and received frame rate, and the effect of asymmetry in the binary channel is examined.

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List of Symbols

C	channel capacity of data channel.
d	suffix which denotes that the condition of channel d.c. drift prevails.
D	value of channel d.c. drift in volts.
$E.R.$	error rate.
f_b	transmitted bit-rate.
F_t	frame transmission rate.
F_r	received transmission rate.
H	probability of making an error in a frame bit when the channel is symmetrical.
H_0	probability of a logical zero being changed to a logical one due to channel d.c. drift.
H_1	probability of a logical one being changed to a logical zero due to channel d.c. drift.
H_d	probability of making an error due to channel d.c. drift.
K	ratio of channel d.c. drift to V_0, expressed as a percentage.
m	bit length of synchronization code.
m_r	bit length of reversal code.
n_f	bit length of one frame.
n	number of bits in a data word.
N	number of data words in the frame.
P_d	probability of no error in a data word.
P_s	probability that the synchronization code is correct.
P_e	probability of an error in a data word when the synchronization is correct.
v	instantaneous value of the band-limited white noise voltage at the decoder input.
V_0	peak-to-peak value of the binary signal at the decoder input in the absence of noise.
σ	r.m.s. value of band-limited white noise at the decoder input.
μ	subscript relating to a data word.

1 Introduction

The control of a remote body, such as a missile, can be achieved by the simplex transmission of the control signals as a time division multiplex binary signal on a suitable carrier, provided that the decoding system in the remote body will only accept data whose probability of being in error is small. One method of achieving a low error-rate is to use error detection and correction techniques. However, the topic of this paper is to describe an alternative method which is particularly suitable when the remote body is subjected to the constraints of size, weight and power consumption.

Consider a baseband digital system which transmits data at a rate substantially in excess of the Nyquist rate. In the presence of a burst of noise, the receiver in the remote body rejects data which it considers to have a high probability of error. In this way the system error-rate is maintained at a low level at the expense of received channel capacity. Thus in the absence of reliable information the receiver operates on previously accepted data which have a high probability of being correct.

The multiplex data are composed of frames. Each frame contains N data channels and a synchronization code. This code has two functions: the first is to identify the time location of each data channel, and the second is to assist the receiver in determining whether the frame data should be accepted or rejected. If the code is correctly identified the frame data are accepted, but if the code is in error the remote body operates on the frame data contained in the last frame whose synchronization code was correctly received. The synchronization code bits are asymmetrically distributed throughout the frame, and the data have a negligible probability of imitating this code. A refinement to this method of using the correct identification of the synchronization code as the criterion of accepting frame data is described in Section 7 and results in a profound reduction in the error rate.

In the next two Sections a description of the baseband systems at the transmitter and receiver is presented. This is followed by an analysis of error-rate and channel capacity for both symmetrical and asymmetrical non-return-to-zero binary data waveforms. It will then be shown that the theoretical equations are in good agreement with the experimental results.

We emphasize that the terminal equipments and transmission channel are *not* described.

2 The Encoder

The output of the encoder consists of non-return-to-zero binary data having a bit rate of f_b bits per second. These binary data are divided into identical frames. Each frame consists of a synchronization code which is m bits in length, and N data channels having different word lengths. Some of the information channels convey binary data, while other channels contain analogue signals which have been pulse code modulated by the encoder.

Figure 1 is the block diagram. The analogue inputs are connected to sample-and-hold circuits. When a particular analogue input is being encoded the output of its hold circuit is connected to one input of a comparator while the other input is derived from a binary weighted resistor ladder connected to a constant reference voltage, which produces discrete voltage steps switched by commands from a ring counter. The voltage steps are successively summed at the comparator input until the nearest approximation to the input is reached (i.e. $\pm$ the smallest voltage step). The final code indicates the presence or lack of each voltage step in the voltage level which is to be represented. The analogue inputs are alternatively switched to the input of the comparator by commands from the ring counter.

To obtain stability a high frequency crystal oscillator is divided to give a clock of the desired bit repetition rate. The clock is fed to the synchronization code generator which produces the m_r bit reversal code and clocks it through the output gate. This reversal code which forms part of the synchronization code consists of an alternate 01 pattern which is generated in order that the clock in the decoder can be easily locked on to the incoming bit stream.

After the reversal code is generated, the clock is switched to a $(n_f - m_r)$ bit ring counter, where n_f is the number of bits in one frame. The ring counter gates the bits of information from the analogue inputs and the digital inputs through the output gate in correct sequence and inserts the remaining synchronization code bits.

These remaining bits of the synchronization code are introduced from the information gating and are distributed in ones or twos among the data bits to ensure that the probability of data bits imitating the synchronization code is negligible.

A $(n_f - m_r)$ bit counter produces a single pulse when $(n_f - m_r)$ clock pulses have been applied to the ring counter. This single pulse, called the frame reset pulse, resets the ring counter and switches the clock back to

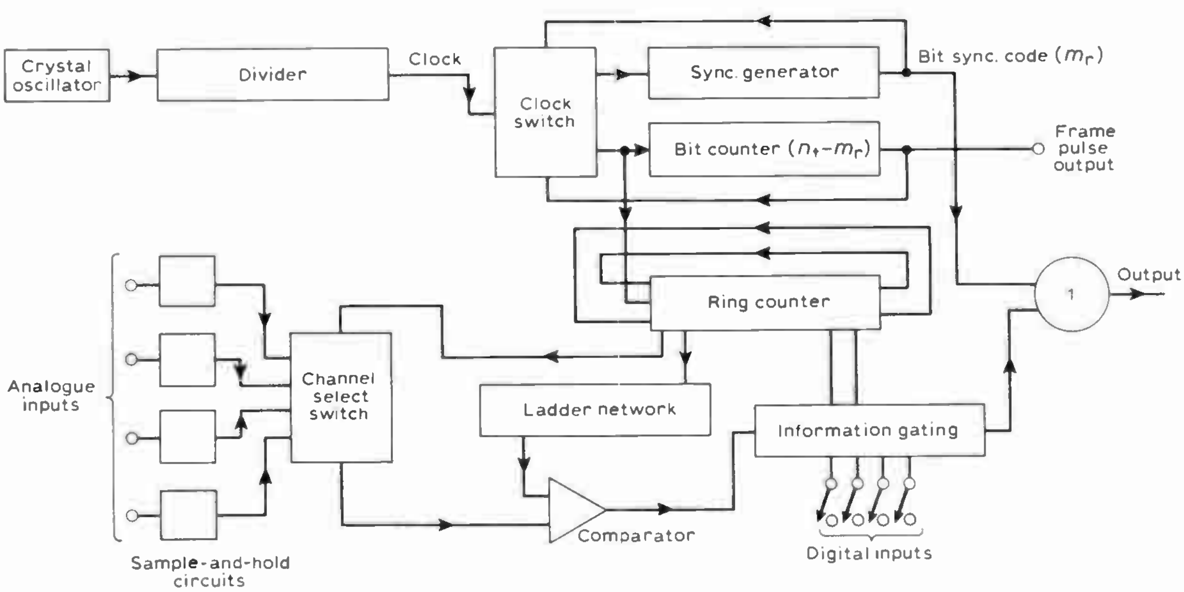

Fig. 1. Encoder block diagram.

the synchronization code generator to commence the next frame.

3 The Decoding Process

The receiver system demodulates the incoming signal to produce the original t.d.m. signal together with unwanted noise. The non-return-to-zero binary waveform is passed into a zero crossing comparator which changes the t.d.m. signal into the logic levels employed in the decoder. The binary waveform at the output of the comparator has more binary changes than the waveform which left the encoder due to the presence of noise. The changes due to noise only occur when the modulus of the noise voltage is in excess of $V_0/2$ where V_0 is the peak-to-peak value of the signal at the comparator input in the absence of noise.

To synchronize the clock in the decoder to the incoming bit stream the system employed is shown in Fig. 2. The output signal from a stable crystal oscillator is divided by divider D1 to form a square wave CLK1. The synchronization system adjusts the phase of CLK1 so that after further division by divider D2 a clock waveform CLK2 will be generated which is almost synchronous with the incoming bit stream. In order to accommodate the phase differences between the transmitted and received bit streams at the decoder in the presence of noise, and to produce a synchronized clock CLK2, the synchronization system differentiates V_c and produces a waveform V_D which has nagetive pulses representing the positions of changes in the binary levels in the V_c waveform. Clock CLK2 allows these negative pulses to be counted by counter A when CLK2 is positive and to be counted by counter B when CLK2 is at zero volts. Both counters A and B are able to count up to four. When either of the counters has reached a number 4, the information is passed to a differentiating circuit, and then into a phase-shift network.

The phase-shift network is basically a synchronous divide-by-4 circuit and is displayed in Fig. 3. Figure 4 shows how the waveform Q_2 is advanced or delayed.

When counter A has reached 4 the preset pulse P occurs thereby causing the output of flip-flop FF1, namely Q_1, to change to a logic 1, resulting in the output

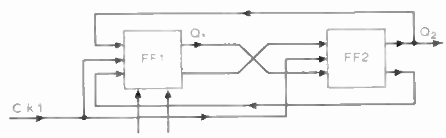

Fig. 3. Phase-shift network.

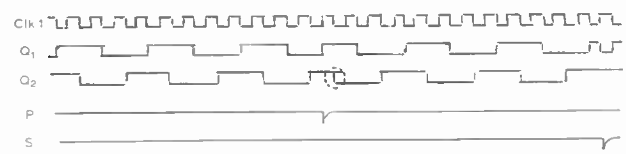

Fig. 4. Waveform associated with phase-shift network.

of FF2 prematurely returning to logic 0 which implies that the negative edge of the current Q_2 pulse is advanced.

When counter B has counted 4 pulses pulse 5 is produced causing Q_1 to go to logic 0 and causing the negative edge of the current Q_2 pulse to be delayed. Thus the negative-going edges in the Q_2 waveform will be adjusted where necessary in a continuous attempt to make counter A and B receive the same number of pulses over a given period, thereby making the clock CLK2 synchronous with the incoming bit-rate. The synchronization system performs satisfactorily in the presence of noise provided the extra zero crossings of the comparator output due to the noise voltage do not occur such that one counter counts more pulses then the other counter.

The synchronized clock CLK2 is then used to clock the information into the decoder shift register. The decoder shift register has a length of n_f bits, i.e. one frame length. The m_r bits which comprise the reversal code, together with the other bits which make up the m bit synchronization code, occupy known positions in the frame and these positions in the shift register are connected to the transfer pulse generator (see Fig. 5).

When the synchronization code (m_r bits) is correctly recognized a transfer pulse is generated which transfers the information in the data channels into the parallel store, i.e. the data in the parallel stores are only updated

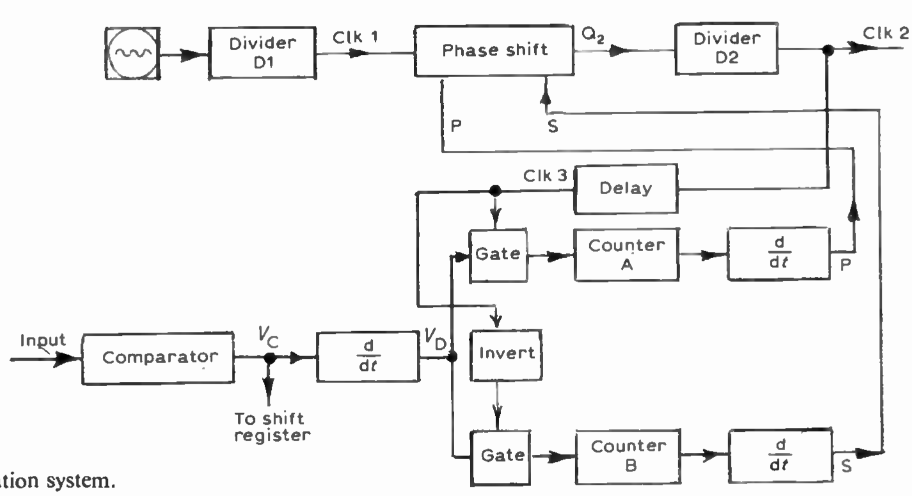

Fig. 2. Bit synchronization system.

