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## The Next Stage in Television?

NUMEROUS television wire broadcasting systems have been set up in Great Britain over recent years, and indeed figures show that approximately 1·9M or 11% of homes receive the television broadcasts of the BBC and Independent Television by this means, the service usually including several radio channels. The licences issued to the operators of relay networks by the Post Office have limited the service offered to programmes normally receivable in the areas, and even such technically feasible 'bonuses' as the programmes of adjacent Independent Television areas or of European stations have not been permitted.

During the past year, however, the Minister of Posts and Telecommunications has authorized five individual relay operators to set up experimental stations to originate and transmit programmes by cable. These are in the very diverse areas of Greenwich, Sheffield, Bristol, Swindon and Wellingborough and the numbers of subscribers range from 6,000 in Wellingborough up to 31,000 in Sheffield. One of the difficulties of these experimental operations, from the point of view of the owners of the networks, is that they have been made wholly responsible for financing the operations without the opportunity of meeting the costs by increased charges to subscribers, from advertising, or through grants or subsidies from a third party, e.g. a local authority.

The Cable Television Association has recently put forward proposals for adding further channels to the ones already indicated and envisages using 'pay television' for some of the channels as a means of financing the services. The Association has now also proposed the setting up, by the Minister, of a Cable Television Council to encourage and control development in the public interest.

The whole question of providing more television channels is a complicated one into which several factors enter. First and foremost is the basic problem of fitting additional channels into the available radio frequency spectrum, while there is a growing feeling that the 'environmental pollution' of the countryside by hundreds of transmitter masts and the urban scene by millions of domestic television aerials must be reduced. This means that some other form of distribution is required, but in looking at these problems from a realistic engineering standpoint, maximum utilization of any proposed system seems imperative.

Cable television under the present system is really only feasible within areas of reasonably high population density—services to outlying country viewers must presumably continue to rely on 'over the air' transmissions. The use of satellites directly beaming transmissions to outlying areas is a solution which is especially attractive in many parts of the world, but from cost-effective considerations one would wish to provide the maximum possible number of viewers. While s.h.f. satellite transmissions have the potential for a vast number of channels, their use for local television programmes would be a rather wasteful use of this capability.

Technical advances have made possible extremely wide bandwidth transmissions either by overmoded millimetre waveguide or by optical fibre and it is readily conceivable that these could be connected to individual homes, although the development of suitable equipment and its full exploitation puts this perhaps two decades away. There have been for several years proposals along these lines which envisage a high capacity transmission link, capable of operating in both directions with a multitude of facilities, some already provided by other, often very different techniques, some completely new. The terms 'electronic grid' favoured by Lord Mountbatten, Chairman of the National Electronics Council, and 'wired city' current in the United States, for these technically challenging proposals have embraced, for instance, conventional television and radio, telephone (audio and video), domestic meter reading, facsimile transmission of newspapers, and direct linkage to and from stores or stockbrokers; the opportunities for a subscriber's participation in central activities seem almost limitless, given the necessary hardware and demand.

Both in Parliament, in a debate on Independent Television last July, and outside, there is controversy over the implications of the present modest cable television experiments which are with us and clearly could be capable of rapid expansion. Would these commit the country for the foreseeable future to a diversified system of distribution, diversely operated, and thereby put even further into question the electronic grid/wired city concepts which must involve an essentially centrally operated enterprise? The latter will certainly need a much longer time to achieve its much wider aims—and on this basis the apparent 'second best' of separate, local, cable television networks is the more practical if the social and political decision is made that more channels in the early future should be the next stage in the development of television.

# Contributors to this issue

**Dr. R. A. Waldron** (Fellow 1961, Member 1957) has recently completed a three-year Principal Research Fellowship at the Post Office Research Centre, Martlesham Heath, during which he has been mainly concerned with work in connection with trunk waveguide communications. On October 13th he took up the post of Head, with the title of Director of Studies, of the School of Mathematics and Computer Science at Ulster College, The Northern Ireland Polytechnic.

Dr. Waldron graduated from the University of Cambridge in 1951 and joined the Baddow Research Laboratories of the Marconi Company, where he worked mainly on microwave theory. During this period he contributed numerous papers on waveguide subjects to the Institution's Journal for several of which he received premiums. In 1968 he was awarded the degree of Sc.D. from Cambridge University on the basis of this and other published work. From 1968 to 1971 Dr. Waldron held a research appointment at the Lincoln Laboratories of the Massachusetts Institute of Technology, working on surface-acoustic-wave-theory. His publications also include the expression of some unorthodox views on relativity theory, on which he has spoken at seminars in several universities and elsewhere.

On the formation of the East Anglian Section in 1965, Dr. Waldron was appointed Programme Secretary, a position he held until 1968, and which he resumed in 1971; he has been Honorary Secretary of the Section Committee for the past year.

**Dr. L. F. Turner** served an apprenticeship with AEI, during which time he gained a Higher National Certificate. He then studied at the University of Birmingham and obtained his Ph.D. degree. Following a period as a Research Fellow in the Department of Electronic and Electrical Engineering at Birmingham, Dr. Turner went to Standard Telecommunication Laboratories Limited, and for the next two years carried out research on various aspects of communications and data transmission. He subsequently joined the staff of Imperial College as a Lecturer in Communications in the Department of Electrical Engineering.

**Dr. G. B. Morgan** obtained the degrees of B.Sc. and Ph.D. at the University College of Swansea, Wales, where his researches were concerned with the high voltage avalanche breakdown of electronegative gases. From 1964 to 1968 he was employed at the Royal Radar Establishment, Malvern, England, working initially on the *Q* switching of high power, solid-state lasers and lidar, and subsequently on the microwave side of a high power, multi-function, search radar. From 1968 to 1970 he worked at RCA, Burlington, Mass., USA, where he was concerned with electro-optic communication and radar systems. Dr. Morgan at present lectures in electronics at the University of Wales Institute of Science and Technology, Cardiff, and his research interests are in avalanche diode amplifiers and pulse compression.

**Mr. K. W. Stanley** obtained his technical training with the Royal Navy at H.M.S. *Collingwood* and subsequently at Acton Technical College. He entered industry with EMI Engineering Development Ltd. in 1947 and worked on the development of components for use in radio and television receivers. In 1956 he joined his present company, Mullard Ltd., as a Product Engineer concerned with ferrite materials. He is now the Senior Product Engineer responsible for the company's ranges of linear and non-linear resistors.

**Dr. J. K. Stevenson** has recently moved to the Polytechnic of the South Bank, London, as a Lecturer in the Department of Electronics and Electrical Engineering. He was previously with Marconi Space and Defence Systems. A fuller note on his career was published in the August 1973 issue of the *Journal*.

**Mr. J. R. Edwards** has held a Senior Lectureship at the Royal Military College of Science since 1969. Details of his career were given in the September 1973 issue of the *Journal* when a companion paper was published.

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# Data compression techniques as a means of reducing the storage requirements for satellite data: a quantitative comparison

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*Based on a paper presented at the Joint IERE-IEE-IEEE-RTS-SMPTE Conference on Video and Data Recording held in Birmingham on 10th to 12th July 1973.*

**SUMMARY**

This paper reports on an investigation into the application of data compression techniques as a means of reducing the 'on-ground' data storage requirements that are associated with many space research programmes. The paper presents and compares quantitatively various compression techniques based on the Shannon–Fano, 'run-length' and Hadamard transformation methods of source encoding. The compression ratios obtained when applying the techniques to actual satellite data are given and some new basic theory relating to 'run-length' encoding is presented.

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## 1 Introduction

In many research applications the information obtained by satellites is transmitted back to earth where it is then stored for subsequent processing and use. Even today, the quantities of data involved are so large that the necessary 'on-ground' storage capacity is becoming excessive, and the situation will certainly become more serious with the development of satellites that transmit information back to earth at increasingly high rates.

The investigation reported in this paper was part of a general study supported by the European Space Operations Centre of the European Space Research Organization into techniques for the 'on-ground' compression and archiving of satellite data. The object of the data compression part of the study was to investigate compression techniques and develop associated software packages that could be used in conjunction with a general-purpose computer to compress data in readiness for storage in an archive. The general-purpose computer and associated software packages can be considered as a data compression unit.

In the study data from the ESRO 1 satellite were used and five different data channels were considered. Figure 1 shows the data obtained from one of the channels during part of the satellite's orbit around the Earth. The input to the compression unit is the satellite data which are received on the ground as samples that have been quantized (in the satellite) and represented by 8-bit p.c.m. words. The output from the compression unit is a sequence of binary digits which represents the input data sequence, and which can be converted back by a decoding process into the original input data sequence. The object of the data compression is to represent an input data sequence by an output data sequence that contains as few binary digits as possible. The data compression ratio is generally defined as:

$$\text{data compression ratio} = \frac{\text{number of binary digits in input sequence}}{\text{number of binary digits in output sequence}}$$

In this paper various aspects of data compression are considered. In Section 2 the Shannon–Fano encoding technique of data compression is outlined briefly, and the results obtained when applying the technique to ESRO 1 satellite data are given. Some of the fundamental difficulties associated with the application of the Shannon–Fano technique are discussed and a modification of the application is developed which, to some extent, overcomes the difficulties. The results of applying the modified method to satellite data are presented.

In Section 3 the basic idea of data compression by 'run-length' encoding is considered and the results of applying various versions of the basic run-length scheme to the compression of ESRO 1 data are given. Some new basic theory relating to 'run-length' encoding is developed and presented in Appendix 2.

In Section 4 the Hadamard transformation technique, which has been used recently as a means of picture bandwidth compression,[1] is applied to ESRO 1 data. In Section 5 some general conclusions are drawn and the data compression techniques discussed in this paper are compared with other known compression techniques.

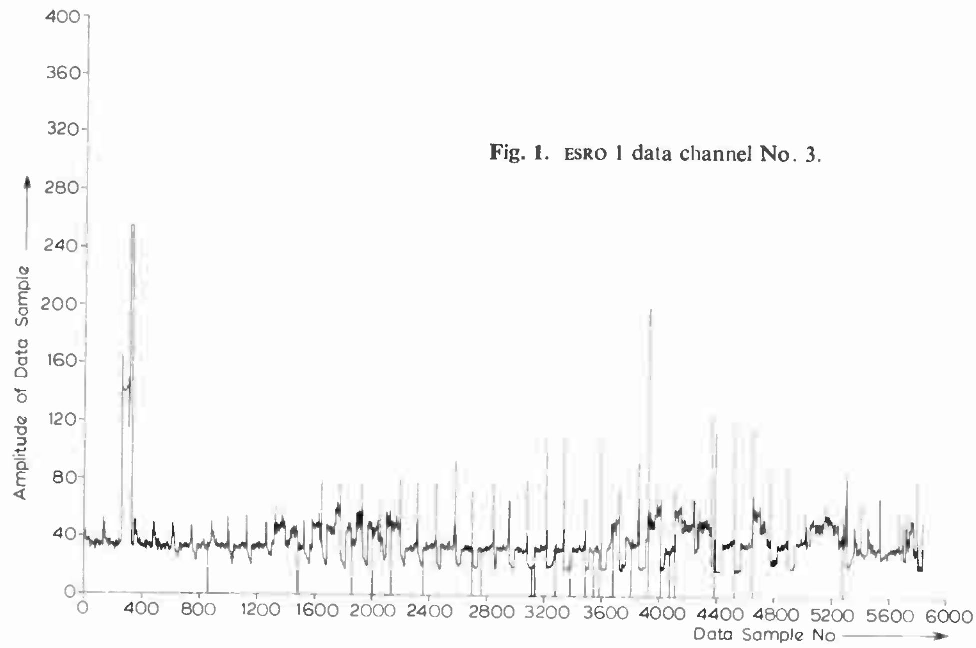


**Fig. 1.** ESRO 1 data channel No. 3.

## 2 Shannon–Fano Encoding as a Means of Data Compression

### 2.1 *The Basic Idea of Shannon–Fano Encoding*[2]

A simple technique for encoding the output of a data source is that due to Shannon and Fano.† In applying this technique each source output, generally termed a 'source symbol' (an 8-bit word in the case of ESRO 1 data) is encoded using a sequence of binary digits and the shortest-length sequences are used for the source outputs that occur most frequently. By encoding in this way the average number of binary digits used to represent a source output symbol is reduced to a minimum and data compression is thereby achieved in certain cases. The rules for constructing Shannon–Fano codes are given in Appendix 1, and a simple example illustrating the application of the rules is shown in Fig. 2.

### 2.2 *Application of Shannon-Fano Encoding to ESRO 1 Data*

The Shannon–Fano encoding technique was applied to ESRO 1 data. In order to determine the effectiveness of the technique when applied to a range of data statistics, and in order to determine the sensitivity of the technique to changes in data statistics, five different data channels were used. In applying the technique the following four operations were performed on each of the five data channels:

(i) The probability $P(x_i)$ associated with each possible output amplitude, $x_i$, (an 8-bit p.c.m. word in the case of ESRO 1 data) was determined by inspection of the data and the appropriate Shannon–Fano code book was constructed using the rules presented in Appendix 1.

(ii) The zero-order entropy, defined by

$$H = -\sum_{i=1}^{i=n} P(x_i)\log_2 P(x_i) \quad \text{was computed.}$$

(iii) The data on the channel were encoded using the Shannon–Fano code book and the compression ratio was computed.