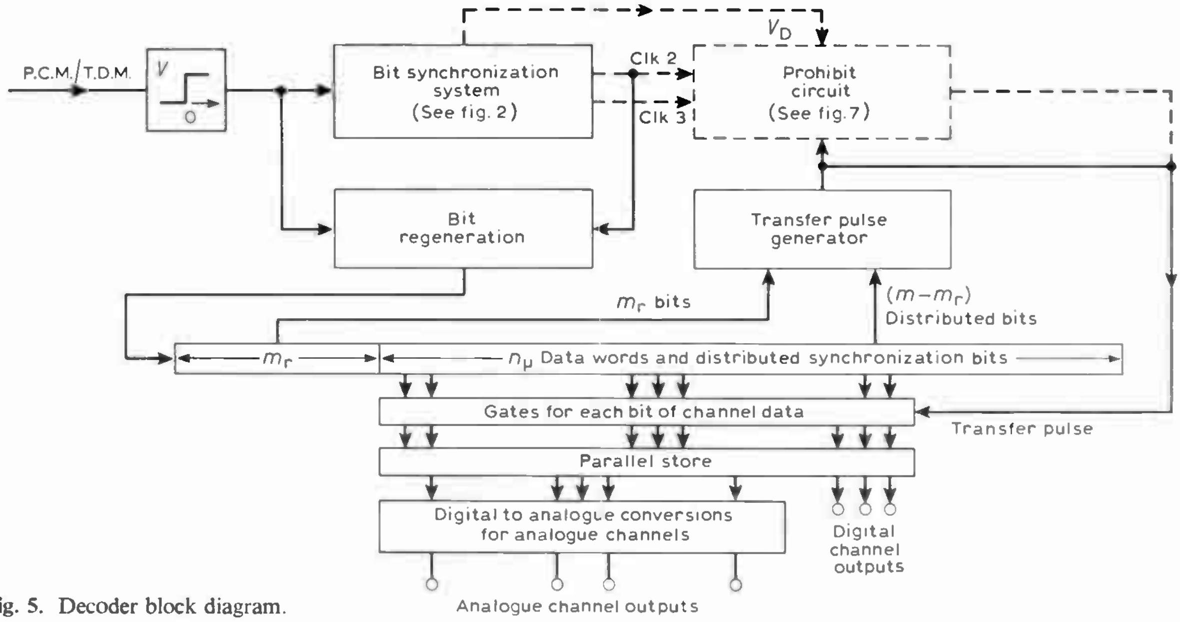

Fig. 5. Decoder block diagram.

when the transfer pulse is generated. Given that bandwidth is not a premium then the sampling rate at the encoder can be made substantially greater than the Nyquist rate enabling some data rejection to occur in order for the data which are accepted to have a high probability of being correct, and yet causing little degradation to the decoded analogue channels.

When the signal-to-noise ratio at the input to the decoder is low a large amount of data is rejected and degradation of the decoded analogue signal may be considerable due to working on reliable data which may not have been updated for a number of frames. However, in this system it is better to have degradation of the analogue channels due to working below the Nyquist rate, than to accept data which have a high probability of being erroneous.

4 Error Rate

The received t.d.m. signal in the absence of noise has non-return-to-zero binary levels of $\pm V_0/2$. In the presence of additive band-limited white noise, zero mean, variance σ^2, the optimum decision level of the received t.d.m. signal is zero volts. This noise-contaminated t.d.m. signal is applied to a zero crossing detector, i.e. a comparator, and if at the times when the output of the comparator is inpected (these occur at approximately the centre of the periods) the input signal to the comparator is $\geqslant 0$, a logical 1 $(+V_0)$ is generated, otherwise a logical zero (zero volts) is formed.

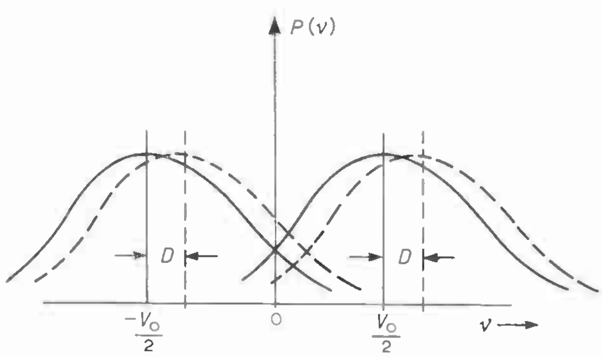

Fig. 6. Probability density curves for symmetrical and asymmetrical channels.

In Fig. 6 the solid curves are the probability density functions for the noise signal relative to the binary levels of the input signal. The probability H of the noise causing the binary levels of $\mp V_0/2$ to be interpreted by the comparator as $\pm V_0/2$ respectively is well known[1] and is

$$H = \frac{1}{2}\left\{1 - \mathrm{erf}\left(\frac{V_0}{\sigma 8^{\frac{1}{2}}}\right)\right\} \tag{1}$$

where

$$\mathrm{erf}\,(x) = \frac{2}{\sqrt{\pi}}\int_0^x \exp(-y^2)\,\mathrm{d}y.$$

Over a large interval of time the average power of the t.d.m. waveform after it has been regenerated is $V_0^2/2$, provided that the noise has not introduced any errors.

The long-term signal-to-noise ratio S/N is

$$\frac{S}{N} = \frac{V_0^2}{2\sigma^2}. \tag{2}$$

Substitute equation (2) into equation (1) gives the probability of a binary error in terms of S/N.

$$H = \frac{1}{2}\left[1 - \mathrm{erf}\left\{\frac{1}{2}\left(\frac{S}{N}\right)^{\frac{1}{2}}\right\}\right]$$

or

$$H = \tfrac{1}{2}\,\mathrm{erfc}\left\{\frac{1}{2}\left(\frac{S}{N}\right)^{\frac{1}{2}}\right\}. \tag{3}$$

The probability of no error in a code bit is $(1-H)$, hence the probability of no error in n digits, assuming that the probability of any bit being in error is statistically independent of other bits being in error, is

$$P_d = (1-H)^n. \tag{4}$$

The probability that the m bit synchronization code is completely correct is

$$P_S = (1-H)^m. \tag{5}$$

Consider the kth frame. Suppose the synchronization code is not correctly recognized in this frame, and that during the next r frames the synchronization code remains unrecognizable. In the $(k+r+1)$ frame the synchronization code is correctly received. The decoder rejects the channel data associated with frames k to $k+r$. These data are only accepted when the synchronization code is correctly received.

However, when the synchronization code is correct the channel data may be false. The probability of this situation occurring is

$$P_e = P_s(1-P_d). \tag{6}$$

From equations (4), (5) and (6):

$$P_e = (1-H)^m[1-(1-H)^n]. \tag{7}$$

If the frame transmission rate is F_t then the received transmission rate F_r, i.e. the rate at which the synchronization code was correctly recognized, is $P_s F_t$, and from equations (3) and (5) this gives

$$F_r = \left[1 - \tfrac{1}{2}\,\mathrm{erfc}\left\{\frac{1}{2}\left(\frac{S}{N}\right)^{\frac{1}{2}}\right\}\right]^m F_t. \tag{8}$$

Now P_e gives the probability of the decoder correctly recognizing the synchronization code when the channel data are in error. This can be expressed as the number of frames for which this condition occurs divided by the number of frames transmitted over a given period, or it can be expressed in a form of error-rate $E.R.$ as

$$P_e = \frac{E.R.}{F_t}. \tag{9}$$

In order to express error-rate of a particular data channel, say μ, in the t.d.m. frame as a function of S/N at the input to the regenerator, equations (3) and (7) are substituted into equation (9) to yield

$$E.R._\mu = \left\{1 - \mathrm{erfc}\left(\frac{1}{2}\left(\frac{S}{N}\right)^{\frac{1}{2}}\right)\right\}^m \times$$

$$\times \left[1 - \left\{1 - \tfrac{1}{2}\,\mathrm{erfc}\left(\frac{1}{2}\left(\frac{S}{N}\right)^{\frac{1}{2}}\right)\right\}^{n_\mu}\right] F_t. \tag{10}$$

Now a frame consists of a synchronization code of m bits and N data channels containing words of length $n_1, n_2, \ldots, n_m$. If the duration of one bit period is $(1/f_b)$ the frame rate is

$$F_t = \frac{f_b}{m_f} = \frac{f_b}{m + \sum_{\mu=1}^{N} n_\mu}. \tag{11}$$

4.1 Effect of Asymmetry in the Non-return-to-zero Binary Waveform

Suppose that in the absence of noise the t.d.m. signal applied to the comparator input has voltage levels $(V_0/2) + D$ for the presence of a code pulse and $-(V_0/2) + D$ for the absence of a code pulse, i.e. the communication channel has a positive drift of D volts. The probability density functions for this situation are given by the dotted curves in Fig. 6.

When there is no pulse present, the noise level must exceed $(V_0/2) - D$ to cause an erroneous code pulse to be formed. The probability H_0 of this occurring is half the probability that the Gaussian noise voltage exceeds $(V_0/2) - D$. Similarly, the probability H_1 of a negative noise voltage causing a code pulse (logical one) to be detected as a logical zero is half the probability that the noise voltage is more negative than $(V_0/2) + D$. The probability H_d of either type of error occurring is

$$H_d = H_0 + H_1 = \frac{1}{4}\left\{\mathrm{erfc}\left(\frac{V_0}{\sigma(8)^{\frac{1}{2}}} - \frac{D}{\sigma(\frac{1}{2})^{\frac{1}{2}}}\right) + \right.$$

$$\left. + \,\mathrm{erfc}\left(\frac{V_0}{\sigma(8)^{\frac{1}{2}}} + \frac{D}{\sigma(2)^{\frac{1}{2}}}\right)\right\}. \tag{12}$$

Let the effect of d.c. drift be expressed as a percentage of the peak-to-peak signal amplitude.

$$K = (D/V_0) \times 100. \tag{13}$$

Then equation (12) can be written with the aid of equation (2) and (13) as

$$H_d = \frac{1}{4}\left[\mathrm{erfc}\left(\frac{1}{2}\left(\frac{S}{N}\right)^{\frac{1}{2}} - \frac{K}{100}\left(\frac{S}{N}\right)^{\frac{1}{2}}\right\} + \right.$$

$$\left. + \,\mathrm{erfc}\left\{\frac{1}{2}\left(\frac{S}{N}\right)^{\frac{1}{2}} + \frac{K}{100}\left(\frac{S}{N}\right)^{\frac{1}{2}}\right\}\right]. \tag{14}$$

The received frame rate is

$$F_{rd} = (1-H_d)^m F_t \tag{15}$$

and the error rate due to d.c. drift in the channel is calculated from equation (6) and (9) as

$$E.R._d = (1-H_d)^m\{1-(1-H_d)^n\}F_t. \tag{16}$$

4.2 Channel Capacity

The channel capacity of a particular channel in the frame which has a word length n_μ bits is

$$C_\mu = F_r n_\mu$$

$$= (1-h)^m f_b(n_\mu/m_f) \tag{17}$$

where h is H or H_d applies for a symmetrical channel (equation (3)) or an asymmetrical channel (equation (12)) respectively. H and H_d have been defined in terms of signal-to-noise ratio, and as this ratio is made very large the probability of channel data having an error P_e approaches zero as h approaches zero. The system now works at its maximum capacity as F_r approaches F_t. The graph of error rate as a function of signal-to-noise ratio at the input to the decoder has a single peak. At very low S/N the error-rate is very low. This condition follows from the way the error rate has been defined, and although error rate becomes negligible as S/N tends to unity so does the channel capacity. This can be observed from Fig. 10, as received frame rate is directly proportional to channel capacity.

5 Reduction of Error-rate

In order to reduce the probability of accepting data which is in error, all of the bits of each frame at the output of the zero-crossing comparator are inspected for binary changes over a half a bit period symmetrically located about the centre of each bit. During this inspection period no binary change will occur unless a noise spike greater than $V_0/2$ is present. If any of the frame bits do have binary changes during the inspection process then all the frame data are rejected, even if the transfer pulse generator correctly recognized the synchronization code.

Clearly this inspection process will greatly increase the data rejected, but will result in data at the output of the decoder having a lower probability of being in error.

This amendment of the decoder is implemented by forming a strobe pulse train CLK3 from the synchronized clock CLK2 (see Fig. 2) having a duration of a half bit period, and applying this strobe train CLK3 to a prohibit circuit shown in Fig. 7, and represented in schematic form by the dotted lines in Fig. 5.

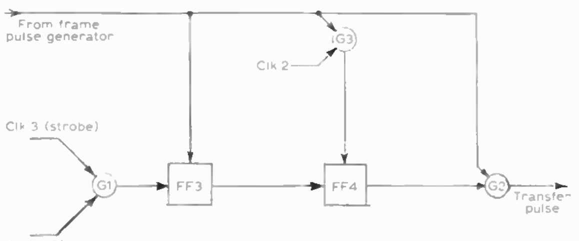

Fig. 7. Prohibit circuit.

The V_D waveform established in the bit synchronization system (see Fig. 2) is applied to the gate G1 which is strobed by CLK3. If *any* of the frame bits have a change of binary levels during the strobe period due to a noise spike then the output of gate G1 will set flip-flop FF3 to a one state. During the frame period, i.e. prior to the end of the frame, gate G3 allows the clock pulses CLK2 to be applied to flip-flop FF4. This flip-flop has an output of logic 1 until a noise spike results in FF3 being set, when the next CLK2 pulse causes FF4 to go to logic 0 thereby inhibiting gate G2. In the meantime the synchronization system is behaving as described previously, i.e. 'instantaneous' sampling at the centre of each bit period. If the transfer pulse generator produces a transfer pulse and gate G2 is inhibited by FF4 then the transfer pulse will be inhibited from transferring frame data to the output of the decoder. Throughout the period of the transfer pulse the gate G2 will be inhibited due to transfer pulse inhibiting G3 and removing the clock from FF4. The frame pulse resets FF3 to its original state, but FF4 does not revert to its original state until the inhibiting effect of the transfer pulse is removed and a subsequent clock CLK2 pulse occurs.

If G2 is inhibited due to a noise spike in one or more of the frame bits during the strobe period, and the transfer pulse generator does not produce a transfer pulse, then not only will frame data be prohibited from passing to the output of the decoder, but bistable FF3 will not be reset. This results in frame data being rejected in the subsequent frame irrespective of whether noise spikes are present or absent in the frame bits during the strobe period, and the situation will continue until a frame pulse is generated thereby resetting FF3.

Noise bursts which completely change a data bit from one to a zero or vice versa during the strobe period will not be detected and constitute an error, providing that the synchronization code is correctly identified. These types of noise bursts are the only types which can result in data errors, and the probability of occurrence is reduced by increasing the system bandwidth well above the minimum required to transmit the data.

6 Measurements

Data of a known and exact cyclic binary pattern whose periodicity is equal to the frame are produced by a pattern generator acting as an encoder simulator. These binary data representing the output of the encoder are connected to a channel simulator whose function is to simulate the non-baseband part of the digital simplex telemetry system. Also connected to the simulator via a variable attenuator is a white Gaussian noise source. The signal emanating from the simulator is a noisy band-limited binary signal and is presented to the input circuitry of the decoder shown in Fig. 5.

Auxiliary circuitry connected to the parallel store in the decoder ascertains if the known transmitted signal has been correctly processed by the pulse reforming circuits. In order to simplify the auxiliary circuitry cyclic binary patterns representing the p.c.m. analogue signals are arranged to have equal numbers of logical ones and zeros, but the positioning of these logical ones and zeros is at the discretion of the experimenter. The auxiliary circuits compare the cyclic binary pattern from the pattern generator with those in the decoder's parallel store and if one or more bits are in error a logical one is produced. If the two sets of data are identical a logical zero appears at the output of the auxiliary circuit.

Because the system is designed so that it prefers to function with the most reliable information that can be obtained, an error in a data channel will not appear at the decoder output unless the frame pulse is generated, i.e. if the synchronization code is correctly received. If the frame pulse is not produced the decoder channel outputs are those in the last frame which did correctly receive the synchronization code. Accordingly a channel error is considered to have occurred only if the output of the auxiliary circuit associated with that channel is a logical one *and* the frame pulse is generated.

Figure 8 shows the experimental arrangement for the measurement of error-rate *E.R.* and received frame-rate F_r. The signal power of the binary signal at the out-

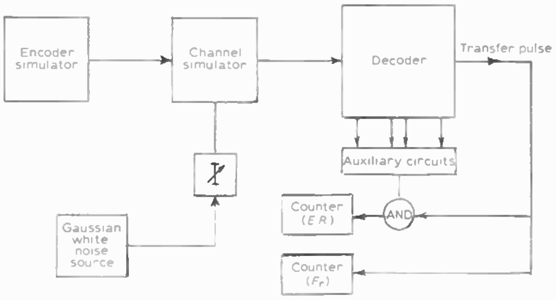

Fig. 8. Experimental arrangement for the measurement of error-rate and received frame rate.

put of the encoder simulator is maintained at a constant value. For a given measurement of *E.R.* and F_r, the noise power is observed at the input of the decoder by an r.m.s. voltmeter. Experiments are then made for different settings of the variable attenuator (placed after the noise source) and for each noise voltage measured the signal-to-noise ratio is calculated. By this method error-rate and received frame-rate are determined as a function of signal-to-noise ratio.

A counter is used to measure *E.R.* in one of the N binary channels, and another counter records F_r. Ultra-violet recorders, not shown in Fig. 8, are also used to determine *E.R.* and F_r. These recorders perform a check on the counters, which may act erroneously in the presence of impulse noise generated from surrounding equipment, and also indicate the positioning of the errors and frame pulses as a function of time. For each value of S/N, 20 readings each lasting over 100 seconds are taken from the recorders and counters. The transmitted frame-rate F_r is unchanged for different signal-to-noise ratios.

The same experimental arrangement as shown in Fig. 8 was employed for the measurements of received frame-rate in the presence of d.c. drift in the channel. This d.c. drift was introduced by adjusting the d.c. bias to a differential amplifier in the channel-simulator thereby making the t.d.m. waveform at the decoder input asymmetrical with respect to earth.

7 Results

Employing the measurement technique described in Section 8 for F_t = 500 frames per second and m = 20, the curves of error-rate as a function of signal-to-noise ratio at the input to the zero-crossing detector are presented in Fig. 9 for a data channel having an 8-bit word. The theoretical calculations relating to this condition which applies to the method of detection of the

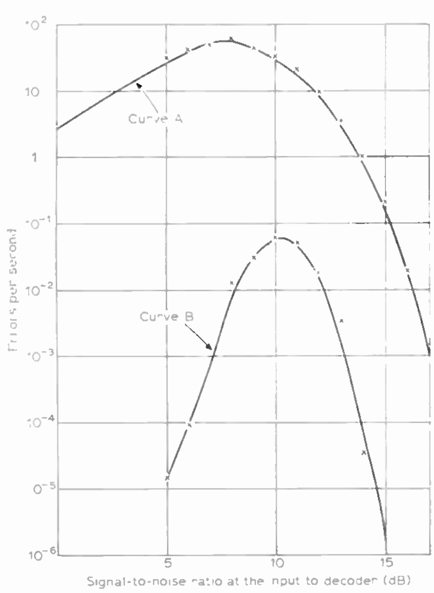

Fig. 9. Error rate against signal-to-noise ratio.
A. Without strobing. B. With strobing.

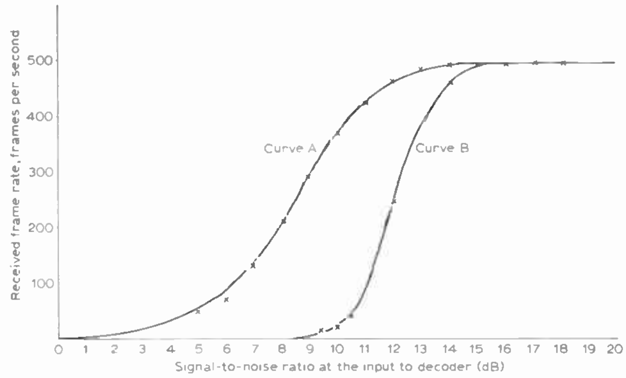

Fig. 10. Received frame rate against signal-to-noise ratio.
A. Without strobing. B. With strobing.

presence or absence of a code pulse by sampling it at the centre of the bit period, i.e. by instantaneous sampling, is represented by curve A. This curve has been computed from equation (10). The crosses in the proximity of curve A relate to experimental results.

When the technique of strobing the code pulses for half a period centred symmetrically about the centre of every pulse in each frame was adopted, the experimental results obtained are displayed by curve B in Fig. 9.

The corresponding variation of received frame-rate as a function of signal-to-noise ratio is displayed in Fig. 10, where curve A again represents instantaneous sampling and is derived from equation (8) and the crosses are the related experimental results. The effect of strobing produces the experimental curve B.

The effect of various values of asymmetry of the t.d.m. waveform at the input to the decoder are shown in Fig. 11. The smooth curves were derived from equation (15), and the crosses are experimental observations.

8 Discussion

The experimental results are displayed in the form of crosses in Figs. 9, 10 and 11, and relate to the same channel simulator. When using the prohibit circuit shown in Fig. 5 the probability of detecting a noise spike during the strobe period is increased if high-frequency noise is allowed to be present. Consequently, the channel simulator had its fixed frequency response suitable for the decoder using strobing, i.e. the response was five

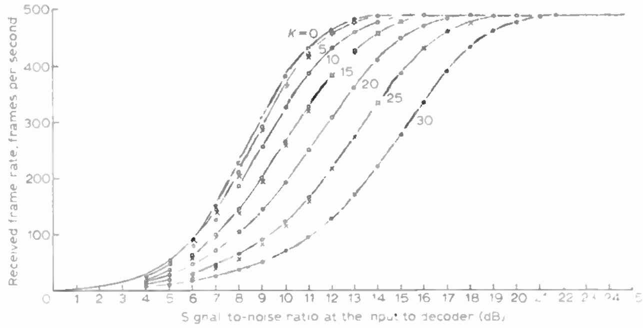

Fig. 11. Received frame rate against signal-to-noise ratio for different values of asymmetry of the t.d.m. signal at the decoder.

times the value required for signalling at the Nyquist rate.

Both theoretical curves in Figs. 9 and 10 contain one peak. For S/N at the input to the decoder which are in excess of $\sim$11 dB, curves A and B in Fig. 9 are approximately parallel. Consequently for the same error-rate the decoder using strobing can operate with $S/N \sim 4$ dB lower than the decoder with instantaneous sampling, i.e. without strobing. However, the penalty of using strobing is the reduction in received frame-rate F_r, see Fig. 10. The guidance system is expected to operate where in general $S/N > 18$ dB. On occasions when $S/N \ll 18$ dB (due to jamming or electrical storms) it is essential that the data accepted have a very high probability of being correct, even if this is achieved at the expense of an accompanying low received channel capacity.

The decoders using instantaneous sampling and strobing have similar performances for $S/N > 18$ dB. However, when $S/N < 10$ dB the error-rate for the decoder using instantaneous sampling is many orders greater than the decoder using strobing. As S/N decreases progressively the error-rate decreases, but this is achieved by less frame data being accepted for decoding.

It is interesting to note that if the channel bandwidth is reduced to half the transmitted bit rate (Nyquist rate) curves A and B in Fig. 9 have similar shapes. However, while curve A is shifted by 7 dB horizontally to the left due to the noise power decreasing as the bandwidth is reduced, curve B experiences this same lateral shift but also moves vertically upwards such that the peak error-rate becomes approximately 4 errors per second. The reason why the peak error-rate increases in this later case is because the low-frequency noise can obliterate or simulate a pulse without the high-frequency noise causing any binary changes during the inspection process. It is the presence of the high-frequency noise which enables the probability of error detection to have a high value.

Now for high S/N the error-rate is reduced if the channel bandwidth is half the Nyquist rate, but as the error-rate for high S/N is satisfactory for the wide channel bandwidth situation this is no advantage. What is required is that the error-rate is as low as possible when the S/N is low. Thus the reason for the channel bandwidth being in excess of its Nyquist value is that in the relatively rare situation when the $S/N < 10$ dB the error-rate is reduced, and so of course is the number of frames accepted for decoding.

The theory, presented in Section 4, relating error-rate and channel capacity in terms of the probability H of making an error in a code bit is applicable to decoders employing either instantaneous sampling or strobing. The value of H has been analytically derived in terms of S/N for the case of instantaneous sampling. The writers could not find a rigorous method of deriving H in terms of S/N when the strobing method (described in Section 5) was employed in the decoder, although others[2–5] approached a related problem and obtained an approximate solution. Accordingly the approach used here was to obtain experimental results as displayed by curve (ii) in Figs. 9 and 10 for the system performance using strobing and by-pass the problem of computing H.

The effect of asymmetry in the non-return-to-zero waveform at the input to the zero-crossing comparator for the decoder employing instantaneous sampling is shown theoretically in Section 4.1. The value of H_d is greater than the value of H derived in Section 4 and this is illustrated by the theoretical curves shown in Fig. 11, which show that as the amount of asymmetry in the binary waveform is increased the received frame-rate falls for a given signal-to-noise ratio. The effect of this asymmetry in the binary waveform on error-rate is deduceable from equation (16).

The measured results displayed in the form of crosses in Figs. 9, 10 and 11, show good agreement with the theoretical equations established in Sections 4, 4.1 and 4.2.

9 Acknowledgments

The authors thank Marconi-Elliott Avionic Systems Ltd. and Loughborough University of Technology for permission to publish this paper.

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Manuscript first received by the Institution on 26th June 1975 and in final form on 18th December 1975. (Paper No. 1737/AMMS 73.)

IERE News and Commentary

Ballot for Elections to Council 1976–77

The President, Dr. P. A. Allaway, announced on 7th October that the result of the postal ballot for the election of three Members to Council had been communicated to him by the scrutineers appointed at the Annual General Meeting on 6th October as follows:

Elected: N. G. V. Anslow
K. Copeland
J. M. Walker

Biographies of these Members were published on page 451 of the August/September Journal.

The total number of ballot forms posted was certified as 6321 of which 1984 were returned.

The scrutineers were: S. J. H. Stevens (Fellow), J. B. Bennett, J. D. Buffin, M. G. Hansen, V. J. Phillips (Members).

Facilities for Members through EUREL

An agreement was signed at the Fourth General Assembly of the Convention of National Societies of Western Europe in Milan last November whereby members of the 17 constituent societies, of which the IERE is one, may benefit from privileges extended by the other member societies.