(iv) After the data had been encoded, decoding was performed and the original data were thus obtained. This operation was performed in order to determine the decoding time.

### 2.3 *Results of Applying the Shannon–Fano Technique: Discussion of Results*

The results obtained when applying the Shannon–Fano encoding technique are summarized in Table 1, which gives the zero-order entropy, the compression ratio and the decoding time for each data channel. Inspection of the Table shows that the compression ratios range from 1·81:1 to 1·44:1 and it can also be seen from the Table that the compression ratios that are achieved are very close to the maximum ratios that the zero-order entropies suggest is possible. The conclusion to be drawn from this is that, for the channels being considered, the Shannon–

| Source Symbol | Probability $P(x)$ | Code Word |
|---|---|---|
| $X_1$ (=00) | ·81 | 0 |
| $X_2$ (=01) | ·09 | 1 0 |
| $X_3$ (=10) | ·09 | 1 1 0 |
| $X_4$ (=11) | ·01 | 1 1 1 |

Average code-word length = 1·29

$$\text{Data compression} = \frac{2}{1\cdot 29} = 1\cdot 55$$

**Fig. 2.** Example of Shannon–Fano encoding.

† In the investigation of compression techniques the Shannon–Fano scheme of compression was used, rather than the Huffman scheme.[3] Although the Huffman encoding scheme is theoretically slightly superior to the Shannon–Fano scheme, the Shannon–Fano scheme is considerably easier to program and, furthermore, it is found in practice that, more often than not, the two schemes result in identical compression ratios. It is also found that when the compression ratios do differ, the difference is negligible.

**Table 1**

Shannon–Fano coding applied to individual source symbols

| Data channel No. | Zero-order entropy (bits/symbol) | Compression ratio | Maximum compression ratio based on zero-order entropy | Decoding time (ms/sample) |
|---|---|---|---|---|
| 1 | 5·35 | 1·48 | 1·49 | 0·90 |
| 2 | 4·96 | 1·60 | 1·61 | 0·80 |
| 3 | 5·26 | 1·50 | 1·52 | 0·90 |
| 4 | 5·46 | 1·44 | 1·46 | 0·93 |
| 5 | 4·35 | 1·81 | 1·83 | 0·79 |

Fano technique is very efficient when applied as a means of encoding single source symbols. The closeness of the compression ratios to the values of the compression ratios based on the zero-order entropy should not, however, be taken to mean that the Shannon–Fano code, when applied to single source symbols, has produced compression ratios that are close to the best that could be achieved if longer blocks of source symbols were encoded. The zero-order entropy is based on the assumption that successive data symbols (samples) are statistically independent and this assumption is not in fact valid. The successive data samples obtained from the satellite are statistically dependent, and the actual entropy of the source, which takes this dependence into account, is less than the zero-order entropy. If the Shannon–Fano technique had been applied so that blocks of source symbols, rather than individual source symbols, had been encoded then the compression ratios obtained would have been better than those of Table 1. This was not done on account of the size of the Shannon–Fano code book that would have been necessary. The Shannon–Fano code book, which has to be stored, contains one code word for each different source symbol and, since ESRO 1 data samples are in the form of 8-bit p.c.m. words, this means that the Shannon–Fano code book contains 256 code words for single-symbol encoding and contains $256^2 = 65\,536$ code words for symbol-pair encoding.

### 2.4 *A Modification of the Application of the Shannon–Fano Encoding Scheme*

It can be seen from the rules presented in Appendix 1 for the construction of the Shannon–Fano code book, and from the example of Fig. 1 that the code-word structure is very dependent on the probabilities with which the various source symbols occur. From this it is clear that the Shannon–Fano technique is likely to be sensitive to any changes in the probabilities associated with the input data samples. These changes may take the form of variations of the probabilities as a function of time, or variations of the probabilities from one channel to another if a single Shannon–Fano code is to be used for the compression of data on more than one channel. With ESRO 1 data it was found, for example, that if the Shannon–Fano code book for data channel No. 1 was applied to data channel No. 5 then the data were expanded rather than compressed.

In an attempt to overcome these difficulties, a modification of the application of the Shannon–Fano encoding scheme was developed and tested. The modification was simple and consisted of applying the Shannon–Fano code to differences between adjacent data samples, rather than to the samples. (This is an established method oı removing non-stationarities from random data.[4]) The basic idea underlying the modification is that, for anything other than rapidly-fluctuating data, there will be a considerable degree of similarity between adjacent data samples.

### 2.5 *Results Obtained from Application of Modified Method: Discussion of Results*

The results obtained when applying the modified version of the Shannon–Fano encoding scheme are given in Fig. 3 and Table 2. Figure 3 shows that the basic idea underlying the modification is justified for the channels considered, since there is a very considerable degree of similarity between the 'difference probability density functions' of the various channels. The Figure shows that for all channels the high probability differences are those in the range $\pm 6$ and that larger differences are of very much smaller probability.

The zero-order entropies of the differences, the compression ratios and the decoding times for the five data channels are given in Table 2. It can be seen from Tables 1 and 2 that the zero-order entropies of the differences are less than the zero-order entropies associated with the individual data samples and that the application of Shannon–Fano encoding to differences results in a greater data compression than that achieved when applying the technique directly to the source symbols. It can also be seen that the compression ratios achieved by Shannon-Fano encoding of the differences are very close to the maximum compression ratios that the zero-order entropies of the differences indicates as possible. The actual compression ratios obtained ranged from a maximum of 2·46:1 for channel No. 1 to a minimum of 2·04:1 for channel No. 4.

The Shannon–Fano code book developed for the amplitude differences of channel 1 was used to encode the amplitude differences of channel 5. It was found when channel 5 was encoded in this way that the compression ratio deteriorated from 2·37:1 to 2·01:1. The deterioration, although still present, is much less than that which occurred when the Shannon–Fano code was applied

**Table 2**

Shannon–Fano coding applied to amplitude differences

| Data channel No. | Zero-order entropy of differences (bits/symbol) | Compression ratio | Maximum compression ratio based on zero-order entropy | Decoding time (ms/sample) |
|---|---|---|---|---|
| 1 | 3·22 | 2·46 | 2·48 | 0·13 |
| 2 | 3·71 | 2·08 | 2·16 | 0·25 |
| 3 | 3·53 | 2·24 | 2·27 | 0·16 |
| 4 | 3·84 | 2·04 | 2·08 | 0·31 |
| 5 | 3·31 | 2·37 | 2·41 | 0·14 |

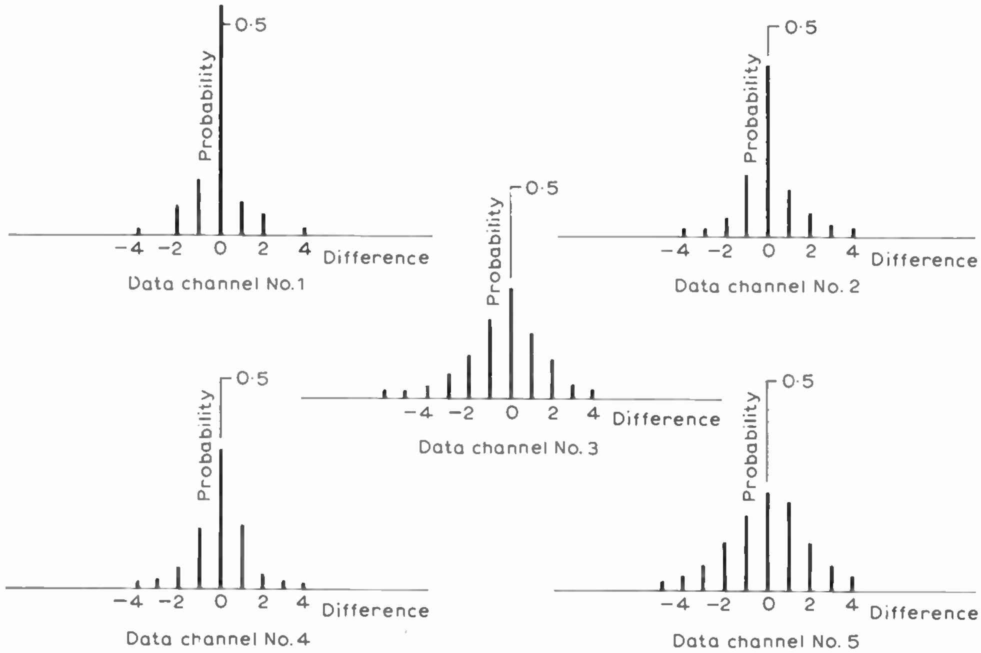


**Fig. 3.** Probability functions associated with amplitude differences between adjacent data samples.

directly to the symbol amplitudes and this, therefore, shows that the modified technique is much less sensitive to variations in channel statistics.

## 3 'Run-Length' Encoding as a Means of Data Compression

### 3.1 *Some Initial Comments and an Explanation of the 'Run-Length' Encoding Technique*

The technique that has come to be known as 'run-length encoding' has proved successful[5, 6] as a means of reducing the bandwidth necessary for the transmission of television signals. This Section reports on an application of the technique to the compression of satellite data.

The basic idea of 'run-length' encoding is as follows: the data are examined and if a number of adjacent samples are of the same amplitude, or are the same to within certain acceptable limits, then a 'run' is said to have occurred. In order to transmit or store the data it is necessary only to given the amplitude of the first sample in the run and the length of the run—hence the name 'run-length encoding'. In the situation in which adjacent samples are considered to be the same only if they are exactly equal, the encoding is said to be 'run-length encoding with zero aperture'. In those situations in which adjacent samples are considered to be the same if they differ by less than a finite amount,† the encoding is said to be 'run-length encoding with a finite aperture'.

In this section of the paper three forms of the basic run-length encoding scheme are considered and the results obtained from the application of the techniques to the compression of ESRO 1 data are given. The effects of aperture size on the constitution of the run-lengths and the resulting effects on the obtainable compression ratios are also considered. In Appendix 2 some new basic theory relating to run-length encoding is presented.

† A consequence of the use of a non-zero aperture is that the reconstructed data are in error as compared with the original input data. The error, which is the difference between the reconstructed and original data, can be expressed as an r.m.s. value.

### 3.2 *Application of 'Run-Length' Encoding to ESRO 1 Data*

Three forms of the basic run-length encoding technique were used to compress ESRO 1 data. The techniques used are described below and the operations involved in the application of each method are outlined.

*Run-length encoding technique No. 1.* The first run-length encoding technique investigated was that in which the lengths of all the naturally occurring runs were encoded using a Shannon–Fano code. In applying the technique the first sample of each run was stored, together with the Shannon–Fano code word representing the length of the run. An expression for the theoretical compression ratio obtainable with this method is given as equation (15) of Appendix 2.

When applying encoding technique No. 1, the following operations were performed on each of the five data channels:

(i) The probability $P(l_i)$ associated with each possible run of length $l_i$ was determined by inspection of the data, and the appropriate Shannon–Fano code book was constructed.

(ii) The zero-order entropy of the run-lengths, defined by

$$H = -\sum_{i=1}^{i=i\,\max} P(l_i)\log_2 P(l_i),\ \text{was computed.}$$

(iii) The data were encoded and the compression ratio was computed.

*Run-length encoding technique No. 2.* Although run-length encoding technique No. 1 has the advantage of producing the highest data compression ratio of all run-length encoding schemes, it suffers from the disadvantage that when all naturally occurring run-lengths are permitted, it is possible for there to be a very large number of different run-lengths, and the associated Shannon–Fano encoding becomes complex. If, however, the

**Table 3** Natural runs broken down into sequences of permissible runs

| Length of natural run $l$ | Sequence of permissible runs into which natural run is broken down |
|---|---|
| 1 | 1 |
| 2 | 2 |
| 3 | 2,1 |
| 4 | 2,2 |
| 5 | 5 |
| 6 | 5,1 |
| 7 | 5,2 |
| 8 | 5,2,1 |
| 9 | 9 |
| 10 | 9,1 |
| 11 | 9 2 |
| . | . |
| . | . |
| . | . |
| 30 | 9,9,9,2,1 |

number of different run-lengths is limited then the encoding becomes relatively simple.

With run-length encoding technique No. 2 the number of different run lengths was limited. This was done by using a fixed set of permissible run lengths and breaking all natural run lengths down into sequences of permissible run lengths. In the investigation, an arbitrarily chosen set of permissible run lengths was used. The permissible run lengths were chosen to be 1, 2, 5 and 9; and Table 3 shows how naturally occurring runs of length $l$ were broken down into sequences of permissible run lengths.

In applying run-length encoding technique No. 2, the same operations were performed as when applying technique No. 1; but with the difference that the operations were applied to the permissible rather than to the naturally occurring run lengths.

*Run-length encoding technique No.* 3. With run-length encoding techniques Nos. 1 and 2, Shannon–Fano code words were used to represent the lengths of the runs. However, as mentioned earlier, the effectiveness with which Shannon–Fano codes compress data is very dependent on whether or not the data statistics are stationary. A possible alternative way of encoding the lengths of the runs, so that the encoding is insensitive to changes in data statistics, is to use fixed-length code words. Run-length encoding technique No. 3, which was a repeat of technique No. 2; except that fixed-length 2-bit words rather than variable length Shannon–Fano code words were used to specify run lengths, was used to test the effectiveness of this idea.

The theory of run-length encoding technique No. 3 is considered in Appendix 2 and an expression for the compression ratio is given as equation (10).

When applying each of the three previously mentioned run-length encoding techniques, apertures of sizes 0, $\pm 1$ and $\pm 7$ p.c.m. quantization levels were used. This was done in order to determine how the data compression ratio could be increased at the expense of the fidelity with which the data could be stored, and hence reconstructed.

### 3.3 *Results Obtained with Run-Length Encoding: Discussion of Results*

The results obtained when applying the three run-length techniques are given in Tables 4 to 6. The results presented in Table 4 were obtained when using a zero aperture, those presented in Table 5 when using an aperture of size $\pm 1$ p.c.m. quantization unit and those presented in Table 6 when using an aperture of size $\pm 7$ p.c.m. quantization units.

Examination of Table 4 shows that when using run-length encoding technique No. 1 the compression ratios varied from a maximum of 1·71:1 to a minimum of 1·12:1. The poorer compression ratios, which in fact are not at all good, were found to be due to the large number of runs of length 1 that occurred in the data. It was found, for example, that 69% of the runs in data channel No. 1 were of length 1 and that 79% of the runs in data channel No. 5 were of length 1.

It can be seen from Table 4 that run-length encoding techniques Nos. 2 and 3 give poorer compression ratios than technique No. 1. The reason for this is that in going to a fixed set of run lengths (and particularly with the short-length runs used) the long naturally occurring runs are destroyed. The consequence of this is that the compression ratios are reduced, since it is the encoding of long runs that leads to effective data compression. It can also be seen from Table 4 that run-length technique No. 3 yields the worst results. The reason for the particularly poor performance of this technique is again associated directly with the high proportion of runs of length 1. When applying the technique, runs of length 1 are encoded using ten binary digits—eight digits are used to signify the amplitude of the sample and two digits are used to indicate that the run is of length 1. From this it

**Table 4** Run-length encoding with zero aperture

| Run-length encoding technique No. | 1 | 2 | 3 |
|---|---|---|---|
| Channel No. | | | |
| 1 | c.r. = 1·71<br>r.m.s. = 0 | c.r. = 1·53<br>r.m.s. = 0 | c.r. = 1·26<br>r.m.s. = 0 |
| 2 | c.r. = 1·38<br>r.m.s. = 0 | c.r. = 1·27<br>r.m.s. = 0 | c.r. = 1·11<br>r.m.s. = 0 |
| 3 | c.r. = 1·15<br>r.m.s. = 0 | c.r. = 1·11<br>r.m.s. = 0 | c.r. = 1·02<br>r.m.s. = 0 |
| 4 | c.r. = 1·25<br>r.m.s. = 0 | c.r. = 1·16<br>r.m.s. = 0 | c.r. = 1·07<br>r.m.s. = 0 |
| 5 | c.r. = 1·12<br>r.m.s. = 0 | c.r. = 1·07<br>r.m.s. = 0 | c.r. = ·987<br>r.m.s. = 0 |

*Note:* In Tables 4, 5, 6, 9 and 10 'c.r.' denotes compression ratio and 'r.m.s.' denotes the r.m.s. error in the stored, and hence, reconstructed data. The r.m.s. errors are given in p.c.m. quantization levels and one p.c.m. quantization level corresponds to an error of 1/256 = 0·394% of the peak signal level.

**Table 5** Run-length encoding with aperture = ±1 p.c.m. quantization level

| Run-length encoding technique No. | 1 | 2 | 3 |
|---|---|---|---|
| Channel No. | | | |
| 1 | c.r. = 2·42<br>r.m.s. = ·48 | c.r. = 2·07<br>r.m.s. = ·65 | c.r. = 2·00<br>r.m.s. = ·65 |
| 2 | c.r. = 2·18<br>r.m.s. = ·58 | c.r. = 1·88<br>r.m.s. = ·74 | c.r. = 1·80<br>r.m.s. = ·74 |
| 3 | c.r. = 1·74<br>r.m.s. = ·57 | c.r. = 1·58<br>r.m.s. = ·87 | c.r. = 1·52<br>r.m.s. = ·87 |
| 4 | c.r. = 2·05<br>r.m.s. = ·56 | c.r. = 1·78<br>r.m.s. = ·77 | c.r. = 1·70<br>r.m.s. = ·77 |
| 5 | c.r. = 1·81<br>r.m.s. = ·59 | c.r. = 1·62<br>r.m.s. = ·89 | c.r. = 1·56<br>r.m.s. = ·89 |

**Table 6** Run-length encoding with aperture = ±7 p.c.m. quantization level

| Run-length encoding technique No. | 1 | 2 | 3 |
|---|---|---|---|
| Channel No. | | | |
| 1 | c.r. = 3·72<br>r.m.s. = 2·46 | c.r. = 2·78<br>r.m.s. = 2·49 | c.r. = 2·77<br>r.m.s. = 2·49 |
| 2 | c.r. = 3·26<br>r.m.s. = 2·65 | c.r. = 2·65<br>r.m.s. = 2·87 | c.r. = 2·60<br>r.m.s. = 2·87 |
| 3 | c.r. = 7·50<br>r.m.s. = 3·11 | c.r. = 4·48<br>r.m.s. = 3·00 | c.r. = 4·36<br>r.m.s. = 3·00 |
| 4 | c.r. = 3·91<br>r.m.s. = 2·42 | c.r. = 2·73<br>r.m.s. = 2·99 | c.r. = 2·73<br>r.m.s. = 2·99 |
| 5 | c.r. = 9·90<br>r.m.s. = 3·00 | c.r. = 4·94<br>r.m.s. = 2·76 | c.r. = 4·77<br>r.m.s. = 2·76 |

is clear that the data are expanded by a factor of 10/8 whenever a run of length 1 occurs, and hence it is necessary that a significant proportion of the runs should be long if effective compression is to be achieved.

Tables 5 and 6 show clearly that by increasing the aperture size higher compression ratios can be obtained. The basic reason for the improvement is that the use of a finite aperture leads to an increase in the occurrence of longer runs. From Table 5 it can be seen that when using an aperture of size ±1 p.c.m. quantization unit, which corresponds to an error of less than 0·4% of the peak signal amplitude level, encoding technique No. 1 yields compression ratios ranging from a maximum of 2·42:1 to a minimum of 1·74:1; and that encoding techniques Nos. 2 and 3 yield compression ratios ranging from a maximum of approximately 2:1 to a minimum of approximately 1·5:1. It is important to note that there is little difference between the compression ratios obtained with Techniques Nos. 2 and 3. Table 6 shows that for an aperture of size ±7 p.c.m. quantization units, which corresponds to an error of less than 3% of the peak signal amplitude, the actual r.m.s. errors are generally of the order of 3 p.c.m. units (approximately 1% of peak signal amplitude) and that with this size of aperture encoding technique No. 1 yields compression ratios ranging from a maximum of approximately 10:1 to a minimum of approximately 3:1. The compression ratios obtained with techniques Nos. 2 and 3 range from a maximum of approximately 5:1 to a minimum of approximately 2·6:1. Again it will be noted that there is little difference between the two encoding techniques. The highly important practical consequence of this is that encoding in which fixed-length code words are used is virtually as efficient as encoding in which variable-length Shannon–Fano code words are used. This means that in practice it would be sensible to use fixed-length encoding, since it is simpler to implement.

## 4 The Use of the Hadamard Transformation as a Means of Data Compression

### 4.1 *Basic Ideas and Theory Underlying the Use of the Hadamard Transformation as a Means of Data Compression*

Recently the suggestion has been made[1, 7, 8] that picture bandwidth compression can be achieved by using signal transformation techniques. This section of the present paper is devoted to a report on the application of the technique to the compression of ESRO 1 data.

The basic idea underlying the signal transformation technique of compression is that if the signal (satellite data in the present case) can be transformed into some new domain, and the transformation can be carried out in such a way as to remove redundancy, then the signals or data can be either transmitted or stored in the less redundant form that they have in the new domain, and compression can thereby be achieved. One specific transformation that has been suggested[1] as a possible means of picture bandwidth compression is the Hadamard transformation. The main advantage of the Hadamard transformation, as compared with other transforms such as the Fourier transform, is that it is easy to compute and apply.

The Hadamard transformation technique is based on the symmetric Hadamard matrix $[H(u, v)]$.

If $[f(x, y)]$ denotes an $N \times N$ array (matrix) of data elements then the Hadamard transformation $[F(u, v)]$ is obtained using the equation:

$$[F(u, v)] = [H(u, v)] \cdot [f(x, y)] \cdot [H(u, v)]. \qquad (1)$$

Equation (1) is a matrix equation in which $[F(u, v)]$ is an $N \times N$ matrix that represents the transformed data, $[H(u, v)]$ is an $N \times N$ Hadamard matrix and $[f(x, y)]$ is the original data matrix.

The original data can be reconstructed from $[F(u, v)]$ by carrying out the following matrix operation:

$$[f(x, y)] = \frac{1}{N^2} [H(u, v)] \cdot [F(u, v)] \cdot [H(u, v)]. \qquad (2)$$

Data compression can be achieved if the transformed data $[F(u, v)]$ is represented only by the important elements in the matrix, and all others are set to zero or some small value. If the compressed (coarsely quantized) transformed matrix is represented by $[\widehat{F(u, v)}]$, then, on

**Table 7**

Compression modes used when applying Hadamard 1

| Compression mode No | Encoding (quantization) scheme |
|---|---|
| 1 | use 10 bits to encode element $a$<br>use 3 bits to encode each of elements $b, c, d$ |
| 2 | use 10 bits to encode element $a$<br>use 2 bits to encode each of elements $b, c, d$ |
| 3 | use 10 bits to encode element $a$<br>use 5 bits to encode element $b$<br>use 2 bits to encode each of elements $c, d$ |
| 4 | use 8 bits to encode element $a$<br>use 3 bits to encode each of elements $b, c, d$ |

**Table 8**

Compression modes used when applying Hadamard 2

| Compression mode No. | Encoding (quantization) scheme |
|---|---|
| 1 | use 12 bits to encode element $a$<br>use 3 bits to encode each element $b, \ldots, p$ |
| 2 | use 12 bits to encode element $a$<br>use 3 bits to encode each element $b, f, e$<br>use 2 bits to encode each other element |
| 3 | use 12 bits to encode element $a$<br>use 2 bits to encode each element $b, \ldots, p$ |
| 4 | use 12 bits to encode element $a$<br>use 3 bits to encode each element $b, f, e$<br>Neglect all other elements |

performing the operation defined by equation (2), an estimate $[f|\widehat{(x, y)}]$ of the original data is obtained as follows:

$$[\widehat{f(xy)}] = \frac{1}{N^2}[H(u, v)] \cdot [\widehat{F(u, v)}] \cdot [H(u, v)]. \qquad (3)$$

## 4.2 *Application of the Hadamard Transformation Technique to the Compression of ESRO 1 Data*

Two versions of the basic Hadamard transformation technique were applied to the compression of ESRO 1 data. In the first version of the technique (called Hadamard 1) input data were broken down into blocks of 4 samples and each block was then arranged as a $2 \times 2$ matrix, $[f(x, y)]$. In the second version (called Hadamard 2) blocks of 16 data samples were used and these were arranged as a $4 \times 4$ matrix, $[f(x, y)]$.

The operations involved in the application of the two versions of the Hadamard technique were as follows:

Hadamard 1

(i) The data were broken down into blocks of 4 samples (let $x_1$, $x_2$, $x_3$ and $x_4$ denote the samples in a block).

(ii) The data were encoded using the Hadamard transformation

$$\begin{bmatrix} 1 & 1 \\ 1 & -1 \end{bmatrix} \cdot \begin{bmatrix} x_1 & x_2 \\ x_3 & x_4 \end{bmatrix} \cdot \begin{bmatrix} 1 & 1 \\ 1 & -1 \end{bmatrix} \rightarrow \begin{bmatrix} a & b \\ c & d \end{bmatrix}.$$

The matrix

$$\begin{bmatrix} x_1 & x_2 \\ x_3 & x_4 \end{bmatrix}$$

is called the 'data' matrix and the matrix

$$\begin{bmatrix} a & b \\ c & d \end{bmatrix}$$

is called the 'information' matrix.

(iii) A compression mode was selected and the information matrix was encoded (quantized) using the selected mode.

(iv) The data were reconstructed from the encoded information matrix using the transformation of equation (3), and the r.m.s. error was computed.

The four compression modes shown in Table 7 were tested with Hadamard 1.

Hadamard 2

(i) The data were broken down into blocks of 16 samples (let $x_1, \ldots, x_{16}$ denote the samples in a block).

(ii) The data were encoded using the Hadamard transformation

$$\begin{bmatrix} 1 & 1 & 1 & 1 \\ 1 & -1 & 1 & -1 \\ 1 & 1 & -1 & -1 \\ 1 & -1 & -1 & 1 \end{bmatrix} \cdot \begin{bmatrix} x_1 & \cdots & x_4 \\ \vdots & & \vdots \\ \vdots & & \vdots \\ x_{13} & \cdots & x_{16} \end{bmatrix} \cdot \begin{bmatrix} 1 & 1 & 1 & 1 \\ 1 & -1 & 1 & -1 \\ 1 & 1 & -1 & -1 \\ 1 & -1 & -1 & 1 \end{bmatrix} \rightarrow \begin{bmatrix} a & b & c & d \\ e & f & g & h \\ i & j & k & l \\ m & n & o & p \end{bmatrix}$$

(iii) and (iv) As for Hadamard 1.