These privileges are available to a member provided that he has paid the current annual subscription to his own society. Some of the facilities require the presentation of a special card of accreditation valid for the current calendar year. The agreement covers the following matters:

— a member resident abroad may use the facilities of the host society (e.g. common rooms, libraries, etc.) on the same terms as members of the host society, on production of the card of accreditation

— a member may attend meetings, conferences, colloquia etc. of the host society on the same terms as that society's own members, on production of the card of accreditation

— a member may obtain publications of the other member societies at reduced prices, normally a 20% discount on prices to non-members, if the order is placed through the secretariat of his own society.

The following societies are the parties to the EUREL agreement:

AUSTRIA: Österreichischer Verband für Elektrotechnik
BELGIUM: Société Royale Belge des Electriciens
Association des Ingénieurs Electriciens sortis de l'Institut Electrotechnique Montéfiore
DENMARK: Ingeniør-Sammenslutningen
FINLAND: Sähköinsinööriliitto r.y. Elektroingenjörsförbundet r.y.
Elektroniikkainsinöörien Seura r.y.
FRANCE: Société des Electriciens, des Electroniciens et des Radioélectriciens
FEDERAL REPUBLIC OF GERMANY: Verband Deutscher Elektrotechniker e.V.
ITALY: Associazione Elettrotecnica ed Elettronica Italiana
NETHERLANDS: Koninklijk Instituut van Ingenieurs
Nederlands Elektronica- en Radiogenootschap
NORWAY: Norsk Elektroteknisk Forening
SPAIN: Asociación Electrotécnica Española
SWEDEN: Svenska Elektroingenjörers Riksförening
SWITZERLAND: Schweizerischer Elektrotechnischer Verein
UNITED KINGDOM: Institution of Electrical Engineers
Institution of Electronic and Radio Engineers

Members who expect to visit any of those societies with a view to using any of the facilities available under the agreement may obtain cards of accreditation from the Membership Secretary, IERE, 8–9 Bedford Square, London WC1B 3RG. Please enclose a stamped addressed envelope (UK members) or a self-addressed envelope with International Reply Coupons (outside UK).

A handbook giving the Statutes of EUREL, the terms of the agreement referred to above, and full details of each society such as its address, number of members, publications and activities and special interests, has been published. Copies may be obtained from the IERE Publications Department, price 50p.

Proposed Institution Activity in Sweden

Although the number of Institution members throughout Sweden is relatively small, it has been suggested that there would be support for occasional informal meetings between members living and working there as well as with members visiting the country on business.

Members interested in furthering these initial proposals are invited to write to:

Mr Lars O. Stromberg, C.Eng. M.I.E.R.E.,
Director, Utec Electronics, S-931 Skelleftea, Sweden.

Those who find that at some future date they are to visit Sweden may like to make contact through the Institution's offices in London.

Another Centenary

One hundred years ago, Melvil Dewey (1851–1931) published his 'Decimal Classification and Relativ Index', a system for arranging books in libraries. The scheme assigns numbers to subjects in such a way that books on similar subjects were grouped together. The great innovation was that the numbers were put on the books and not on the shelves, thus introducing flexibility into the system. Another advance was the way the numbers were constructed: from the general to the particular, a longer number for a more detailed subject. New subjects could be added without re-arranging the whole library or its catalogue.

One reason for the success and continuing use of Dewey's scheme was its early adoption by public libraries both here and in the United States. Although very good for such general libraries, it was felt that it was not detailed enough for specialist libraries. An international scheme, based on Dewey's but allowing more detail was compiled by the Institut International de Bibliographie, Brussels. This is the Universal Decimal Classification adopted as a British Standard (BS 1000) and used by a large proportion of special libraries such as our own. Recent editions of Dewey have, however, accepted feedback from other schemes, and UDC and Dewey could converge and be united in the future.

Melvil Dewey was also very interested in rationalizing the spelling of English, believing in a phonetic system, hence the 'relativ' of his title. At one time he went so far as to spell his own name 'Dui'!

A Confederation to Protect Professional Interests

LORD HAILSHAM, of St. MARYLEBONE, P.C., C.H.

The professions should be more united and self-conscious of their separate interests in putting forward their distinctive point of view, emphasized Lord Hailsham when he delivered the Fourth Rivers Lecture to The Institute of Chartered Secretaries and Administrators. Saying that the professions were unrepresented at the highest levels Lord Hailsham asked: 'If there can be a Confederation of British Industry why can there not be a Confederation of British Professional Associations, however loosely associated, to identify the problems which affect us all'?

A speech by one professional man to others should normally be unemotional, objective, and if possible, full of learning, but I speak to you in some anger. I believe our services are undervalued in modern society and in particular in modern Britain. I believe our rewards are insufficient. I believe our earnings are taken away by excessive taxation—in this country vastly excessive as compared with others on either side of the Iron Curtain. I believe our savings are being destroyed by inflation even when they are not being deliberately confiscated by unjust fiscal measures.

I believe that our most valuable possession, which is our integrity and our independence of judgment, is under threat. In the meantime we have suffered a catastrophic fall in our standard of life and the current negotiations between the Government and the TUC (neither of whom represent our interest) are predicated upon the proposition that this situation shall not merely continue, but that the fall should be accelerated. I believe that all this is happening not as a mere matter of chance, still less the result of irresistible changes or impersonal forces. It is happening partly as the result of deliberate policy and partly by our own failure to present a united front and proclaim in a coherent manner our philosophy of life and the individual contribution which professional men and women make to society at large.

Of course, my plea would not be valid unless I were able to establish that professional men and women were a valuable component in society, and unless I were able to establish in addition that their continued existence as a separate component was the essential condition of this contribution being made. But that, of course, is precisely my contention and it is around this contention that this lecture is being built.

I must begin with a disclaimer. By saying that our contribution is unique and individual, I do not mean that it is better than that of others. This would be foreign to the whole outlook on society of professional men and women. I only mean that it is different, and that its continuance is essential to a free and civilized society. The professional man has no class antagonisms. He is not hostile to manual labour, organized or not organized. He is neither hostile nor subservient to management in industry, to the agriculturalist, to the landlord, to the tenant, or to any other class. On the contrary, he provides services to them all. But the view of society to which he subscribes and which alone is congenial and conducive to his interest is that it should not be uniform. The mark of a free and civilized nation is not its rule by majority vote, but the treatment it accords to minorities, and, where they exist as a majority also to what are called ungrammatically 'underprivileged groups.' Uniformity—and for that matter equality whether of income, property or esteem—can be bought, if at all, which is doubtful, only at the price of repression. Uniformity is the enemy of freedom; imposed equality is inconsistent with justice, social or individual.

Reprinted from *Professional Administration*, the Journal of the Institute of Chartered Secretaries and Administrators, June 1976.

Defining a Profession

In the age of Trollope's novels, the professions *par excellence* were, of course, the Church, the Law, Medicine and the Officer grades in the armed services. But this is not merely wrong today, even if it were ever right; it is in fact positively misleading. The growth in the number of professions shows it to be false; an attempt at definition simply by enumeration of the individual members of a class is bound to be misleading. Professions today are more numerous than they have ever been in the past, and their services more than ever essential to the public weal.

The first, and obvious, feature of a professional is that even when a lawyer would define him as being under a contract of service, which is true of an increasing number of professionals today, he is in fact the provider of skilled services for which both an academic and a vocational qualification are required. These require previous training, and years of delay before earning capacity is achieved. All professional men and women therefore begin their lives with a sacrifice of leisure and income during the period for which training is required. There are, of course, some professions like the administrative civil service and the foreign service recruited by competitive examination, where vocational training is mainly post-entry. But most professions required pre-entry training and most characteristically set the required standards themselves.

This last characteristic is of great importance. That a profession should set its own standards of qualification is part of its general claim to autonomy of self government as a separate group within the community. Obviously if a profession were to set its standards much too low, with the result that insufficiently qualified persons were able to practise, or impossibly high so that an insufficient number of adequately competent persons could not be reached, that would be a matter of legitimate public concern. No profession or group of professions can set itself up as a state within a state, claiming to be immune from public criticism or public legislation. But autonomy is essential to professional status and it is essential as a safeguard to the public which requires in such matters to be protected against shoddy standards and political interference.

Rules of Conduct

This autonomy does not limit itself to qualification for entry. It is essential that a profession should be in control of its own special disciplines and professional ethics. Each profession has rules of conduct which are binding on its members. It is of course important that these, too, should not conflict with the generally accepted ethical code to which all are expected to conform, and that the codes involved should be aimed at the public advantage and not merely the advantage of the profession itself. But the idea that a profession can exist without a special code of conduct binding upon its members is not, I think, acceptable. The rule that chartered accountants, barristers, solicitors and doctors in private practice should not advertise, and the various regulations which are made from time to time varying, relaxing, or extending this general prohibition are aimed as much at the protection of the public against the quack as the protection of the practitioner against unfair competition. The rule that the barrister or the medical consultant does not accept instructions direct from members of the public is to protect the specialized character of the service he provides, and not simply to provide jobs for solicitors or GPs.

If a profession is to be autonomous in respect of its ethical rules, it must equally be autonomous as to its means of enforcement. This is not to say that I necessarily criticize the co-option of lay members on the the Solicitors Disciplinary Committee, or the particular, and publicly promulgated, constitution of that or the General Medical Council. It is important that members of the public should have confidence in the tribunal and that both the profession and the public should be protected. But the general rule should be that the profession itself should declare and police its own rules subject only to overriding considerations of public interest and public policy. In this respect the professions differ from the requirements of industry and commerce where quality control of the product is best assured by competition and advertising and the danger of the quack and the charlatan is in general less acute, and where present, should be the subject of legislation.

A third, but essential, ingredient in the continued existence of a profession as such is that the professional should remain independent in his judgment and in the integrity of his advice whether he remains in private practice as a freelance, or whether he operates under a contract of service as part of an industrial or public concern. I imagine that the overwhelming number of chartered secretaries and administrators are so employed. So is an increasing number of solicitors and barristers and so are the doctors employed in hospitals or by the NHS. Chartered engineers and chartered accountants are not characteristically one or the other. Nor are architects. But no employer has the right to ask a salaried professional to prostitute his professional judgment, and his professional association ought always to be prepared to fight for his interest if it is threatened in this respect.

Having said all this it is manifestly a matter of degree when a professional becomes part of management. I have no doubt whatever that there are hundreds of barristers, solicitors secretaries, administrators, accountants, surveyors, architects, engineers who have moved upwards, downwards or sideways, out of purely professional practice and into purely directorial, managerial or financial posts. Moreover, there are certain business activities, like that of a banker or bank manager, which partake of many of the characteristics of a profession. On the other side of the scale there are professions which grade imperceptibly into the ranks of skilled labour, and vice versa.

Ever-changing Society

Society is not a caste system, nor a class system with fixed boundaries. It is more like a constantly shifting biological group. There are journalists who regard themselves primarily as professionals. There are others who prefer to think of themselves primarily as skilled operatives. There are school teachers who think of teaching primarily as a profession. There are others who seem to behave as if it were only another skilled occupation. There are engineers who think of themselves as technical assistants; others regard themselves as professionals. Who is to say who is right and who is wrong? We are not talking about watertight compartments. A man is very often exactly what he thinks himself to be.

Having thus discussed rather than attempted exactly to define our terms of reference, I now come to discuss the place of the professional in society. In the first place I would point out that by nature the professional desires to have and by preference chooses to play no part whatever in class warfare. He is neither, in any big way, a have or a have not. No doubt the professions contain by achievement or the lack of it many who are one rather than the other, or vice versa. But, characteristically a professional is a member of the middle class, hoping in his best years to earn more than the average wage as the result of the years he has spent in acquiring his qualification, trying to save, attempting to found a family which will do him credit, and conform to his own standards of ethics and social acceptability.

Characteristically he will wish to retire on something more than the state pension and leave property to his widow and family, at least sufficient to achieve comfort. Characteristically he will be a patron of literature and the arts at least to the extent of acquiring a modest general library, going to the theatre and concerts, and perhaps collecting objects of beauty and value. I do not mean to say that no one else does this. I only mean to say that I know of few professionals who do not do at least some of these things, that they do so because the nature of their education and training inclines them to do so, and they do so characteristically, as much as or more than any other segment of society.

I have said enough, I hope, to show that the professional adds a sophisticated, educated, critical, independent minded note essential to the continuance of a free society. He is not, characteristically, and perhaps happily, a member of any particular political party. A number of us belong to one or the other of these, and we hope to bring with us when we do the characteristics which belong to our chosen way of life. When we do not play a part in controversial politics, as is true, perhaps, of a majority amongst us, we find our services much in demand on the committees and in the organization of a vast number of charitable, social and cultural organizations, whose very variety precludes description, but who form the glory and provide the substance of a free and civilized nation.

Diminishing Standards of Life

I began these remarks with words of some bitterness. But at this stage, I feel sure an unfriendly critic would be inclined to ask why it is, if all this be true, and if we are really the high minded, cultivated, public spirited, talented group that I have claimed us to be, do we find ourselves to be persecuted, do we suffer a deliberate denigration of our ideals, and a continuous, progressive and increasingly rapid diminution in our standards of life.