The four compression modes shown in Table 8 were tested with Hadamard 2.

## 4.3 *Results of Applying the Hadamard Transformation Technique: Discussion of Results*

The results of applying the techniques denoted as Hadamard 1 and 2 are shown in Tables 9 and 10 respectively.

If the compression ratios and the r.m.s. error values of Table 9 are compared with the corresponding ratios and r.m.s. values given in Tables 5 and 6 they are seen to be significantly inferior. The r.m.s. errors are higher than those resulting with finite-aperture run-length encoding and the compression ratios are lower. In those cases in which the Hadamard technique gives compression ratios that are nearly equal to those obtained by run-length encoding, the r.m.s. errors are always higher.

It should be noted from Table 9 that compression mode 4 gives r.m.s. errors that are very much worse than those obtained with the other three compression modes. The reason for this is that in mode 4 the element $a$ of the information matrix was truncated to 8 bits, which covers the range 1 to 256. Inspection of the information matrices showed, however, that the element $a$ often had a value close to its maximum of $4 \times 256 = 1024$, and it therefore needs a 10-bit specification. The effect of limiting it to a maximum of 256 introduces large errors. As a general conclusion, it appears that it is necessary to use the full range for the representation of the element $a$.

Table 9 Results obtained from application of Hadamard 1

| Compression mode No. / Channel No. | 1 | 2 | 3 | 4 |
|---|---|---|---|---|
| 1 | c.r. = 1·68<br>r.m.s. = 32·2 | c.r. = 2·0<br>r.m.s. = 32·5 | c.r. = 1·68<br>r.m.s. = 31·98 | c.r. = 1·88<br>r.m.s. = 46·9 |
| 2 | c.r. = 1·68<br>r.m.s. = 14·4 | c.r. = 2·0<br>r.m.s. = 14·65 | c.r. = 1·68<br>r.m.s. = 14·16 | c.r. = 1·88<br>r.m.s. = 22·3 |
| 3 | c.r. = 1·68<br>r.m.s. = 4·96 | c.r. = 2·0<br>r.m.s. = 5·21 | c.r. = 1·68<br>r.m.s. = 4·96 | c.r. = 1·88<br>r.m.s. = 13·8 |
| 4 | c.r. = 1·68<br>r.m.s. = 13·28 | c.r. = 2·0<br>r.m.s. = 13·58 | c.r. = 1·68<br>r.m.s. = 13·13 | c.r. = 1·88<br>r.m.s. = 29·8 |
| 5 | c.r. = 1·68<br>r.m.s. = 3·91 | c.r. = 2·0<br>r.m.s. = 4·16 | c.r. = 1·68<br>r.m.s. = 3·89 | c.r. = 1·88<br>r.m.s. = 3·93 |

If the results given in Tables 9 and 10 are compared, it will be seen that, as a general rule, the use of longer blocks of data samples results in improved compression ratios. This improvement in compression ratio appears to be gained at the expense of larger r.m.s. errors, but it is felt, however, that by proper choice of compression mode encoding, the r.m.s. errors may be reduced somewhat. The structures of the compression modes used were chosen on an intuitive basis rather than as a result of analysis. With more work, it should be possible to present rules relating to the choice of compression mode strategy.

## 5 Some General Comments and Conclusions

In this paper various techniques of data compression have been examined, and the results obtained when applying the techniques to the compression of ESRO 1 satellite data have been presented and compared. It only remains to summarize the results and make some general comments.

From an inspection of Tables 1 to 10 it is clear that, in general, run-length encoding tends to give the best results, particularly when small apertures are used. If apertures are not used then the Shannon–Fano technique applied to amplitude differences tends to yield the best results.

Although it is desirable to be able to compare the techniques discussed in this paper with other known compression techniques this is a difficult task. The difficulty arises from the fact that the compression ratios that are obtainable with a given technique tend to be very dependent on the data statistics. In order to obtain a meaningful comparison it is necessary to be able to compare techniques when they are applied to the same data.

In an earlier study[9] relating to the compression of satellite data it was found that pulse code modulation methods of compression tended to give poorer results than methods based on the ideas of prediction and interpolation. The compression methods discussed in this present paper have been compared[10] somewhat superficially with the techniques based on the ideas of prediction and interpolation. The results of the comparison showed that there is little difference between run-length encoding with an aperture and compression techniques based on prediction and interpolation; and that Shannon–Fano encoding of amplitude differences gives slightly better results than compression using differential pulse code modulation (d.p.c.m.) with fixed-length code words. This latter result is to be expected since the Shannon–Fano technique applied to differences is in fact a form of d.p.c.m., but with differential elements being encoded in the optimum manner.

## 6 Acknowledgments

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Table 10 Results obtained from application of Hadamard 2

| Compression mode No. / Channel No. | 1 | 2 | 3 | 4 |
|---|---|---|---|---|
| 2 | c.r. = 2·25<br>r.m.s. = 23·5 | c.r. = 2·85<br>r.m.s. = 23·6 | c.r. = 3·05<br>r.m.s. = 23·7 | c.r. = 6·10<br>r.m.s. = 23·7 |
| 3 | c.r. = 2·25<br>r.m.s. = 9·68 | c.r. = 2·85<br>r.m.s. = 9·8 | c.r. = 3·05<br>r.m.s. = 9·86 | c.r. = 6·10<br>r.m.s. = 9·9 |
| 4 | c.r. = 2·25<br>r.m.s. = 21·4 | c.r. = 2·85<br>r.m.s. = 21·6 | c.r. = 3·05<br>r.m.s. = 21·7 | c.r. = 6·10<br>r.m.s. = 21·7 |
| 5 | c.r. = 2·25<br>r.m.s. = 6·50 | c.r. = 2·85<br>r.m.s. = 6·67 | c.r. = 3·05<br>r.m.s. = 6·71 | c.r. = 6·10<br>r.m.s. = 6·77 |

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## 8 Appendix 1: Rules for Constructing a Shannon–Fano Code

The Shannon–Fano encoding procedure involves the following steps:

Step 1. Arrange the source symbols in order of decreasing probability.

Step 2. Divide the source symbols into two groups of as nearly equal probability as possible, and assign a binary zero to each symbol in the upper group, and a binary one to each symbol in the lower group.

Step 3. Repeat step 2 by dividing each of the two groups into two sub-groups of as nearly equal probability as possible, and assign a binary zero to each symbol in an upper subgroup, and a binary one to each symbol in a lower subgroup.

Step 4. Repeat the previous steps until each subgroup contains only one element.

## 9 Appendix 2: Some General Theory Relating to Run-Length Encoding

In this Appendix we shall consider some general theory relating to run-length encoding. In particular we shall derive expressions for data compression ratios.

Let us begin by considering a data source whose statistics are such that whenever it generates a very long sequence of output symbols the sequence can be broken down into runs of lengths $l_1, l_2, \ldots, l_n$. Let the probabilities associated with the runs be $P_1, \ldots, P_n$ respectively.

Suppose the source generates a long sequence of output symbols and that the complete sequence is found to contain a total of $N$ runs of the various lengths $l_1, l_2, \ldots, l_n$. Provided the source generates a sufficiently long sequence, that is, provided $N$ is very large then the sequence will contain $NP_i$ runs of length $l_i$; $i = 1, 2, \ldots, n$. It thus follows that the number of symbols in the sequence generated by the source is $S$ where,

$$\begin{aligned} S &= \sum_{i=1}^{i=n} NP_i\, l_i \\ &= N \,.\, \sum_{i=1}^{i=n} P_i l_i = N \,.\, l_{av}, \end{aligned} \tag{4}$$

where $l_{av}$ is the average run length.

Suppose the symbols generated by the source are $K$-bit p.c.m. words (in ESRO data channels we are dealing with $K = 8$). Under these conditions it follows that the number, $Q$, of binary digits in the data stream is,

$$Q = KN \,.\, \sum_{i=1}^{i=n} P_i l_i. \tag{5}$$

In the absence of any data compression, this is the number of binary digits that we should have to use in order to either transmit or store the data stream.

A number of possible encoding schemes which might lead to a compression of the data will now be considered.

*Encoding scheme A.* Assume that the data are encoded and stored, using the following scheme:

'Whenever we reach the start of a run a 9-binary digit word (a zero plus an 8-bit p.c.m. word) is stored and a binary 1 is stored as marker digit for each of the other samples in the run.'

We see that, effectively, this scheme is providing a 1-bit marker for each $K$-bit symbol produced by the source and it is also providing an 8-bit word representing the first sample in a new run. Clearly, the system can be decoded without ambiguity.

The number of binary digits used in encoding the sequence of source symbols is thus:

$$\begin{aligned} R &= S + KN \\ &= N \,.\, \sum_{i=1}^{i=n} P_i l_i + KN. \end{aligned} \tag{6}$$

The data compression ratio is thus:

$$\text{compression ratio} = \frac{KN \,.\, \sum_{i=1}^{i=n} P_i l_i}{N \,.\, \sum_{i=1}^{i=n} P_i l_i + KN}. \tag{7}$$

It should be noted from equation (4) that if the data are slowly varying so that very long sequences arise then the term

$$\sum_{i=1}^{i=n} P_i l_i$$

becomes large and the compression ratio tends to $K$. This is just as we would expect since we are using one 'marker' bit for each $K$-bit p.c.m. symbol generated by the source.

*Encoding scheme B.* Suppose that data are encoded and stored using the following scheme:

'When the start of a new run is reached the $K$-bit word representing the sample is stored and this is then followed immediately by a *fixed-length* code word which gives the length of the run.'

Suppose we find that the longest run is of length $l_{i\,max}$. If we assume that it is possible for all runs of length $l_1, l_2, \ldots, l_{i\,max}$ to occur then we can indicate the length of the run without ambiguity if we use a fixed-length binary word of length $L$, where $L$ is the smallest integer satisfying the condition

$$L \geqslant \log_2 l_{i\,max}. \tag{8}$$

For example, if $l_{i\,max} = 32$ then $L = 5$ is sufficient to enable us to specify any of the 32 run lengths. If, however, $l_{i\,max} = 33$ then we require $L = 6$.

In applying encoding scheme B we use $KN$ binary digits to specify the first samples in the $N$ runs and we also use $NL$ digits in order to provide the necessary information about the run lengths. Thus the compression ratio is:

$$\text{compression ratio} = \frac{KN \,.\, \sum_{i=1}^{i=i\,max} P_i l_i}{KN + LN}. \tag{9}$$

On substituting from condition (5) we have:

$$\text{compression ratio} \leqslant \frac{KN \,.\, \sum_{i=1}^{i=i\,max} P_i l_i}{KN + N \log_2 l_{i\,max}} \tag{10}$$

and it can be seen that if the data are slowly varying so that the average code-word length,

$$\sum_{i=1}^{i=i\,max} P_i l_i,$$

is large then the compression ratio becomes:

$$\text{compression ratio} \leqslant \frac{K \,.\, \sum_{i=1}^{i=i\,max} P_i l_i}{\log_2 l_{i\,max}} \tag{11}$$

It is difficult to be very specific about how condition (11) behaves. It depends very much on the way in which the maximum run-length $l_{i\,max}$ is related to the average run length

$$\sum_{i=1}^{i=i\,max} P_i l_i.$$

In those cases in which long runs tend to predominate, one would expect

$$\sum_{i=1}^{i=i\,max} P_i l_i$$

to increase much more rapidly than $\log_2 l_{i\,max}$ and the compression ratio, therefore, to become very good.

*Encoding scheme C.* (optimum run-length encoding.) If the successive runs produced by the source are statistically independent then it is possible for us to derive an expression for the maximum compression that we can achieve through the use of run-length encoding.

Suppose, as before, the runs are of lengths $l_1, \ldots, l_{i\,max}$ and that they have probabilities of $P_1, \ldots, P_{i\,max}$ respectively.

The least average number of binary digits per run that we have to use (or could use) to encode the run-lengths is given by the entropy function.

$$H = -\sum_{i=1}^{i=i\,max} P_i \log_2 P_i.$$

It thus follows that in order to either transmit or store the data sequence produced by the source, we have to use a total of at least

$$\begin{aligned} NK + N \sum_{i=1}^{i=i\,max} P_i \,.\, \log_2 1/P_i \\ = NK - N \,.\, \sum_{i=1}^{i=i\,max} P_i \,.\, \log_2 P_i \end{aligned} \tag{12}$$

binary digits, and that the compression ratio is given by

$$\text{compression ratio} = \frac{KN \,.\, \sum_{i=1}^{i=i\,max} P_i l_i}{NK - N \,.\, \sum_{i=1}^{i=i\,max} P_i \log_2 P_i}. \tag{13}$$

*Encoding scheme D.* (Shannon–Fano encoding of run lengths.) Consider the situation in which Shannon–Fano encoding is used to encode the lengths of the runs.

As before, suppose that the data source generates runs of lengths $l_1, \ldots, l_{i\,max}$ with probabilities $P_1, \ldots, P_{i\,max}$ respectively.

Suppose also that we construct the Shannon–Fano code for the run lengths and that we find that the runs of length $l_1, \ldots, l_{i\,max}$ are encoded by the Shannon–Fano technique into code words of lengths $L_1, \ldots, L_{i\,max}$ respectively.

Under these conditions, the average code-word length of the Shannon–Fano code is

$$\sum_{i=1}^{i=i\,max} P_i L_i.$$

This is, in fact, the average number of digits that we use to provide information about the lengths of a run. It follows that the number of binary digits used to encode the data is

$$NK + N \,.\, \sum_{i=1}^{i=i\,max} P_i L_i. \tag{14}$$

The compression ratio for the encoding scheme is

$$\text{compression ratio} = \frac{NK \,.\, \sum_{i=1}^{i=i\,max} P_i l_i}{NK + N \,.\, \sum_{i=1}^{i=i\,max} P_i L_i}. \tag{15}$$

From what has been said (see Section 2) on the direct use of Shannon–Fano encoding, we know that the average code-word length of the Shannon–Fano code will probably be nearly equal to the zero-order entropy of the run lengths. This means that, probably,

$$\sum_{i=1}^{i=i\,max} P_i L \simeq \sum_{i=1}^{i=i\,max} P_i \log_2 1/P_i,$$

and hence equations (13) and (15) will have very similar values.

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# Non-linear resistors

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*Based on a paper presented at an IERE Colloquium on Fixed and Variable Resistors held in London on 30th January 1973.*

SUMMARY

Non-linear resistors have been known since the early 19th century but have been of commercial significance only over the last 50 years. Today over 700 million pieces are manufactured annually. This paper outlines the three major types commonly found in the electronics industry, namely, negative and positive temperature coefficient thermistors, and voltage-dependent resistors. The electrical properties of each type are examined and from these various fields of application are derived.

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## 1 Introduction

In 1833 Michael Faraday discovered that the resistance of silver sulphide varied with temperature. This was the beginning of our knowledge of non-linear resistors but little further advance took place for nearly a century when the rapid expansion of the electronics industry and developments in semiconductor materials produced the first practical non-linear resistors. Today, 140 years after Faraday, it is estimated that 700 million non-linear resistors are produced annually.

The variations in type have become so numerous that a simple definition of a non-linear resistor becomes difficult. It is easier to consider first a linear resistor, which maintains its resistance value irrespective of changes of temperature, light, potential, etc. When large changes of resistance accompany changes in these parameters, the component is described as a non-linear resistor. Further subdivision is possible but the two main groups are thermally-sensitive resistors from which the name 'thermistor' is derived,[1] and voltage-dependent resistors usually referred to simply as v.d.r. or occasionally varistors.[2]

## 2 Manufacture

The manufacture of both these classes follows techniques well established in the ceramics industry. The raw materials, frequently metallic oxides, are ball-milled together to give the correct particle size and an intimate mix. The resultant slurry is vacuum filtered and dried. A preliminary firing to induce the correct chemical or physical form is usually required, followed by further ball-milling and mixing with an organic binder. The shape of the component is produced either by pressing in the case of disks, or by extrusion for rods. The final sintering is around 1000°C under carefully controlled conditions of temperature and atmosphere. Electrical connexions are made by means of metallic paints, metal spray or evaporation and connecting leads are added. Finally, the component is protected by lacquer coating or plastic moulding. The electrical properties of a non-linear resistor are primarily dependent on the initial raw materials, the physical size of the component and the manufacturing processes. Each of these variables must be fully understood and subjected to strict quality control to ensure a high yield of acceptable components. Similarly, new components may be developed by variations to these processes but each variation must be subjected to complete investigation to determine their reproducibility.

## 3 Thermistors

Thermally-sensitive resistors or thermistors, can be divided into two groups:[1] those with a negative temperature coefficient of resistance (n.t.c. thermistors) and those with a positive temperature coefficient (p.t.c. thermistors). The resistance/temperature characteristics of the two types are shown in Fig. 1. The resistance of the n.t.c. falls following an exponential characteristic over a wide temperature range. The p.t.c. thermistor shows a large increase of resistance over a small temperature range.

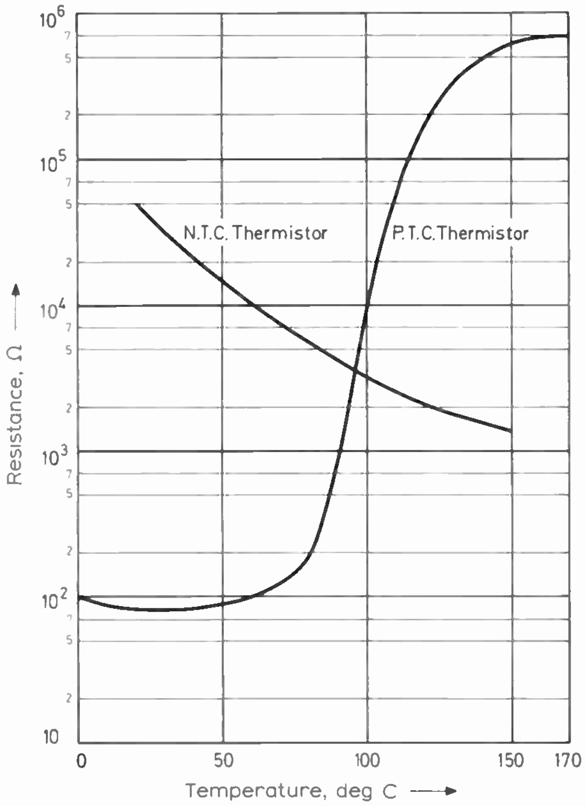


**Fig. 1.** Resistance-temperature characteristics of negative and positive temperature coefficient thermistors.

### 3.1 *N.T.C. Thermistors*

These can be made with room temperature resistance values ranging from a few ohms up to megohms. They are manufactured from oxides of the iron group of elements such as Cr, Mn, Fe, Co, Ni. Normally these oxides have a high resistance but are transformed into semiconducting materials by the addition of small amounts of ions having a different valency. Two examples are (i) iron oxide in which a small part of the $Fe^{3+}$ ions are replaced by $Ti^{4+}$ ions and (ii) nickel oxide or cobalt oxide with a partial substitution of the $Ni^{2+}$ or $Co^{2+}$ ions by $Li^{1+}$ ions.

The relationship between resistance and temperature is given by

$$R_2 = R_1 \exp B\left(\frac{1}{T_2} - \frac{1}{T_1}\right)$$

where

$R_2$ is the resistance at $T_2$ K,

$R_1$ is the resistance at $T_1$ K

and $B$ has a value depending on the material used and has values up to 10 000 K.

Manufacturers' data usually give the resistance of the thermistor at 25°C together with the $B$ factor; therefore, this equation may be used to calculate the resistance at any temperature within its working range. Further, by using the appropriate limit values of resistance and $B$ value, the full spread of resistance in any type at any temperature can be found.

The resistance of a thermistor is dependent on its body temperature and changes to this can be brought about by changes in ambient temperature or by self-heating due to power dissipation within the component. When thermistors, especially the small bead type, are used for tempera ture measurement, the power dissipation must be kept to a low level to avoid inaccuracies due to self-heating.

Figure 2 shows the voltage-current characteristic of an n.t.c. thermistor. Initially the relationship is linear, since, at low power levels, the dissipation is insufficient to raise the temperature above ambient. At higher power levels the resistance falls and a value of voltage $E_{max}$ is reached when further increases of current cause a fall in potential across the thermistor.

Dissipation factor and thermal time-constant are two further properties frequently quoted. The first of these is the power expressed in milliwatts required to raise the temperature of the thermistor by 1 deg C. The time-constant is the time for the resistance of the thermistor to change by 63% of the total change when subjected to a step function change in temperature.

### 3.2 *Applications of N.T.C. Thermistors*

The main application for n.t.c. thermistors is in the field of temperature measurement and control. The properties which make them very suitable for this purpose are:

(i) large temperature coefficient of resistance
(ii) wide operating temperature range
(iii) small size and ruggedness.

The water temperature indicator in road vehicles uses an n.t.c. thermistor and ammeter in series across the battery supply. A disk thermistor is held at the bottom of a closed metal tube by a spring which also acts as the non-earthed contact. The tube is mounted in the engine's cooling system whilst the indicator, the ammeter is on the facia panel. This is the simplest circuit that can be used to measure temperature. For greater sensitivity and accuracy, various bridge networks may be used together with linearization networks of fixed resistors to offset the exponential thermistor characteristic. Medical thermometers with an accuracy of $\pm 0{\cdot}1$ deg C over the range of 25°C to 45°C are available.

In addition to industrial process control, thermistors used with solid-state switching circuits are being used in domestic appliances such as washing machines and central-heating installations. For these applications,

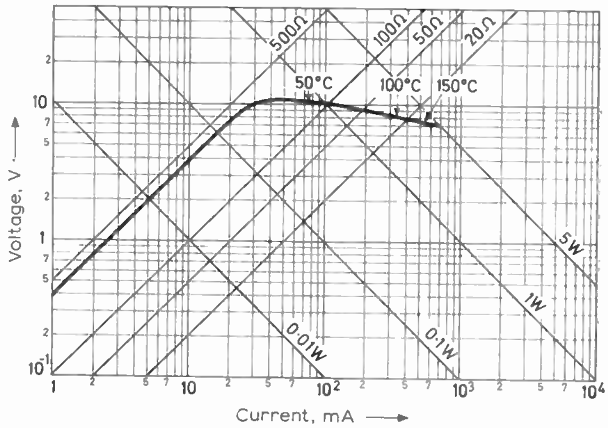


**Fig. 2.** Voltage-current characteristics of an n.t.c. thermistor.

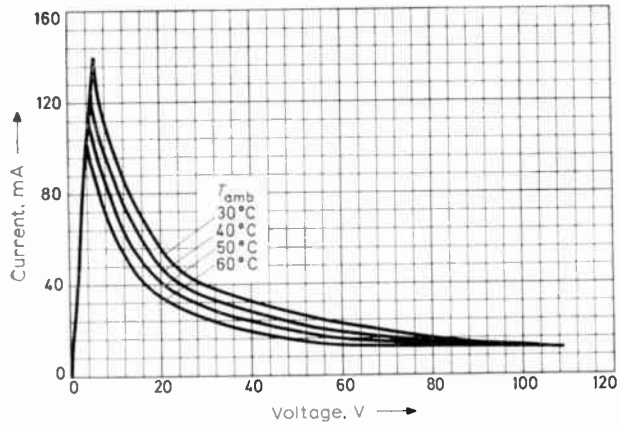


Fig. 3. Voltage-current characteristics of a p.t.c. thermistor.

specially encapsulated thermistors are used to give protection from the environment but to maintain a close thermal contact. To obtain a precise control over temperature, the resistance of these thermistors may be given at several temperatures or a full calibration curve may be supplied. In addition to the above applications in temperature measurement and in temperature control the thermal compensation of transistor circuits to offset performance changes due to temperature fluctuations must also be mentioned. Copper has a positive temperature coefficient of resistance of approximately +0·4% at room temperature. The negative coefficient of an n.t.c. thermistor can be used to offset this change and, for this purpose, thermistors are frequently included on field scanning circuits of television receivers and in series with moving coil meters. One of the first large-scale uses of thermistors was in a.c./d.c. valve operated radio and television receivers; the series-connected heater chain was protected by a thermistor from the initial high current surge at switch-on. For receivers using solid-state devices, a thermistor may still be used to protect the switch or rectifier diode from initial high current surges. The thermal inertia of a thermistor may be used to delay the action of an electromagnetic relay. The delay time ranges upwards from a few seconds and is governed by the selection of coil resistance, thermistor and applied voltage.

Miniature bead thermistors are used to measure infrared radiation and microwave power. They may be used to measure fluid rates of flow and as level indicators in storage vessels. These small thermistors are made by sintering a drop of thermistor paste on to two platinum alloy wires to give a bead approximately 0·5 mm in diameter. These small beads have a very fast thermal time-constant necessary for these applications

### 3.3 *P.T.C. Thermistors*

Thermistors can also be made with a positive temperature coefficient of resistance but, as shown in Fig. 1, their characteristic is not the inverse of the n.t.c. type.

These thermistors are made from barium titanate. When used in its monocrystalline form this material has a resistance which varies inversely with temperature. A p.t.c. thermistor is not however monocrystalline but rather numerous small crystals bonded together during the sintering process. At a certain temperature, barrier layers form at the intercrystalline boundaries and impede the electron flow. As the temperature rises, so does the resistance of these barrier layers until, above a certain limit, the material resumes its normal negative characteristic, but at a much higher resistance value. The nature of this resistance-temperature characteristic prevents a simple mathematical relationship and manufacturers usually quote a resistance at 25°C together with resistance values at other temperatures. The term 'switch temperature, $T_{SW}$' is introduced to denote the temperature at which the resistance starts to rise rapidly. It is defined as that temperature at which the thermistor has a resistance equal to twice its minimum value.

Examination of the voltage-current characteristic (Fig. 3) shows the initial linear portion of the curve where voltage and current rise together followed by the rapid drop in current that occurs once the thermistor has changed to its high resistance state.

### 3.4 *Applications of P.T.C. Thermistors*

Commercial p.t.c. thermistors have been available for a relatively short time, one of the first applications being to protect electric motors against burn-out. Previously, motors had been protected with excess-current cut-outs or bi-metal type thermostats. The small size and low thermal inertia of a p.t.c. resistor embedded directly in the motor windings gives a more satisfactory protection under all fault conditions. Overheating of the motor due to excessive load or blocked ventilation causes the resistance of the p.t.c. thermistor to rise. A simple solid-state switching circuit senses this change and causes either the motor's power supply to be cut off or an alarm given.