I would reply, first, that the nature of the services we provide has been almost continuously undervalued. It is easy of course to praise the operatives who make an object of use or beauty, which is manufactured and perhaps exported. On the other side we occasionally—more seldom nowadays, but still sometimes—read words in praise of the entrepreneur, the man with enterprise, the bicycle repairer turned motor manufacturer, or even the man to discern the economic possibilities of a building site and the drive and capacity to realize and achieve its potential. But, as I believe to our great disadvantage, we do not often hear of the design engineer who first conceived the product, and drew it on the board.

We do not hear much of the men or women who first laid out the factory or the production line, who learned the skills or the languages necessary to sell the product overseas, or who provide the necessary accountancy, secretarial or administrative skills to let the whole business work smoothly, to keep it solvent, or the legal skills to draft its contracts, to fight its battles (few, one hopes) in the Courts, or the medical skills to keep either its managerial or industrial staff in health and at work. This is because you cannot easily measure the value of brains, nor the advantage of technical training. In my own sphere, if you win a case for a client, it is justice which has prevailed, if you lose it, it is the incompetence of the advocate, and since advocacy, and medicine, and accountancy, and administrative capacity largely depend on imponderables and can often be frustrated by ill-luck, you seldom hear of the virtues of their practitioners. On the contrary, it is when things go wrong that they are most heard of, and then, alas, it is almost always by way of criticism and complaint.

But the mere fact that brains are undervalued in this country is not the only reason why the professional becomes unpopular. He is the constant proponent of advice, and advice is seldom popular except when it is palatable, and when it is palatable it is almost always unnecessary. The lawyer is always advising that a scheme which at first sight looked so attractive is, unfortunately, illegal. The physician proscribes as well as prescribes, and what he proscribes are normally the nicest things to eat and drink, or the darling habits which lead to obesity and heart disease. The accountant is always telling his client or his employer that he cannot afford it, and the administrator that you must not or cannot do what you want. We are therefore the eternal whipping-boys of management, and the obvious targets of organized labour. We are considered to be the lackeys of capitalism, when all we are doing is to remind others of the facts of life when it is least to their taste to remember them. If we are to be respected we have always first to fight for our integrity with tooth and claw.

I would now wish to list some of the dangers to which we are exposed. There is the danger from government. There is the danger from corporations, private and public. There is the danger from organized labour. Each and all of these influences is gradually encroaching on the integrity and independence of the professions.

Tip of the Iceberg

This is a gloomy forecast, but anyone with an eye to realities can see that it is true. The recent battle between the doctors and the Health Service is an example of the first. For years now the Health Service has battened upon the devotion and enthusiasm of the junior doctors, driving them to inordinately long hours of duty at impossibly low salaries. Political influence is seeking to destroy the independence of consultants. There is a continual battle between public authorities, and even private industry, and the administrators they employ. Only the worst cases, like Clay Cross, become public. But the worst cases are only the tip of a far larger iceberg. The fiscal policies of successive Governments might have been expressly designed to diminish the standard of life of professionals and middle management, and in some cases have undoubtedly been so designed.

The danger from trade unions has only just begun to be seen. But it is quite clear that a trade union affiliated to the TUC does not and cannot adequately represent the interests of a chartered accountant or engineer or administrator, or doctor, or a barrister whether he be salaried or in private practice. This is not because one is better than the other. It is that they are different kinds of entity. A professional needs a professional association to represent him, and although professional associations, such as your own, or the Bar Council, or the various Engineering Institutes, or The Law Society have many of the characteristics of trade unions, they are by no means the same, and they do not or at least have not until recently felt themselves as free to apply industrial pressures in support of their members as do unions of the other kind.

But, increasingly, of recent years, and especially since last year's Trade Union and Labour Relations Act, pressure has been brought by trade unions to force genuine professionals to leave their own professional associations which understand their needs and requirements and compel them to join with non-professionals in unions of the usual kind on the top salaried fringe of the trade union movement. Whatever may be said of a closed shop for workers of the same grade, an attempt, in the interests of a closed shop to force members of the professions into unnatural membership with those who have not the same interest is something which ought to be resisted by all who have the interest of the professions at heart.

What then is the answer to these problems? I suggest two. The first is that the professions should be aroused to their dangers and should be more united and self-conscious of their separate interest in putting forward their distinctive point of view. This does not mean that professional associations should take sides in party politics. For many years I have thought that the organic relationship between the TUC and the Labour Party has been a corrupt and corrupting influence on our public life, and is indeed a significant contributory cause of our national decline.

I would not wish professional associations to take a similar line. But to say that they should not take sides as between political parties is not the same thing as to say that they should not associate with one another. If there can be a Confederation of British Industry why can there not be a Confederation of British Professional Associations, however loosely associated, to identify the problems which affect us all, to pressurize politicians, to proclaim our distinctive points of view, to defend our interests, and to advertise the importance to society of a strong, independent, incorrupt, public-spirited professional class not identified with management or organized labour, nor with the private or public sector of industry as a whole, but permeating society by reason of the services provided for each sector of the whole?

It is as well to remember that the professional is not without natural allies in his struggle to survive. Although essentially middle class in outlook he is not the whole of the middle class. Although often self-employed, he does not represent the whole of the self-employed. He naturally finds himself in the same plight, and therefore in the same lobby as middle management, as the small shopkeeper or business proprietor, as the farmer, as the skilled technician or workman. The only trouble is that he has not yet learned to organize and fight the political battle either in the field of party politics or outside it. It is time that we came to look at one another and recognize in one another persons who are suffering from the same pressures, and who all need allies not simply to secure personal survival but to ensure the continuance of standards in society without which independent and honourable professions cannot continue to exist.

CEI News for Members

The New Presidents

Mr. Gerald James Mortimer, M.B.E., B.Sc., A.R.S.M., has been nominated President-Elect of the Institution of Mining and Metallurgy and will assume office in May 1977. Mr. Mortimer is Group Chief Executive of Consolidated Gold Fields Ltd. which he joined in 1947 as a mine captain on the Rand.

This year's President of the Institution of Mechanical Engineers is Dr. Ewen M'Ewen, C.B.E., D.Sc., F.R.S.E. He is Vice-Chairman (Engineering), i.e. Group Chief Engineer, of Lucas Ltd. and his public appointments include that of Chairman of the Metrology and Standards Requirements Board, Department of Industry.

Mr. Charles Abell, O.B.E., who retired in May 1974 as Chairman of British Airways Engine Overhaul Ltd., has taken office as President of the Royal Aeronautical Society.

On 1st October 1976, Mr. E. S. Booth, C.B.E., M.Eng., F.R.S., Chairman Yorkshire Electricity Board, Leeds, will take over from Mr. R. J. Clayton, C.B.E., M.A., Technical Director, The General Electric Company Ltd., as President of the Institution of Electrical Engineers for the 1976/77 Session.

At the 66th Annual General Meeting of the Institution of Structural Engineers in London recently, Dr. W. Eastwood, B.Eng. Ph.D., was elected to succeed Mr. Peter Mason as President of the Institution for 1976–77. Before forming his consultancy practice of W. Eastwood and Partners of Sheffield in 1970, Dr. Eastwood was Professor of Civil Engineering at the University of Sheffield.

Professional Engineers should Join Union to Exert Proper Influence

There were three reasons why CEI had recently been urging professional engineers to join a suitable trade union said Mr. Tony Dummett, the Council's chairman. These were the steady erosion of the status of the professional engineer reflected in his salary, closed shop legislation, and industrial democracy.

Mr. Dummett, who was speaking at the annual luncheon of the Institute of Fuel, said: 'There are now many chartered engineers—and there will be more—who are being forced by the crudest power tactics to join unions with which they have no sympathy, which cannot respect their code of conduct and in which their voice will be swamped by the non-professionals.'

Referring to industrial democracy, Mr. Dummett said: 'In all the schemes so far proposed, whether it be for extra seats on the board, for consultative councils or whatever, only the trade unions are written in. And this means—make no mistake—the big militant unions and none other if they know it. Yet this will exclude the profession whose voice, one would think, would be essential in discussing the operation and development of our great technological enterprises in the national interest.'

CEI had thus concluded, continued Mr. Dummett, that engineers and indeed other professional men in industry would be unable under present circumstances to exert their proper influence unless they were organized in trade unions. 'But they must be the right sort of trade unions', he emphasized, 'and in particular they should be ready to take due account of the profession's code of conduct.' They must be basically, though not necessarily exclusively, professional in membership.

If strong trade unions—or even better a single trade union—could be developed with the sort of membership he had outlined, continued Mr. Dummett, it would be governed by men of conscience. 'No effective union can completely abrogate industrial action; it is its only ultimate sanction. But the sort of trade union we are recommending will use it only as a last resort, not as a first one as so many do. It will not *threaten* action until every possibility of conciliation and arbitration has been tried and failed and it will not *take* such action until all the membership has been balloted by secret procedure and has declared itself overwhelmingly in favour.'

Earlier Mr. Dummett had said that the most important problem facing CEI was the position of the engineer in modern society. It was a simple fact that most of the things we took for granted would not exist or continue to be, without professional engineers. 'As such, one might expect that the key role of the profession would be recognized and that it would be universally honoured. But, on the contrary, the status of the engineer is inferior, his rewards relatively low and his image poor. This is a state of affairs that, in the interests of the country, cannot be allowed to continue.'

The steady erosion of the status of the professional engineer said Mr. Dummett, was recognized by CEI as a grave threat to the well-being of the profession it was set up to protect. It was this that had led the Board of CEI to set up its working party whose report 'Professional Engineers and Trade Unions' was published recently. This had revealed a real need for professional engineers to join trade unions, a need which had not been generally recognized.

Changes at the Institutions

The Council of the Institution of Mechanical Engineers has announced that Mr. K. H. Platt, C.B.E., C.Eng., F.I.Mech.E., will be retiring as Secretary of the Institution in the autumn of 1976, and that they have appointed Mr. Alex McKay, C.B., C.Eng., F.I.Mech.E., F.I.E.E., as Secretary.

Mr. McKay is at present General Secretary of the Institution of Chemical Engineers. Prior to taking up this post he served for thirty-two years in REME; he was Director of Electrical and Mechanical Engineering (Army) in the rank of Major General from 1972 to 1975.

Dr. Trevor J. Evans, B.Sc., Ph.D., C.Eng., M.I.Chem.E., has been appointed General Secretary of the Institution of Chemical Engineers in succession to Major General A. M. McKay, C.B., C.Eng., F.I.E.E., F.I.Mech.E., with effect from 1st October 1976. Dr. Evans is at present Deputy Secretary of the Institution, with special responsibility for learned society and technical matters.

Mr John Sparey, M.A., has been appointed Secretary of the Institution of Municipal Engineers in succession to Mr. A. Banister, O.B.E., B.Sc., C.Eng., F.I.C.E., F.I.Mun.E. who is retiring at the end of the year. Mr Sparey has been with the Royal Institution of Chartered Surveyors since 1969, since 1974 as Secretary of two of its Divisions.

Change in I.Mech.E. Membership Requirements

Currently a person who is a Chartered Engineer by virtue of membership of another CEI institution may, subject to the acceptability of his or her professional training, subsequent experience and responsibility, be elected a corporate member of I.Mech.E. on the basis of the 'old' – pre-1974 – academic requirements. This concession will be withdrawn on 31st December this year after which all candidates for corporate membership, whether or not they are already Chartered Engineers will have to satisfy the academic requirements now in force.

Letter to the Editor

From: **C. E. S. Ridgers**, C. Eng., M.I.E.R.E.

'Fighting for the Professional Engineer'

Linda Dickens's article 'Fighting for the Professional Engineer' in the June 1976 issue of the Journal makes very sad reading. To realize that there is now inter-union rivalry to snap-up Professional Engineers into various organizations (none of which is truly representative of the Professional Engineer) and, in some cases, to use the closed shop 'press-gang' principle to ensure compliance, is to realize that professionalism in industry may well become extinct.

Few Engineers wish to be part of a large, amorphous, remotely-controlled Trade Union, whose aims and politics are the antithesis of their own. The article mentioned, in a rather disparaging tone, that few Engineers actually joined a certain union. This tone colours all articles written about Professional Engineers and Trade Unions, and their reluctance to join such organizations. It is as if Linda Dickens and all such writers feel that it is the Professional Engineer's duty to join a union, and by not doing so, he was being less than patriotic (or whatever passes for patriotism today).

Linda Dickens made another point that, according to the CEI report, almost 90% of the Professional Engineers are in unions affiliated to the TUC. This fact she accepts as proof that *all* engineers should be TUC-affiliated, and that *all* engineers would have no objection to being TUC-affiliated. It appears that all articles written about the Professional Engineer and Trade Unionism are biased towards unionism and the TUC. Nobody appears to have asked the Professional Engineer (from any institution) what his feelings are. When we consider that the future of this country lies with the skill, inventiveness, dedication and enthusiasm of the Professional Engineers one wonders why he has been so neglected by those in power and authority (not necessarily vested in the same group) in this country.