Similarly the change in resistance of a p.t.c. thermistor can cause an alarm to be given in many applications where overheating occurs under fault conditions. Warning of overheated axle-boxes in railway rolling stock is but one example.

Demagnetization of the magnetic shield surrounding a shadow-mask colour television tube must be effected each time the receiver is switched-on. For this purpose, degaussing coils fitted to the shield are initially energized and the demagnetization current gradually reduced to a negligible level. This is achieved automatically by a p.t.c. thermistor and a voltage-dependent resistor. More recently, dual p.t.c. elements consisting of two p.t.c. thermistors electrically in series and thermally in contact with each other have been introduced.

## 4 Voltage-Dependent Resistors

The third major group of non-linear resistors are voltage sensitive;[2] their resistance falls as the applied voltage increases. Voltage-dependent resistors (v.d.r.) are made from silicon carbide and their voltage dependence is caused by the contact resistance between the carbide crystals forming a complicated network of series and parallel paths through the material.

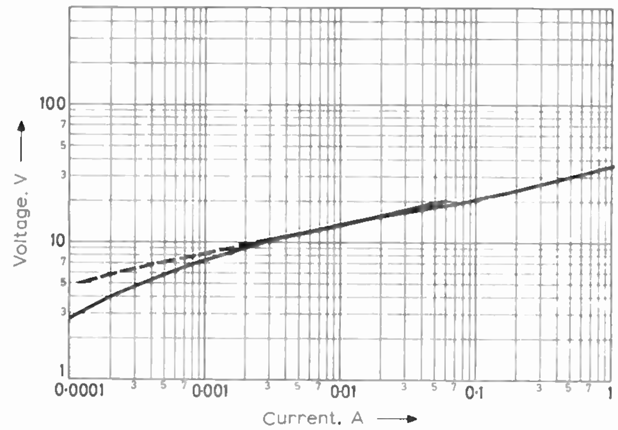


Fig. 4. Voltage-current characteristic for a v.d.r.

The relationship between voltage and current for a v.d.r. can be approximately written as

$$V = CI^B$$

where $V$ and $I$ are in volts and amperes respectively and $C$ and $B$ are constants for a given v.d.r.

This relationship is shown in Fig. 4. The values of $C$ and $B$ depend on the composition of the material and the dimensions of the disk. The $C$ value, defined as the voltage for a current of 1A, ranges from 10 V to 1000 V for disk types and up to 4000 V for rods. Typical $B$ values are from 0·13 to 0·25.

In using a v.d.r., it is important to remember that the power dissipation is proportional to the 6th power of the applied voltage for a $B$ value of 0·2. A voltage increase of only 12% will double the power dissipation. The circuit conditions must be fully investigated to ensure that the component ratings are not exceeded. The change of resistance with change of applied voltage is instantaneous —there is no delay whilst the temperature of the resistor changes as there is with thermistors. If a sinusoidal voltage waveform is applied to a v.d.r. the resultant current waveform will be non-sinusoidal with a high third-harmonic content.

### 4.1 *Applications of Voltage-Dependent Resistors*

Voltage stabilization, switch contact protection and surge suppression are the main applications for v.d.r. The voltage surge caused by breaking an inductive circuit will cause a spark at the switch contacts and in due course the switch contacts may be damaged by burning, resulting in faulty operation. A v.d.r. connected in parallel with the coil, which is usually a relay, absorbs the energy of the surge and protects the contacts. In small battery-operated motors, sparking at the commutator limits the useful life of the motor and may cause interference in any associated electronic circuits. Voltage-dependent resistors connected between the commutator segments will eliminate this problem. In television receivers, v.d.r.s are extensively used for peak voltage limitation to protect valves and semiconductors and the stabilization of circuits against voltage fluctuations.

### 4.2 *New Developments*

Recent years have shown two new developments in voltage-dependent resistors. The first is an asymmetric v.d.r. in which the current-voltage characteristic is dependent upon the polarity of the applied voltage. In many respects they may be likened to a Zener diode with a 'knee' at 1·5 V and extremely low current flow in the reverse direction. Secondly, a new material based on zinc oxide has been developed.[3] This material has a very low $B$ value of about 0·05. This low $B$ value makes these components most suitable for surge suppression on the a.c. mains supply.

## 5 Conclusions

This paper has attempted to outline the main types and properties of non-linear resistors and in so doing has deliberately avoided a detailed or mathematical approach. It can be seen that there are numerous applications in very varied fields for these components. Their unusual and extremely useful characteristics warrant detailed study by all electronics applications engineers. New applications are continually arising and new materials are being developed to broaden the fields of use so that the future of the non-linear resistor is well assured.

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# The choice of data transmission system for efficient use of transmitter power

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SUMMARY

The relationship between system error probability and transmitter power is investigated for block and convolutional codes operating in conjunction with binary phase-reversal keying. The following assumptions are made: (i) all systems accept the same data rate, (ii) a very large bandwidth expansion is not acceptable, (iii) the channel is disturbed by white Gaussian noise, (iv) a matched filter receiver is used.

The performances of these coded systems are compared with those of optimal binary and *M*-ary modulation schemes to give an indication of the systems to be considered in order to achieve efficient use of transmitter power for given system bit error probabilities.

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## List of Symbols

| | |
|---|---|
| $E_B$ | energy per data bit ($J$) |
| $E_n$ | energy per signal element ($J$) |
| $E(R)$ | reliability function |
| $k$ | number of data bits per block or symbol |
| $k_0$ | number of data bits per convolutional code block |
| $m$ | number of convolved blocks |
| $mn_0$ | constraint length of convolutional code |
| $M=2^k$ | number of discrete symbols in a multi-level system |
| $n$ | number of signal elements (= total bits per binary code word) |
| $n_0$ | total bits per convolutional code block |
| $N_0$ | noise power per hertz (W/Hz) |
| $p$ | channel bit error probability |
| $P_B$ | system bit error probability |
| $P_w$ | system word (symbol, block) error probability |
| $q=(1-p)$ | probability of channel bit being correct |
| $R$ | system data rate (bit/s) |
| $R_0$, $R_2$, $R_{2B}$, $R_c$, $R_L$ | coding performance bounds (bit/s) |
| $R_n$ | code ratio, (bits/dimension) = 1/bandwidth expansion ratio |
| $t$ | maximum number of bits which can be corrected by a block code |
| $v$ | number of channel bits in error per word |
| $W$ | bandwidth (Hz) |

$$\text{erfc}(x) = \frac{2}{\sqrt{\pi}} \int_x^\infty e^{-u^2}\, du$$

$$\binom{n}{r} = {}_nC_r = \frac{n(n-1)\ldots(n-r+1)}{r!}$$

## 1 Introduction

The designer of a communication system for the transmission of digital data is faced with a choice of a number of binary and *M*-ary modulation schemes, together with the possibility of using digital error-correcting encoding and decoding. Figure 1 shows in block form the units in such a system.

The relative merits of various modulation schemes operating over a channel perturbed by white Gaussian noise have already been considered in an earlier paper.[1] It was found that *M*-ary orthogonal binary codes in conjunction with correlation detection could theoretically provide error-free communication at the Shannon capacity, provided that the code length and bandwidth occupancy tend to infinity. Very long orthogonal binary codes can be generated and decoded more economically if a hard binary decision is made at the receiver, and such a scheme is likely to be attractive for communication systems having very large available bandwidths.

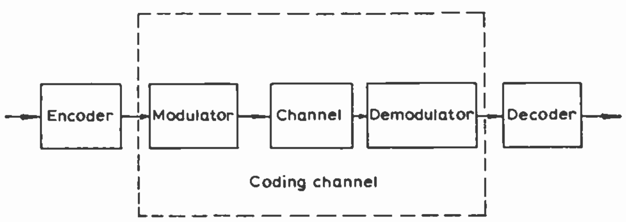



**Fig. 1.** Block diagram of a digital communication system.

This paper is concerned with systems which have a limited bandwidth availability, and consequently only binary codes needing a restricted bandwidth expansion are considered. The theoretical information losses due to binary signalling and hard decisions at the receiver together with the performance bounds of block and convolutional codes are reviewed before specific codes are analysed.

The performance of block and convolutional codes are evaluated when transmitting over a binary phase-reversal keying channel. The ability of binary error-correcting codes to reduce the probability of bit error for the decoded data stream is well known. Many papers, however, compare the error probability of the coded system with that of the uncoded system for a fixed probability of bit error in the communication channel. This implies that the energy per channel bit is the same in each case, and since redundant bits have to be added to the data then either the data rate is less for the coded system or the signal power is increased in proportion to the rate of redundant bits. To achieve a fair comparison between systems the information rate $R$ into the system should be kept constant and error performance plotted as a function of energy per data bit into the system. In this paper care has been taken to express the results in these terms. Shastova[2] has compared the performance of Hamming codes on this basis, but surprisingly little else seems to have been published.

Error probability curves for block and convolutional codes are compared with those of selected modulation schemes. The results indicate which systems should be considered to achieve efficient use of transmitter power as a function of the system bit error probability required.

## 2 The Effects of Signal Quantization

To simplify the implementation of transmitter and receiver it should be possible to quantize the signals, preferably to a binary form. This can be conveniently considered in two steps:

(i) quantizing the transmitted signals, but assuming correlation detection at the receiver,

(ii) replacing the correlation detection by a quantized decision on each received signal element.

Using Shannon's arguments related to an ensemble of codes it may be shown[3] that the probability of error of a code word may be bounded by an equation of the form:

$$P_w < 2^{-n(R_L - R_n)} \tag{1}$$

where $n$ = number of signal elements per code word. Each element may be binary or multi-level in nature, and may be considered as a dimension in a multi-dimensional space, so that the $n$ signal elements in a code word specify a position in an $n$-dimensional hyperspace.

$R_L$ represents the limiting number of information bits which may be transmitted per dimension as $n$ tends to infinity. $R_L$ is a function of the channel signal-to-noise ratio and of the system constraints.

$R_n$ is the code ratio in bits per element. Note that $1/R_n$ is the bandwidth expansion of the coded system relative to uncoded binary transmission. Hence provided that $R_L > R_n$ and $n$ is made large enough it should be possible to attain an infinitesimally small probability of error. For finite $n$, $(R_L - R_n)$ gives a measure of how rapidly the error probability reduces with increasing $n$.

Figure 2 shows how $R_L$ varies with signal-to-noise ratio, $E_n/N_0$, for different systems. $E_n$ is the mean energy per dimension ($J$), not the energy per data bit, and $N_0$ the noise power (W) per hertz in the channel. If a constraint is placed only on the energy per word the transmitted signal vectors for each word are contained within an $n$-dimensional hypersphere, and theoretically it should be possible to achieve the Shannon capacity. This performance is represented by the curve $R_L = R_c$.

If the power in each dimension (element) of the signal is limited to a maximum value, the signal vectors of each word are contained within an $n$-dimensional hypercube, and $R_L = R_0$. As would be expected, some loss of performance occurs; this tends to 3 dB at low signal powers.

When the transmitter is constrained to sending binary symbols, $n$ becomes the number of bits transmitted per channel word. Obviously in this case not more than 1 bit can be sent per dimension, and this places an upper limit

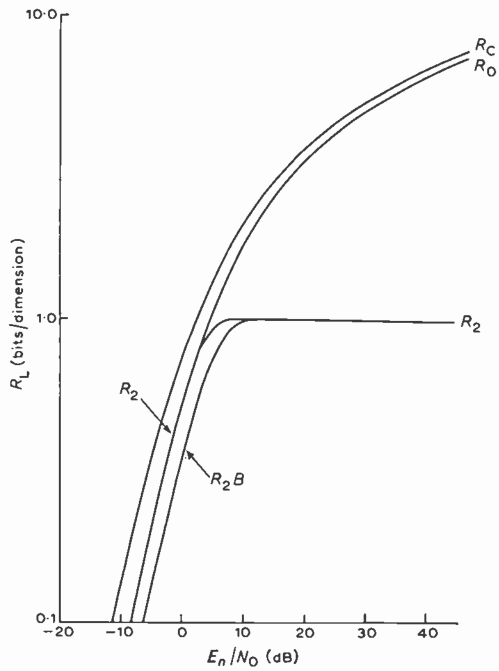



**Fig. 2.** $R_L$ versus $E_n/N_0$ for various quantization constraints.

on $R_2$. However, if the channel signal/noise ratio is less than +2 dB, no loss of performance is incurred due to the binary restriction, as $R_2$ and $R_0$ are equal. It should be noted that the argument of $R_L$ in these curves is energy per element, not energy per data bit.

Once true correlation at the receiver is replaced by a binary decision on each element of the waveform a loss of information occurs at the receiver. The system performance is given by $R_L = R_{2B}$. At low signal/noise ratios a system degradation of only 2 dB is experienced, and at high signal/noise ratios performance of the hard decision system is as good as that of the cross-correlated binary signal.

### 2.1 *Bounds on Transmitter Power*

In order to transmit information with a low word error probability using hard quantization, it is necessary for $R_n$ to be less than $R_{2B}$ (equation (1)). Since $R_{2B}$ is never greater than 1, it is therefore advisable to insert redundancy in the binary word. Is there, then, some optimum amount of redundancy, i.e. an optimum code ratio? If $n$ is allowed to be very large, then $P_w$ will tend to zero provided that $R_{2B} > R_n$. Using the equations[3]

$$p = \tfrac{1}{2}\,\mathrm{erfc}\sqrt{\frac{E_n}{N_0}} \qquad (2)$$

$$R_{2B} = 1 - \log_2\left(1 + 2\sqrt{p(1-p)}\right) \qquad (3)$$

$$R_n = R_{2B} \qquad (4)$$

$$\frac{E_B}{N_0} = \frac{1}{R_n}\cdot\frac{E_n}{N_0} \qquad (5)$$

the lower bound for $E_B/N_0$ for zero word error probability as $n \to \infty$ can be computed against the code ratio, $R_n$, $E_B$ is the energy transmitted per data bit into the system, so $E_B \times R$ is the total transmitter power required to transmit at a data rate of $R$ bit/s. The bound is shown in Fig. 3(i), and indicates that low ratio codes perform best. This is as we should expect since these codes occupy a large bandwidth. Half-ratio and one-third-ratio codes suffer penalties of 1·1 dB and 0·7 dB respectively compared with very low ratio codes. Curve (ii) shows the bound for binary transmission using correlation detection using[3]

$$R_n = R_2 = 1 - \log_2\left(1 + \mathrm{e}^{-E_n/N_0}\right) \qquad (6)$$

and curve (iii) the Shannon bound for a system having the same bandwidth, $W$, as the binary coded system.

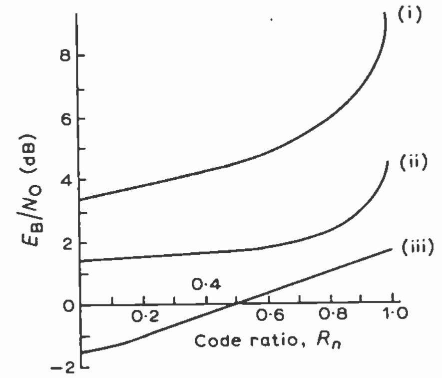



**Fig. 3.** Minimum $E_B/N_0$ required to approach zero word error probability as $n \to \infty$ as a function of code ratio, $R_n$.
(i) binary transmission, hard limiting reception,
(ii) binary transmission, correlation detection,
(iii) Shannon limit ($W = \frac{1}{2}R/R_n$).

## 3 Classes of Code

Block coding and convolutional coding are the two main classes of code used for adding redundancy to digital messages. Some of these codes are applicable to coding with bases other than 2, however in practice binary codes are used because they are convenient to implement.

### 3.1 *Block Coding*

Data bits are grouped at the transmitter into blocks of $k$ bits. These are re-coded into $n$ ($n > k$) bits containing redundancy, which are then transmitted serially through the system. Because of the redundancy in the coding the receiver is able to detect and possibly correct bit errors in the received block. $M$-ary orthogonal coding is one example of block coding which can correct digit errors.

It is possible to implement many codes as cyclic codes. When such codes are employed the $k$ data bits are first transmitted, unaltered. By circulating these bits in a feedback shift-register the $(n-k)$ redundant bits can be conveniently generated[4] and sent after the data bits. A similar feedback shift-register can be used at the receiver to generate an $(n-k)$ bit syndrome which indicates whether the code word is in error, and which can be used to correct the most likely error pattern in the code word.

### 3.2 *Convolutional Coding*

In a convolutional code the data and check bits may again be considered as being in blocks of $k_0$ information and $(n_0 - k_0)$ check bits. However, the check bits are not derived from the current block only, but also from the contents of the previous $(m-1)$ blocks. The product $mn_0$ is the total number of channel bits which depend on each data bit, and is known as the constraint length of the code. Commonly $n_0$, $k_0$ are small numbers, e.g. 2,1: 3,2 producing a $\frac{1}{2}$ or $\frac{2}{3}$ ratio code respectively. Since the check bits are convolved with the data bits a different approach is needed in decoding this class of code. Sequential decoding has been developed by Fano[5] and Jelineck[6] whilst the Viterbi algorithm[7, 8] gives maximum likelihood decoding.

## 4 Performance Bounds

As would be expected the probability of error of a code depends upon its complexity, which is indicated by the block length $n$ in the case of block codes and by the constraint length $mn_0$ in the case of convolutional codes. For each class of code exponential error bounds have been derived which show the limits of capability of the coding systems.

### 4.1 *Block Coding Bounds*

The probability that a block will be erroneously decoded is bounded by[9, 10]

$$A_1\,\mathrm{e}^{-nE_1(R)} < P_w < A_2\,\mathrm{e}^{-nE_2(R)} \qquad (7)$$

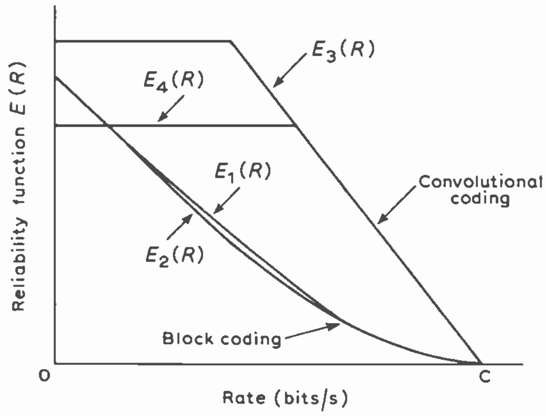



**Fig. 4.** Reliability functions for block and convolutional codes.

where $A_1$, $A_2$ are constants or slowly varying functions of $n$, the block length and $E_1(R)$, $E_2(R)$, are reliability functions of the data rate, $R$.

Curves of $E_1(R)$ and $E_2(R)$ for a typical discrete memoryless channel are shown in Fig. 4. The lower bound in equation (7) represents the performance of the best possible code, whereas the upper bound is achieved by averaging over an ensemble of all possible codes. There should, therefore, be at least one code which performs as well as this.

### 4.2 *Convolutional Coding Bounds*

Similarly the probability that a block of a convolutional code will be erroneously decoded is bounded by[7, 11]

$$A_3\,\mathrm{e}^{-mn_0E_3(R)} < P_w < A_4\,\mathrm{e}^{-mn_0E_4(R)}. \tag{8}$$

Curves of $E_3(R)$, $E_4(R)$ versus $R$ are also shown in Fig. 4, for the same channel. It can be seen that for both classes of code the reliability function falls to zero at a rate equal to channel capacity. Except where the information rate is very low compared with capacity, the convolutional codes have a larger reliability function than the block codes and should therefore provide better performance for equal code lengths. The next section is concerned with the evaluation of the performance of practical codes for comparison with these bounds.

## 5 Performance of Block Codes

### 5.1 *Choice of Modulation System*

In principle, binary error-correction coding can be used in conjunction with either a binary or a multi-level modulation scheme, provided that suitable code conversion is employed at the transmitter and receiver. Many papers have been written showing that for a fixed channel bit error probability the addition of redundant bits to the data bits, i.e. coding, can improve the reliability of received data. Such results are over-optimistic in that they ignore the fact that additional power is necessary to transmit these extra bits if a constant data rate is to be maintained. A more realistic assumption is that the data rate of the system and the power available are constant. The addition of redundant bits will reduce the energy available per channel bit and hence increase the channel bit error probability.

In order to calculate the change in bit error probability it is necessary to have a knowledge of the relationship between bit error probability and signal-to-noise ratio in the channel. This relationship was considered in an earlier paper,[1] where it was found that phase-reversal keying (p.r.k.) was the most efficient of the binary modulation methods. Such a system has a bit error probability of

$$p = \tfrac{1}{2}\,\mathrm{erfc}\sqrt{\frac{E_N}{N_0}} \tag{9}$$

where $E_n$ is the energy per channel bit and $N_0$ the noise power spectral density.

Other binary and $M$-ary modulation schemes have identical or very similar relationships between $p$ and $E_n$, the primary difference being that $E_n$ is multiplied by an appropriate factor.[1] Throughout this section the modulation system will be assumed to be binary p.r.k. with matched filter reception. The performance of coding over other modulation systems can be estimated by application of the appropriate multiplying factor to $E_B$.

### 5.2 *Choice of Code Family*

The choice of code should be such that its performance in terms of the number of channel errors it can correct will be as large as possible, the code ratio $k/n$ as large as possible and $n$ should be small for ease of implementation. The Hamming bound places an upper limit on the performance of codes. Any $(n, k)$ block code has $(n-k)$ redundant bits in it. These have $2^{(n-k)}$ possible combinations, and each combination must be able to specify a particular error pattern for the word. If the code can correct errors of up to $t$ bits then there are

$$\binom{n}{1}+\binom{n}{2}+\binom{n}{3}\cdots\binom{n}{t}$$

allowable error combinations plus the possibility of no error, hence

$$2^{(n-k)} \geqslant 1+\sum_{i=1}^{t}\binom{n}{i}. \tag{10}$$

An optimum code would meet the equality condition for equation (10). Only a restricted range of such codes is available,[4] including an assortment of codes with values of $n$ up to 25, the best single-error correcting Hamming codes and double-error correcting BCH codes. Because of this restriction of choice, practical systems commonly employ non-optimum BCH codes, allowing a wide choice of code length, ratio and error correcting capability. Peterson[4] has shown that BCH codes of length up to $n = 1023$ provide error correcting capabilities better than or as good as the Hamming bound (which is for very long codes). In the following sections the performance of Hamming and BCH codes is evaluated with the assumption of a power-limited system.

### 5.3 *Calculation of Channel Bit Error Probability*

To compare the performance of block codes over a hard-limited binary channel with other modulation schemes it is necessary to use $S/RN_0$, i.e. $E_B/N_0$ as the variable, where $E_B$ is the energy per data bit into the system. The total word energy $kE_B$ is spread over $n$ bits, and consequently the channel bit signal to noise ratio is $(k/n)E_B/N_0$. This is the value used to calculate the channel bit error probability, $p$, from equation (9).

## 5.4 *Calculation of Word Error Probability*

The following calculation is applicable to any $(n, k, t)$ block code having $n$ bits per block, of which $k$ are information bits, and capable of correcting errors of $t$ or less channel bits per word. Let $p$ be the probability of any channel bit being in error and let $q=(1-p)$, the probability of the bit being correct. The probability that the data bits will be decoded correctly is the probability that the number of channel errors is less than or equal to $t$. This is

$$q^n+\binom{n}{1}q^{(n-1)}p+\binom{n}{2}q^{(n-2)}p^2+\ldots\binom{n}{t}q^{(n-t)}p^t$$

Hence the probability of a system word error is

$$P_w=1-\sum_{i=0}^{t}\binom{n}{i}q^{(n-i)}p^i. \qquad (11)$$

Alternatively one may consider the probability of error as being the sum of the probabilities of having a $(t+1)$, a $(t+2)$, etc. error, and hence obtain the equation

$$P_w=\binom{n}{t+1}q^{(n-t-1)}p^{(t+1)}+\binom{n}{t+2}q^{(n-t-2)}p^{(t+2)}+\ldots \qquad (12)$$

Provided $p$ is small enough then $p^{(t+1)}$ will be much larger than $p^{(t+2)}$ and a simple expression for $P_w$ may be obtained:

$$P_w\simeq\binom{n}{t+1}q^{(n-t-1)}p^{(t+1)}. \qquad (13)$$

Equation (11) is subject to rounding-off errors in computation as $p$ becomes small, whereas equation (13) is accurate for small $p$. $P_w$ was calculated from both equations, and except for very long codes the two answers were found to agree over a wide range of $P_w$.

## 5.5 *Choice of Code Length*

A typical requirement of a digital transmission system might be to send a 16-bit computer word to a receiver. How many redundant digits should be attached to the information digits to get the most efficient use of transmitter power? Figure 5 shows the error probability for $k=16$ BCH codes with values of $n$ from 16 to 121. It can be seen that a considerable improvement in performance is obtained by using a $\frac{1}{2}$ ratio code, but decreasing the ratio to 16/121 results in a degradation in performance relative to the $\frac{1}{2}$ ratio code except at very low error probabilities. Thus there is a limit to the amount of redundancy that should be used. It is well known that low ratio BCH codes do perform poorly;[4] better performance might be obtained with other codes.

The bound on word error probability, $P_w<2^{-n(R_L-R_n)}$ given in equation (1) indicates that reduced word error probability can be achieved by increasing $n$. If we have a fixed length data word to be transmitted, a number of these could be grouped 2, 3 etc. together to form a larger block. Provided the error probability of this larger block is less than that of shorter blocks, a nett gain in system

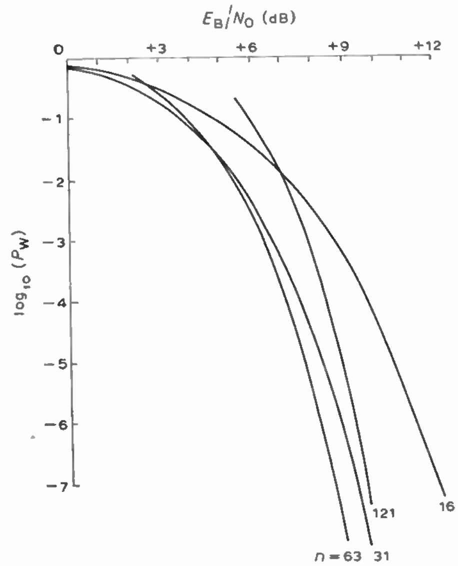



Fig. 5. Block error probability, $P_w$ for BCH codes, $K=16$, $n=16$ to 121.

performance has been achieved. Figures 6(a)–(c) show that for codes of ratio $\frac{1}{3}$ to $\frac{2}{3}$ a substantial reduction in word error probability is achieved with increasing $n$. Alternatively less transmitter power is needed to achieve a given error probability as $n$ increases.