Neglect of the Professional Engineers' status and erosion of their differentials will not force them to join the contending unions, but will persuade them that their profession and/or the country is not worth working for. Such a result will be a disaster for Britain.

The advice of the CEI has been met with cold belief from most engineers. I doubt if any now believe that the CEI has any useful purpose, and some are beginning to wonder if the Institutions themselves have a viable future. Certainly if Professional Engineers are forced, against their will, to join certain Trade Unions, a time will inevitably come when they must choose between professionalism (i.e. the Rules of Professional Conduct of their Institution) or the orders of their union.

How long can the Institutions survive attacks of this nature against their very structure? Not for very long.

There is one answer to the dilemma: a Trade Union for the Professional Engineer, run by Engineers, entirely independent of the Institutions and of any other organizations. Such a union would give advice to its members, negotiate salaries and contracts and the myriad other matters which help the Professional Engineer to fulfil his place in Industry. We are told that there are between 200,000 to 300,000 Chartered Engineers. If only 10% paid £10 per annum, the annual income would be nearly £¼M. Enough for a fulltime staff of advisers and negotiators, housed in the Union's own offices.

The Doctors have their union, the Air Pilots theirs. So too have the Electrical Power Engineers in the Electrical Power Supply Industry. The Teachers are organized and vocal. The future of Britain lies with the Professional Engineer of all the Institutions. The Institutions alone do not give him a voice in the country's affairs.

But a Trade Union with muscle power will.

71 Waltham Avenue, C. Ridgers
Guildford, Surrey.

9th August 1976

Launch of Decca Ship Simulator

Designed for the training of bridge teams, including ships' officers, pilots and helmsmen, the Decca Ship Simulator is not only the first of its kind in the UK but includes features which are not to be found in any other ship simulator in the world. The manoeuvring behaviour of the simulated vessel is governed by a computer mathematical model, developed jointly by Decca Radar Ltd. and the National Physical Laboratory Ship Division, who will be taking delivery of the Simulator, which has been ordered by the Department of Industry.

It is built into two 36 ft (11m) Portakabins, one containing the wheelhouse, equipment room and Captain's day cabin, and the other housing the bridge window projection system and screen. The general design is based on extensive trials with experienced Masters and Pilots, and provides a high degree of realism. Approach channels of any length and complexity can be set up with no artificial restrictions on position of own ship. Navigation marks and other ships can be included in the same exercise.

The Simulator is capable of providing anti-collision, navigation, pilotage and ship handling exercises for ships between 500 tonnes and 500,000 tonnes, with the vessel responding correctly to wheel and engine controls, and including the effects of tidal stream and depth of water under the keel.

Incorporated are a wheelhouse with a bridge control console including wheel and autopilot, engine controls, anti-collision radar, ship's telephone, radio communications, warning annunciators and chart table.

A major innovation is a bridge window through which can be seen the bows of own ship, lights of navigation marks and lights of other ships. Up to 16 lights can be shown at one time, for example 8 buoys and 2 other ships. The lights move with the correct perspective as own ship and the other ships move, and are properly correlated with the echoes on the radar display. Mounted above the bridge window are instruments, including heading repeater, log, rate of turn indicator, engine revolutions and clock. Engine and propeller noises and vibration are generated, varying correctly with engine revolutions.

The Simulator may be programmed for real or artificial exercise areas, which can be changed in a few minutes by inserting different magnetic tape cassettes. Each exercise is automatically recorded on a track plot for subsequent analysis, together with recordings of rev/min, ship speed, rudder angle, rate of turn, drift angle and heading.

Universities Consortium for Mobile Radio Research

Over the last few years a novel development in collaboration in research programmes between British universities has been evolved by the Universities of Bath, Birmingham and Bradford. The three Electronics and Electrical Engineering Departments have set up a research consortium, currently spending £100,000 a year on mobile radio investigations, the overall purposes of which are to improve techniques generally but also to make such applications as data transmission and visual displays more practicable and efficient.

Much of the work is supported financially by the Home Office Directorate of Telecommunications, which is concerned particularly with police, fire and other emergency services. The Science Research Council, the nationalized Fuel and Power industries and the Ministry of Defence also give support.

The three universities first collaborated in 1974. The success of the experiment has now led them to form the Universities Mobile Radio Research Consortium 1976 (UMRRC) to pool information, share research equipment, define common objectives and ensure that there is no duplication of effort. Broadly speaking, Bath, under Professor William Gosling, concentrates on transmitter and receiver technology; Bradford, under Professor David Howson, concentrates on problems arising from interference between 'co-sited' transmitters; while Birmingham, under Professor Ramsay Shearman, is concerned with what happens to the signal between transmitter and receiver.

The main obstacles to more extensive use of mobile radio are fading of the signals caused by the reflection of radio waves from buildings or other obstructions and also interference from electrical sources such as machines and car ignition. These effects are most serious with teleprinter communications, now being used by the police to replace the laborious transmission of data (e.g. numbers of stolen cars) at dictation speeds. The consortium is attacking fading and interference in various ways. Bath is working on a receiver which can recognize, and partially suppress, interference signals. Birmingham has developed diversity reception techniques in which aerials on different parts of a vehicle pick up the same signal at different points of the wave pattern. An 'add-on' unit, no bigger than a small portable radio has been developed for v.h.f. amplitude modulated equipment and a version for u.h.f. frequency modulated equipment is now in prototype form.

With a similar purpose in mind, Bath is working on the use of multiple transmitters in an alternative 'diversity transmission' arrangements. This too can give a constant received signal for speech or data.

Research at Bradford examines interference at mobile radio base stations from other co-sited transmitters and from the so-called 'rusty bolt effect' when unwanted interference is produced from radiation reflected off rusty metal structures. A suite of computer programs has been developed to predict the magnitude of interference over the whole mobile radio frequency spectrum and equipment to minimize the effects has been constructed.

At the request of the Ministry of Defence, Birmingham is also working to produce a formula whereby optimum siting of base stations for mobile radio coverage of a particular city would be determined in advance. This would enable equipment to be sited by computer calculations, eliminating the lengthy trial-and-error period that is now needed. In the course of this work the mobile laboratory established by the Department of Electronic and Electrical Engineering has compiled 'reception profiles' for the cities of Bath, Birmingham and Bradford.

The work to open more channels is also going ahead. Mobile radio at present uses 70–170 MHz and 425–470 MHz. Originally an interval of 25 kHz was maintained between channels, allowing several thousand for mobile radio in particular geographical area, e.g. West Midlands County. With the coming of improved crystal oscillators and microelectronic frequency synthesizers it became possible to control operating frequencies within much finer limits: the Home Office have reduced the interval between channel allocations to 12·5 kHz, thus doubling their number and the next such step might be 6·25 kHz. The ultimate interval would be about 5 kHz, because a single-sideband speech channel itself requires 3 kHz.

Further improvement of crystal oscillators, frequency selective filters and many other aspects of the equipment will be needed before this ultimate interval can be realized. Bath University have already recently demonstrated some of the new transmitters and receivers required, and with the important research at Birmingham and Bradford on the many fundamental scientific problems involved, there is a good prospect of ultimately finding solutions to the problem of giving increasing numbers of users ever more sophisticated personal and mobile radio.

A considerable fillip to the future financial viability of the research programme planned by the University of Bath in particular has recently been announced by Professor Gosling. The Wolfson Foundation is to make a grant to his Department of £102,000 to support its work on narrow-band system techniques for mobile radio.

Some of the work of the members of the Consortium has been or will be described in The Radio and Electronic Engineer or at IERE Conferences.

J. D. Parsons and M. Henze, 'A proposed pre-detection combining system for mobile radio', Proc. of Conf. 'Radio Receivers and Associated Systems', Swansea, 1972.

M. J. Withers, 'A diversity technique for reducing fast-fading', Proc. of Conf. 'Radio Receivers and Associated Systems', Swansea, 1972.

J. D. Parsons and P. A. Ratliff, 'Diversity reception for v.h.f. mobile radio', *The Radio and Electronic Engineer*, **43**, No. 5, p. 317, May 1973.

J. D. Parsons, M. Henze, P. A. Ratliff and M. J. Withers, 'Diversity techniques for mobile radio reception', *The Radio and Electronic Engineer*, **45**, No. 7, p. 357, July 1975.

S. A. Mawjoud and J. G. Gardiner, 'Some origins of radiated noise from communal transmitting sites', Proc. of Conf. 'Civil Land Mobile Radio', Teddington, 1975.

J. P. McGeehan, 'Design and characterization of a phase-locked v.h.f. land mobile radio receiver', The Radio and Electronic Engineer (to be published).

Members' Appointments

CORPORATE MEMBERS

Mr. E. G. Lamb, B.Sc. (Eng.) (Fellow 1965) has been reappointed for a further period of three years as an education member of the Iron and Steel Industry Training Board. He is Principal of Bell College of Technology, Hamilton, Lanarkshire.

Mr. C. G. Baillie-Searle (Member 1972, Graduate 1968), formerly a Lecturer at the National College for Advanced Technical Education, Durban, South Africa, has returned to the United Kingdom to take up an appointment as a Senior Electronic Design and Development Engineer with AEI Semiconductors, Lincoln.

Lt.-Cdr. D. J. Carver, RN (Member 1973) who has just completed two and a half years as Executive Officer and Second in Command HMS *Sheffield*, has been appointed Staff Communications Officer at Headquarters, British Forces, Hong Kong.

Mr. P. Dunne (Member 1973, Graduate 1970) has for the past two years been a Television Systems Planning Engineer with Robert Bosch Fernsehanlagen in Germany. and has now returned to the Republic of Ireland as a Project Engineer with the Television Design Group of Radio Telefis Eireann in Dublin.

Mr. J. Epps (Member 1973, Graduate 1971) has taken up the post of Chief Electronic Engineer with Delta Controls Ltd., Kingston-upon-Thames, Surrey, following a short period as Chief Electronic Engineer with Newman Industries. Mr. Epp's previous experience includes posts with Moore Reed & Co. Ltd. of Andover, and Masson, Scott, Thrissell Engineering Ltd.

Mr. D. T. Hambley (Member 1973, Graduate 1968) has been appointed Senior Engineer with Gould Advance Instruments, Hainault, Essex. He was previously Chief R.F. Engineer with Green ECE Limited.

Mr. J. King (Member 1970, Graduate 1956), a Professional and Technology Officer Grade 2 with the Ministry of Defence, has moved from the School of Signals, Blandford Camp, to the Royal Signals and Radar Establishment, Malvern.

Mr. D. M. Lindsley (Member 1969, Graduate 1962) who for the last four years has been in charge of the Systems Engineering Department at Bailey Meters & Controls Ltd., has been made Engineering Director responsible for all systems engineering, produce design and development activities of the firm.

Mr. D. H. Mackenney (Member 1971, Graduate 1963) has taken up the post of Product Applications Manager for RCA Picture Tube Division (UK) at Southport. He was formerly Quality Manager at Thorn Colour Tubes Ltd., Skelmersdale.

Mr. S. M. A. Naqvi, B.Sc., Dip.El. (Member 1966) who since 1973 has been an instructor at the Civil Aviation Institute, Tripoli Airport, Libya, has now returned to Pakistan, where he has been appointed Chief Instructor Communications with the Civil Aviation Training Unit, Karachi Airport.

Mr. J. H. Parrott (Member 1973) is now Principal Electronics Engineer with John Davis & Son Ltd., Derby. He was previously Senior Engineer with GEC Elliott Process Automation Ltd., Leicester.

Mr. R. J. Pover (Member 1972, Graduate 1970) has taken up the appointment of Export Manager with ITT Components Group, Europe, based at Harlow, Essex. Mr. Pover was formerly Product Promoter in Marketing with the Group.

Mr. C. D. Putnam (Member 1969, Graduate 1967) who was formerly Senior Engineer with Goring Kerr Ltd., has joined Redifon Flight Simulation Ltd., Crawley, Sussex, as Principal Engineer. Mr. Putnam previously worked on flight simulation systems when he was Project Engineer with General Precision Systems Ltd. from 1962 to 1964.

Mr. J. M. Reid (Member 1973, Graduate 1968) who was Technical Director of South Yorkshire Automatic Control Engineering Ltd., Maltby, Yorkshire, from 1969 to 1976, has taken up the appointment of Managing Director of Jubilee Electronics Ltd., Blyth, Nottinghamshire.

Mr. B. J. Roberts (Member 1969) who was appointed Chief Engineer of the Radio & Television Corporation of Iran in 1972, has returned to the UK and is now Works Manager of Digital Electronics Ltd., Watford.

Wing Cdr. W. A. Russell, M.B.E., B.Sc., RAF (Ret.) (Member 1951, Associate 1950) has been appointed Senior Tutor in Applied Electronics at the Central Institute of Technology, Heretaunga, Upper Hutt, New Zealand. Wing-Commander Russell emigrated to New Zealand in 1963 after 16 years' service in the RAF, latterly as an Education Officer.

Sqn. Ldr. D. G. Salter, RAF (Member 1971) has been promoted on secondment to AGWT3, Ministry of Defence Procurement Executive Headquarters in London, after serving as Officer Commanding *Martel* Support Squadron, RAF Marham, Kings Lynn.

Mr. H. Sharman (Member 1966, Graduate 1960) is now a Project Leader with Perkins Engines Co., Peterborough.

Mr. T. Simms, M.Sc. (Member 1968, Graduate 1960) who went to Canada in 1969, is now Manager, Systems Development, Canada Post Office Headquarters. He was previously at Bell Northern Research.

Mr. R. J. Simpson (Member 1973, Graduate 1963) is now Development Specialist concerned with Resource Management in the Computer Services Department at the Midland Bank in Sheffield.

Mr. D. A. G. Tait (Member 1958, Graduate 1956) who was founder and Managing Director of Culton Instruments Ltd., has formed a new Company, LCR Measurements Ltd.

Mr. M. E. Tickner (Member 1973, Graduate 1969) who was Group Marketing Manager of Jiskoot Autocontrol Ltd., Tunbridge Wells, is now Marketing and Sales Director for the Company.

Mr. G. W. Waters, B.Sc., D.Phil. (Member 1973) is now an Associate Principal Engineer with the Harris Satellite Communications Operation, Melbourne, Florida, USA. Dr. Waters was previously a design analyst with AII Systems, Moorestown, New Jersey, working on the MARISAT satellite terminal, the rights of which are now transferred to the Harris Corporation.

Mr. R. I. Williams, B.Sc., P.E. (Member 1970, Graduate 1969) who has been Senior Control Systems Engineer with the Alyeska Pipeline Service Company since early 1975, has now been appointed Manager, Supervisory Control Systems with the Company. After going to the United States in 1969, Mr. Williams had experience as a designer with Systems Engineering Inc.

NON-CORPORATE MEMBERS

Mr. R. Bransbury (Graduate 1969) who was previously concerned with product marketing with SGS-ATES (UK) Ltd., has been appointed Chief Engineer with CADAC (London) Ltd.

Mr. C. S. Chew, B.Sc. (Graduate 1970) has been appointed Assistant Controller of Telecommunications in the Malaysia Telecommunications Department, Kajang, Selangor. For the past two years he has been reading for his B.Sc. degree at the University of Strathclyde.

Lieut. P. T. Cooper, RN (Graduate 1970) has been seconded to Meteo Dienst, Marine Vliegkamp Valkenburg, Katwijk, Netherlands, on exchange duties with the Netherlands Navy.

Mr. S. H. Gruszka (Graduate 1971) has been appointed Area Sales Engineer with MEMEC Ltd., Aylesbury Bucks. He was formerly Field Sales Engineer with Fairchild Semiconductor Ltd.

Mr. S. Isaksen (Graduate 1965) has been a television engineer in the Institute for Educational Research of the University of Oslo since 1968, and has now been appointed television consultant to the University.

Mr. Y. S. Lee, M.Sc. (Graduate 1966) has been made Senior Lecturer in the Department of Electronic Engineering, Hong Kong Polytechnic, which he joined in 1969 as a Lecturer. He was previously a Telecommunication Assistant with the Hong Kong General Post Office.

Mr. J. F. Pickup (Graduate 1971) who from 1964 to 1975 was a Lecturer in Illumination Engineering at the Polytechnic of the South Bank, has taken up the post of Senior Lecturer in Building Science in the Faculty of Architecture and Building of the University of Singapore.

Mr. C. J. Swinn, B.Sc. (Graduate 1974) joined AFA-Minerva (EMI) Ltd. in 1973 as Systems Engineer, becoming Senior Systems Engineer in 1975, and he has now been promoted to Marketing Product Manager (Communications) with the company.

Mr. A. L. Unthank (Graduate 1964) has taken up an appointment with Kennedy and Donkin, consulting engineers, as Project Engineer (Electrical) in Tehran. He has been a Site Construction Engineer (Electrical) for the company in Northern Ireland for the past three years.

Mr. G. Escritt (Associate Member 1974) has been appointed Instrument Engineer with the North East Region British Gas Corporation. Mr. Escritt was previously Assistant Communications Engineer with the Corporation.

Mr. F. J. Jane (Associate Member 1973) is now General Manager of AFA-Minerva, South West Division, at Bristol. Mr. Jane was formerly Computer Technical Education Manager at the NCR Technical Education Centre, Dundee, before going to AFA-Minerva in 1975 as Deputy General Manager of the South West Division.

Lt.-Cdr. J. W. Smith, RN (Associate Member 1974) has taken up the appointment of Officer in Charge at the Royal Naval Wireless Station, Crimond, Aberdeenshire. His previous appointment was as Radio Officer, Fleet Maintenance Group, Plymouth.

Mr. P. J. Watts, B.A. (Associate Member 1973) is now reading for the postgraduate degree of Master of Education at Bath University. Following experience in industrial and Government research, he has for the past seven years lectured in electrical engineering at the Somerset College of Arts and Technology, Taunton.

Mr. R. Jassal (Associate 1975) has been seconded for one to two years by Hasler AG of Berne as a Maintenance Engineer for fixed communications with the Saudi Arabian Airlines Corporation. For the past eighteen months he has been a test engineer working on programme control switching systems with Hasler.

Obituary

The Council has learned with regret of the deaths of the following members:

William James Brace (Member 1960, Graduate 1957) died on 11th November 1975 aged 47. Since 1960 he had been a Design and Development Engineer with Standard Telephones and Cables Ltd., Newport, Gwent.

Flight Lieutenant John Frederick Cogswell, RAF (Member 1973, Graduate 1969) died on 16th December 1975 aged 39 years. Since March 1974 he had been Officer Commanding Electrical Engineering Support Flight, RAF Locking.

Jack Haskel Davis (Member 1973, Associate 1946) died in May 1975, aged 58. He was a Standards Engineer in the Department of Quality Assurance of Elta at Ashdod, Israel. Mr. Davis had also worked for several years in the United States.

Michael John Graham (Member 1969) died in December 1975 aged 41. Since February 1975 he had worked with Strainstall Ltd. of East Kilbride as an electronic engineer.

Commander George Frederick Edmund Knox, RAN (Ret.) (Fellow 1971, Member 1955) died in December 1972, aged 59. Before his retirement from the Royal Australian Navy in 1958, Commander Knox held staff appointments in Australia and in London, latterly specializing in nuclear engineering. For the next ten years he worked in industry and in 1968 he was appointed a Senior Technical Officer in the Department of Navy, Canberra.

Ralph Thornton Lakin, M.B.E., F.I.E.E. (Fellow 1956, Member 1950) died on 27th July, aged 62 years. He leaves a widow and one son.

Ralph Lakin spent the whole of his professional life with Whiteley Electrical Radio Co. Ltd., which he joined as an apprentice in 1928. He obtained his technical education at the Mansfield Technical College, completing the Higher National Certificate in electrical engineering in 1940 and over the years he progressed through the posts of Technical Assistant, Design Engineer and then Research Engineer, until in 1950 he became Chief Engineer. He joined the Board as Technical Director in 1956 and on the death of the founder of the Company, Mr. A. H. Whiteley, in 1967 he was appointed Managing Director.

Known throughout the radio and associated industries for his work on loudspeakers and audio equipment, Mr. Lakin was also responsible for meteorological equipment and nuclear instrumentation. Between 1941 and 1948 he lectured on electrical and electronic engineering at Mansfield Technical College.

Mr. Lakin was awarded the M.B.E. in the 1953 Coronation Honours for service to the electronics industry, the year after he had supervised on behalf of the Government the installation of Meteorological Stations overseas.

From 1970 to 1972 Mr. Lakin served as a member of the Institution's Council and earlier he had been the IERE representative on the then Nottinghamshire Regional Advisory Council and on the organizing committees of Joint Conferences. He had often lectured to Local Sections on loudspeakers and meteorological equipment.

The Institution was represented at the funeral service in Mansfield on 30th July by Mr. P. A. Bennett, Past Chairman of the Yorkshire Section.

Domingo Michael Tunji Mobolaji Oke (Fellow 1964, Member 1960) died in December 1974 aged 57. For a number of years Mr. Oke was with the Nigerian Broadcasting Corporation, retiring in 1969 as Chief Engineer. He subsequently formed his own Company in Lagos.

Donald James Peffers (Member 1964, Graduate 1955) died in September 1974 aged 45. He was a Senior Engineer with EMI Electronics Ltd.

Arthur Rogers (Member 1963) died on 26th June 1976 aged 48. He had been with the Marconi Company since 1956 as a Radar Development Engineer.

Commander George William Keeton Whittaker, O.B.E., RN (Ret.) (Member 1950, Associate 1946) died in October 1975 aged 55. His last appointment before retiring from the Navy in 1962 was in HMS *Centaur*.

John Robert Wighton, B.Sc.(Eng. (Member 1967) died on 4th April 1976 aged 58 years following a long illness. He leaves a widow, a son and a daughter.

John Wighton served an apprenticeship as an Electrical Fitter and worked with East London engineering companies from 1931 to 1945. During this period he studied part-time at West Ham College of Technology for the Higher National Certificate and City and Guilds qualifications. In 1945 he was appointed an Assistant Lecturer at East Ham Technical College and in 1952 was awarded a degree in Electrical Engineering of the University of London following part-time study at the then Northampton Polytechnic. From 1955 to 1966 Mr. Wighton was a Senior Lecturer at Southend College of Technology, responsible for building up endorsement courses in Electronic and Electrical Engineering for the HNC. His next appointment, which he held until his death, was as Head of the Department of Engineering at North-East Liverpool Technical College and here too he considerably expanded the range of work.

Advice of the deaths of some of these members has only recently reached the Institution.

Applicants for Election and Transfer

THE MEMBERSHIP COMMITTEE at its meeting on 9th September 1976 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.

Meeting: 9th September 1976 (Membership Approval List No. 225)

GREAT BRITAIN AND IRELAND

CORPORATE MEMBERS

Transfer from Graduate to Member

BROWN, Andrew Duncan. *Sevenoaks, Kent.*
CUMNER, Clive Stewart. *Rudgwick, Sussex.*
ELLEN, Martin Trevor. *St. Albans, Herts.*
ELLISON, Louis Frederick. *Huddersfield, Yorkshire*

Direct Election to Member

DINI, Giovan Battista. *Erith, Kent.*
LOWNE, Alan John. *Chesham, Buckinghamshire.*

NON-CORPORATE MEMBERS

Direct Election to Graduate

LLOYD, Alan. *Ashford, Middlesex.*
HORNBY, Robert Maurice. *Northwich, Cheshire.*
LEE, Sik-Fun. *Glenrothes, Fife, Scotland.*
NIKIFOROS, Demosthenes Nicholas Jean. *Leatherhead, Surrey.*
SITHAMPARANATHAN, Manickham. *London.*

Direct Election to Associate Member

EBDON, Barry Edward. *London.*
HANKINS, Anthony. *Glossop, Derbyshire.*
HORLOCK, Colin Thomas. *Basingstoke, Hampshire.*

Transfer from Student to Associate

OBEYESEKERA, Piyasena. *London.*

Direct Election to Associate

HOLMES, Edward Ernest. *Falmouth, Cornwall.*

STUDENTS REGISTERED

JOHNSON, Patrick Gordon. *Exeter, Devon.*

OVERSEAS

CORPORATE MEMBERS

Transfer from Graduate to Member

FASHOLA, Victor Kehinde. *Lagos, Nigeria.*

NON-CORPORATE MEMBERS

Direct Election to Associate Member

CHENG, Kit Kwan. *Hong Kong.*
CHEUNG, Yee To. *Hong Kong.*
FUNG, Yim Ming Paul. *Hong Kong.*
MBELU, Bennett Oduche. *Ogidi, Nigeria.*

Transfer from Student to Associate Member

CHEN, Chin Min. *Perak, Malaysia.*

STUDENTS REGISTERED

GOH, Han Khoon. *Singapore.*
GUNSENA, Chaudra Kumary. *Kandy, Sri Lanka.*
KOU, Yook Fat. *Penang, West Malaysia.*
LIAW, Wee Hian. *Singapore.*
MOK, Keng Cher. *Singapore.*
RAYNAYAKE, Mudiyanselage Anjali Kumari. *Kandy, Sri Lanka.*
TAY, Kiam Kan. *Singapore.*
WIJESISI, Piyaseeli. *Trincomalee, Sri Lanka.*

STANDARD FREQUENCY TRANSMISSIONS

(Communication from the National Physical Laboratory)

JULY 1976	Deviation from nominal frequency in parts in 10^{10} (24-hour mean centred on 0300 UT)	Relative phase readings in microseconds NPL—Station (Readings at 1500 UT)	
	Droitwich 200 kHz	*GBR 16 kHz	†MSF 60 kHz
1	−0·1	697·3	611·8
2	−0·2	697·2	611·8
3	−0·3	697·1	611·7
4	−0·3	697·0	611·7
5	−0·4	697·2	611·7
6	−0·4	696·9	611·7
7	−0·2	697·2	611·8
8	−0·2	697·2	611·8
9	−0·2	697·3	611·6
10	−0·2	697·2	611·7
11	−0·2	697·2	611·6
12	−0·2	697·4	611·5
13	−0·2	697·3	611·7
14	−0·2	697·3	611·8
15	−0·1	697·3	611·8
16	−0·1	697·3	611·9
17	−0·1	697·3	611·9
18	−0·1	697·1	611·8
19	−0·2	697·3	612·0
20	−0·1	697·0	611·9
21	−0·1	697·3	611·9
22	−0·1	697·3	611·9
23	−0·1	696·8	611·9
24	−0·1	696·9	611·9
25	−0·1	697·1	611·9
26	−0·1	696·7	611·7
27	−0·2	696·8	611·9
28	−0·1	696·8	611·9
29	−0·1	696·6	611·8
30	−0·2	696·7	611·8
31	—	696·5	611·8

AUGUST 1976	Deviation from nominal frequency in parts in 10^{10} (24-hour mean centred on 0300 UT)	Relative phase readings in microseconds NPL—Station (Readings at 1500 UT)	
	Droitwich 200 kHz	*GBR 16 kHz	†MSF 60 kHz
1	—	696·8	611·8
2	−0·1	696·5	611·8
3	−0·1	696·4	611·6
4	−0·1	696·5	611·4
5	−0·1	696·6	611·6
6	−0·1	696·5	611·7
7	—	—	611·7
8	—	—	611·7
9	−0·2	696·8	611·7
10	−0·1	696·2	611·5
11	−0·2	695·9	611·5
12	−0·2	696·0	611·7
13	−0·2	696·1	611·7
14	−0·2	—	611·7
15	−0·2	696·1	611·7
16	−0·2	695·9	611·5
17	−0·2	695·9	611·7
18	−0·2	695·8	611·7
19	−0·2	695·9	611·7
20	−0·2	695·7	611·7
21	−0·2	—	611·7
22	−0·2	—	611·5
23	−0·2	695·7	611·7
24	−0·2	695·4	611·5
25	−0·2	695·3	611·5
26	−0·2	695·4	611·5
27	−0·2	695·3	611·7
28	−0·2	695·2	611·7
29	−0·2	695·4	611·5
30	−0·2	695·3	611·7
31	−0·2	695·8	611·7

All measurements in terms of H-P Caesium Standard No. 344 agrees with the NPL Caesium Standard to 1 part in 10^{11}.

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

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

Forthcoming Institution Meetings

London Meetings

Tuesday, 9th November

JOINT IEE/IERE MEDICAL AND BIOLOGICAL ELECTRONICS GROUP IN ASSOCIATION WITH THE BIOLOGICAL ENGINEERING SOCIETY

Physiology for engineers—1

Botany Theatre, University College London 6 p.m. (Tea 5.30 p.m.)

Further details to be announced.

Wednesday, 17th November

EDUCATION AND TRAINING GROUP

Colloquium on TECHNICIANS IN THE ELECTRONICS INDUSTRY

Royal Institution, Albemarle Street, London W1, 2.30 p.m. Advance registration necessary. For further details and registration forms, apply to Meetings Officer, IERE

Thursday, 18th November

COMMUNICATIONS GROUP

Compatible noise reduction in stereo broadcast systems

By Dr. A. R. Bailey (*University of Bradford*)

IERE Lecture Room, 6 p.m. (Tea 5.30 p.m.)

Wednesday, 1st December

JOINT COMPONENTS AND CIRCUITS GROUP AND INSTITUTE OF PHYSICS

Colloquium on INTEGRATED INJECTION LOGIC.

Royal Institution, Albemarle Street, London W1, 10.00 a.m. Advance registration necessary. For further details and registration forms, apply to Meetings Officer, IERE.

Thursday, 2nd December

AUTOMATION AND CONTROL SYSTEMS GROUP

Automated intelligence for an inter-stellar probe

By Dr. W. F. Hilton

IERE, Lecture Room, 6 p.m. (Tea 5.30 p.m.)

Tuesday, 7th December

ELECTRONICS PRODUCTION TECHNOLOGY AND AEROSPACE, MARITIME AND MILITARY SYSTEMS GROUPS

Colloquium on TESTABILITY AND TESTING TECHNOLOGY

Royal Institution, Albemarle Street, London W1, 10 a.m. Advance registration necessary. For further details and registration forms, apply to Meetings Officer, IERE.

Tuesday, 7th December

JOINT IERE/IEE MEDICAL AND BIOLOGICAL ELECTRONICS GROUP IN ASSOCIATION WITH THE BIOLOGICAL ENGINEERING SOCIETY

Physiology for engineers—2

Botany Theatre, University College London 6 p.m. (Tea 5.30 p.m.) Further details to be announced.

Wednesday, 8th December

JOINT IERE/IEE COMPUTER GROUP

Colloquium on FAULT TOLERANT COMPUTER SYSTEMS

IERE Lecture Room, 2 p.m.

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

Thames Valley Section

Thursday, 11th November

Facsimile—a review

By M. S. Bowden and M. J. Malster (*Rank Xerox*)

Caversham Bridge Hotel, Reading, 7.30 p.m

Wednesday, 8th December

The automated warehouse

By A. St. Johnston (*Vaughan Systems and Programming*)

Caversham Bridge Hotel, Reading, 7.30 p.m

East Anglian Section

Thursday, 18th November

Golay triple error correction code

By Alan Croft (*Plessey*)

Civic College, Rope Walk, Ipswich, 6.30 p.m. (Tea 6 p.m.)

Thursday, 25th November

Viewdata—a Post Office interactive information medium for the general public

By S. Fedida (*Post Office Research Establishment, Martlesham*)

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

Thursday, 2nd December

Brain and body scanners

By A. McAtamney (EMI)

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

South Western Section

Wednesday, 24th November

CEI LECTURE

The Engineer, Technology and Society

By J. E. Allen (*Hawker Siddeley Aviation*)

Department of Chemistry, University of Bristol, 7 p.m.

Further details to be announced.

Tuesday, 2nd December

JOINT MEETING WITH IEE

Technology in the service of the police

By D. Wyeth (*Dolby Laboratories Ltd.*)

Room 2E3. 1, University of Bath, 6 p.m. (Tea 5.30 p.m.)

Southern Section

Wednesday, 10th November

The Dolby noise reduction system

By P. Plunkett (*Dolby Laboratories*)

Synopsis: The Dolby A and Dolby B systems will be demonstrated and described, explaining how the classical problems associated with audio noise reduction systems are overcome. Advances in tape recording techniques coupled with higher standards demanded by audiophiles have created the need for audio noise reduction. The requirements for a hi-fi noise reduction system will be laid down and conventional static and dynamic noise reduction systems are described along with their advantages and disadvantages. The professional Dolby A and simpler Dolby B systems will be demonstrated and described explaining how the problems associated with more conventional approaches are overcome.

Electronics (South) Building, Southampton University, 7 p.m.

Friday, 12th November

Electronic ignition

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

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

Wednesday, 17th November

JOINT MEETING WITH IEE

Universal adaptor for computer peripherals

By D. M. Taub (*IBM*)

Southampton University, 6.30 p.m. (Tea 5.45 p.m.)

Wednesday, 24th November

Helium speech

By Dr. N. G. Kingsbury (*Marconi Space and Defence Systems*)

Synopsis: When a diver breathes a mixture of oxygen and helium in order to avoid the problems of nitrogen narcosis, his speech suffers a 'Donald Duck' type of distortion due to the increased sound velocity in the gas. At great pressures when the mixture is almost pure helium, the speech becomes practically unintelligible.

Methods of electronically processing or unscrambling the speech to improve its intelligibility will be discussed with particular reference to the technique of using digital storage registers to expand small segments of the input so as to reduce the formant frequencies by a manually selectable factor while leaving the larynx excitation rate unaltered. This results in relatively natural sounding speech with high intelligibility from the processor output. The formant reduction factor may be continuously varied between unity and 3½, to enable correct operation to depths in excess of 2000 feet. A demonstration will be given of the newest Marconi processor which has a power consumption of only 1 watt and will run off its internal rechargeable battery for up to 20 hours.

Consideration will be given to signal-to-noise ratio and frequency response requirements when helium speech from an operational diving environment is being processed. Solutions to the problems of communication with a diving bell along 2000 feet of umbilical cable which is carrying both power and communication circuits will be discussed. In the event of the umbilical cable being severed, it is desirable to have a through-water communication link and means of solving the additional problems of this requirement will be suggested.

Room ABO-11, Portsmouth Polytechnic, 7.30 p.m.

Tuesday, 30th November

Speech synthesis

By J. N. Holmes (*Joint Speech Research Unit*)

Lecture covers the following: How human speech is produced—vocal cords, air turbulence, acoustic resonances; relationships between linguistic units of message and acoustic signal; methods of modelling human mechanism electrically; applications—saving in information rate for speech transmission, voices from machines, research in phonetics and linguistics.

School of Signals, Blandford Camp, 6.30 p.m. (Tea 6 p.m.)

Wednesday, 1st December

JOINT MEETING WITH IEE

Developments in power electronics

By Dr. J. K. Hall, (*University of Loughborough*)

H.M.S. *Collingwood*, Fareham, 6.30 p.m. (Tea 5.45 p.m.)

Wednesday, 8th December

Electronic ignition

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

Lecture Theatre F, University of Surrey, 7 p.m.

Beds & Herts Section

Tuesday, 9th November

New semiconductor devices

By C. S. den Brinker (*Mackintosh Consultants Co.*)

Synopsis: In spite of the rather prolonged world recession, several new approaches to integrated circuit manufacture have recently become established. This is complemented by a series of new circuit technologies. If history does repeat itself, then at this moment we appear to be at the verge of another major innovative cycle, similar to that of the early sixties when the integrated circuit was introduced. These new technologies will be discussed, and it will be shown that some of the sharp divisions that were prevalent at one time are now rapidly beginning to disappear. Technologies to be described include integrated injection logic, d-mos, v-mos, the latest results in bipolar processes, as well as some of the recent advances in linear professional designs.

Luton College of Technology, 7.45 p.m.

Kent Section

Thursday, 11th November

Microprocessors

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

Further details to be announced.

Thursday, 9th December

Electronics on Saturday

By Dr. K. J. Dean (*Principal, S. E. London Technical College*)

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

West Midlands Section

Tuesday, 16th November

JOINT MEETING WITH IEE

Area traffic control in Coventry

By D. J. Clowes (*West Midlands Traffic Authority*)

To be followed by a visit to the Traffic Control Centre

Lanchester Polytechnic, Coventry, 6.30 p.m. (Tea 5.45 p.m.)

Monday, 6th December

JOINT MEETING WITH IEE

Electronic techniques in archaeology

By Professor E. T. Hall (*University of Oxford*)

North Staffordshire Polytechnic, Beaconside, 7 p.m. (Tea 6.30 p.m.)

South Midlands Section

Wednesday, 8th December

Global communications from land line to satellite

By D. W. Weedon, (*Cable and Wireless*)

Majestic Hotel, Park Place, Cheltenham, 7.30 p.m.

East Midlands Section

Tuesday, 9th November

JOINT MEETING WITH IEE

Ceefax—a new form of broadcasting

By D. T. Wright (*BBC Research Department*)

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

North Eastern Section

Monday, 6th December

JOINT MEETING WITH IEE

Dynamic ship positioning

By R. Bond (*G.E.C.*)

Merz Court, University of Newcastle, 6.15 p.m. (Tea 5.30 p.m.)

North Western Section

Thursday, 18th November

Integrated circuits in the modern world

By a speaker from *Ferranti, Hollinwood*

Renold Building, UMIST, 6.15 p.m. (Light refreshments available before the meeting)

Thursday, 16th December

Engineering in medicine

By Dr. J. A. Hewer (*Middlesex Hospital*)

Renold Building, UMIST, 6.15 p.m. (Light refreshments available before the meeting)

South Wales Section

Wednesday, 10th November

ANNUAL GENERAL MEETING

Followed by: The use of ultrasonics in medicine

By Dr. P. N. T. Wells, (*Bristol General Hospital*)

Room 112, Department of Physics, Electronics and Electrical Engineering, UWIST, Cardiff, 6.30 p.m. (Tea 5.30 p.m.)

Wednesday, 17th November

JOINT MEETING WITH INSTITUTE OF PHYSICS

Liquid crystals

By I. Shanks (*RSRE*)

University College, Swansea, 6 p.m. (Tea 5.30 p.m.)

Wednesday, 8th December

University television services

By Dr. P. Whitaker, (*University of Birmingham*)

Room 112, Department of Physics, Electronics and Electrical Engineering, UWIST, Cardiff, 6.30 p.m. (Tea 5.30 p.m.)