**Table 1**

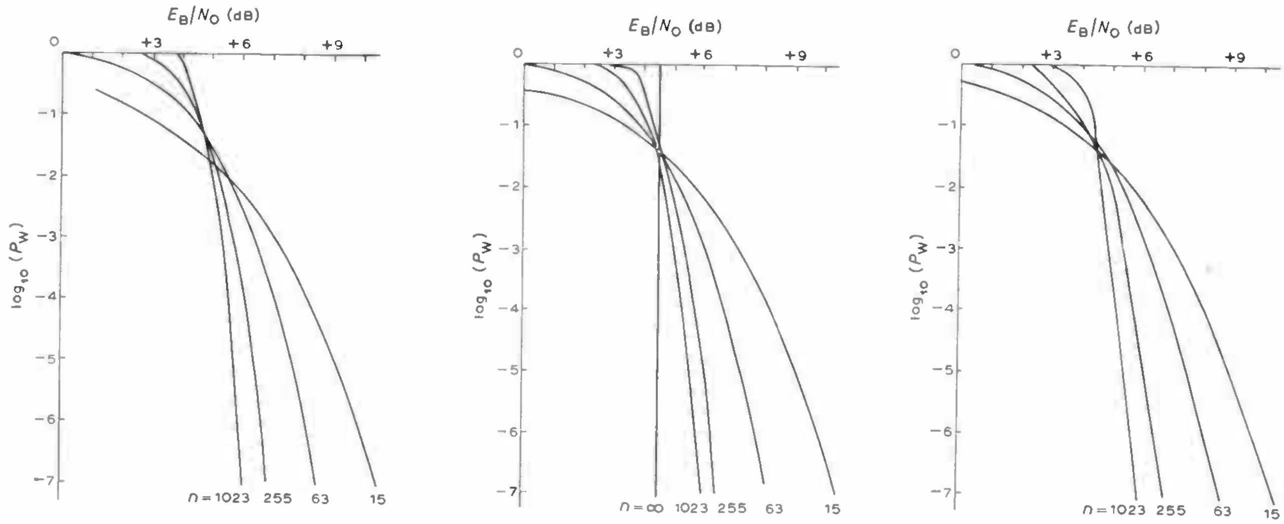
Signal/noise ratio, $E_B/N_0$(dB) required to achieve a word error probability of $10^{-6}$

| $n$ ↑ | $k/n$ → $\frac{1}{3}$ | $\frac{1}{2}$ | $\frac{2}{3}$ |
|---|---|---|---|
| 1023 | 6·0 | 5·8 | 5·8 |
| 255 | 6·7 | 6·3 | 6·6 |
| 63 | 8·2 | 7·8 | 8·2 |
| 15 | 9·9 | 9·8 | 9·7 |

Equation (1) also indicated that $P_w$ is a function of code ratio $R_n$, which is equal to $k/n$ for binary signalling. However $R_{2B}$ is a function of $E_n$, the energy per channel bit and $E_n=(k/n)E_B$ which is therefore also a function of code ratio. Table 1 compares the power requirements of codes of differing length and ratio to achieve an error probability of $10^{-6}$. It can be seen that for a given $n$ the choice of code ratio is not very critical over the range from $\frac{1}{3}$ to $\frac{2}{3}$, but a $\frac{1}{2}$ ratio code is marginally better for values of $n$ of 63 and 255. This is not in agreement with the predictions of the bound on $E_B/N_0$ for binary transmission (Fig. 3(a)) which indicates that low ratio codes are best. Thus there appears to be a further loss in system performance with low code ratios due to the non-optimum structuring of BCH codes.

## 5.6 *Bit Error Probability*

A situation which the communication engineer often faces is that he is presented with a stream of data bits which are delivered in serial form at a fixed rate. These have to be transmitted over a communication channel of limited power capability. The problem is to decide whether to encode the bit stream, and if so how long to




(a) $k/n \simeq \frac{1}{3}$, $n = 15$ to 1023. (b) $k/n \simeq \frac{1}{2}$, $n = 15$ to 1023. (c) $k/n \simeq \frac{2}{3}$, $n = 15$ to 1023.

**Fig. 6.** Block error probability, $P_w$ for BCH codes.

make the blocks and what code rate to choose. Since the segmentation of the data into blocks is purely arbitrary the block error probability is of little significance in this situation; system bit error probability is a more realistic measure of system performance.

To calculate the bit error probability of a decoded block code one has to consider the way in which channel bit errors are corrected. BCH codes with moderate values of $n$ approximate closely to close-packed codes, i.e. codes which conform to the sphere-packing bound. For such codes all valid code-words are uniformly separated in $n$-dimensional space by a Hamming distance[4] of $2t+1$. A one-dimensional representation of the situation is shown in Fig. 7. Only valid code words, of which there are $2^k$, are sent by the transmitter to the receiver. Any channel bit errors will cause the received word to become a non-valid word. The decoding process at the receiver assumes that the number of errors is less than $t$, so that all non-valid words within a distance $t$ of $A$ will be decoded as $A$. If in fact the number of errors is greater than $t$, say $t+1$, then $A$ will be decoded as $B$. In this case the decoding process increases the number of bit errors in the signal. The relationship between channel and decoding errors is as follows:[12]

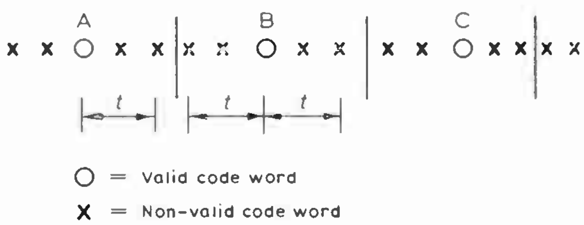



**Fig. 7.** One-dimensional representation of Hamming distance for close-packed $t = 2$ code.

| Input errors, $v$ bits per word | Output errors, bits per word |
|---|---|
| $0 < v < t$ | $0$ |
| $(t+1) < v < (3t+1)$ | $(2t+1)$ |
| $((2t+1)i-t) < v < ((2t+1)i+t)$ | $(2t+1)i$. |

Hence the bit error probability of the $n$ channel bits is

$$P_B = \sum_{i=1}^{\binom{n-t}{2t+1}} \frac{(2t+1)i}{n} \left[ \sum_{v=(2t+1)i-t}^{(2t+1)i+t} P_v \right]$$

where $P_v = \binom{n}{v} p^v q^{(n-v)}$ = probability of a $v$-bit channel error. Now $P_B$ is the bit error probability for the $n$ channel bits after decoding. These are truncated into $k$ data bits which will have the same bit error probability as the $n$ bits.

$P_B$ was calculated for various code ratios and lengths. The results are shown in Figs 8(a)–(c) where $E_B$ is again the energy per data bit (or $S/R$). As would be expected the results show that on the basis of bit error probability the longer codes are better. Table 2 compares the performance of codes of various ratios and lengths on

**Table 2**

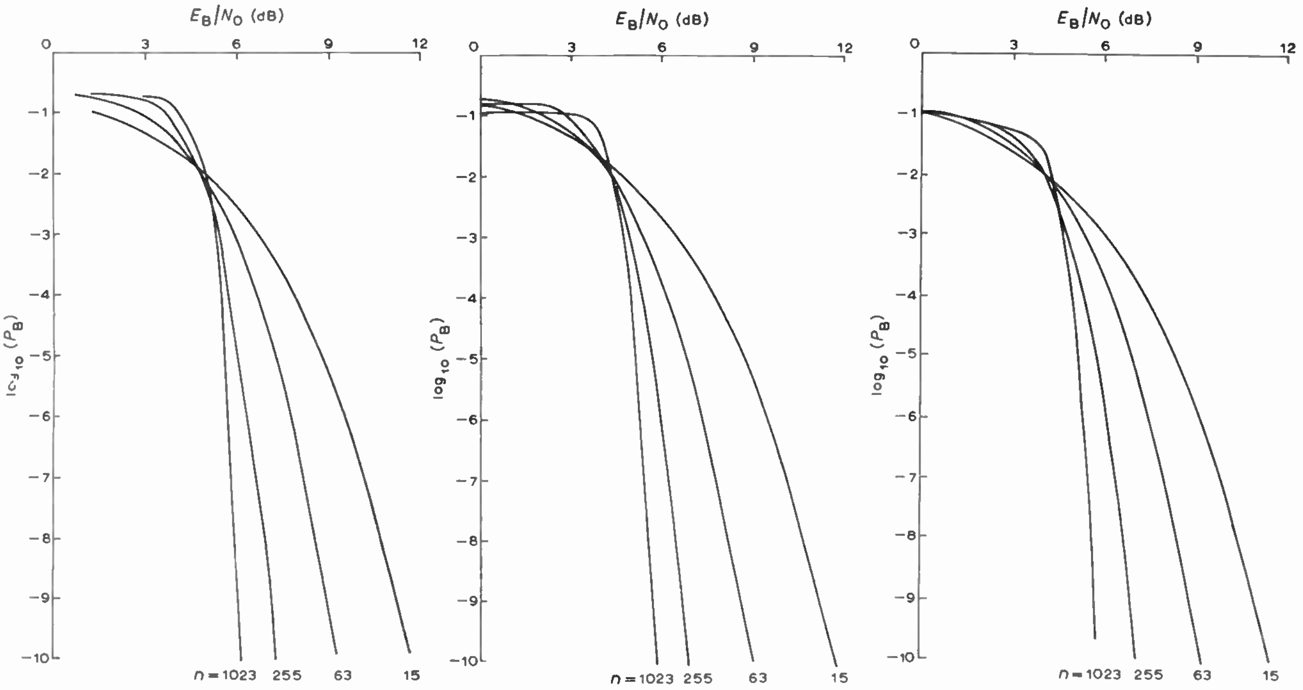
Signal/noise ratio $E_B/N_0$(dB) required to achieve a bit error probability of $10^{-6}$

| $n$ ↑ | $k/n$ → $\frac{1}{3}$ | $\frac{1}{2}$ | $\frac{2}{3}$ |
|---|---|---|---|
| 1023 | 5·9 | 5·4 | 5·4 |
| 255 | 6·6 | 6·1 | 6·2 |
| 63 | 8·0 | 7·4 | 7·6 |
| 15 | 9·6 | 9·5 | 9·3 |

the basis of a bit error probability of $10^{-6}$. These results correlate very closely with those based on word error probability (Table 1), i.e. the choice of code ratio is not very critical, but $\frac{1}{2}$ ratio codes need slightly less power than $\frac{1}{3}$ or $\frac{2}{3}$ ratio codes for $n = 63$ and $n = 255$. The $\frac{1}{3}$ ratio code is never a good choice, but the $\frac{2}{3}$ ratio code is better when short codes are used.

## 6 Performance of Convolutional Codes

The Viterbi[7] decoding algorithm gives optimum maximum likelihood decoding, and would therefore seem the natural choice for convolutional code receivers.



(a) $k/n \simeq \frac{1}{3}$. (b) $k/n \simeq \frac{1}{2}$. (c) $k/n \simeq \frac{2}{3}$.

**Fig. 8.** Bit error probability for BCH codes.

Unfortunately the complexity of the receiver grows rapidly with the constraint length of the code employed, and this imposes a limit upon the size of code which can be used. The complexity of sequential decoders is relatively independent of code length, so that much longer code lengths can be used. Due to the nature of the decoding algorithms it is not easy to calculate the bit error probability of convolutional systems. Heller[13] has used a digital computer simulation of a system employing the Viterbi algorithm to measure the system bit error rate for various values of $E_B/N_0$. The system is assumed to operate over a channel using phase-reversal keying with coherent detection. Figure 9(a) shows the results of half-ratio codes with constraint lengths of 8, 12 and 16 bits and with a hard limiting receiver. Comparing the results for the $mn_0 = 16$ code with that of an $n = 15$ BCH half-ratio code shows an improvement in performance of 2·4 dB at a bit error rate of $10^{-4}$ for the convolutional code. To achieve this performance the Viterbi decoder stores path lengths of 32 bits. It might then be fairer to compare the $mn_0 = 16$ code with the 31 bit BCH code. In this case the convolutional decoder still has an advantage of 1·6 dB.

## 6.1 *Comparison of Code Ratios*

With block codes it was found that the performance of codes of ratio $\frac{1}{3}$, $\frac{1}{2}$ and $\frac{2}{3}$ were very similar for a given value of $n$. The same is true for convolutional codes of similar constraint length. Ratio $\frac{1}{3}$ and $\frac{2}{3}$ codes with $mn_0 = 9$ and ratio $\frac{1}{2}$ codes with $mn_0 = 10$ all require about 6·4 dB to produce a system bit error rate of $10^{-4}$.

## 6.2 *Viterbi Decoding with a Soft-limiting Detector*

By quantizing each bit of received channel signal into eight levels rather than two the information available to the decoder is increased considerably. Heller shows that practically all of the 2 dB loss due to receiver quantization can be recovered in this way. Of course the receiver complexity is increased as an a.g.c. will probably be needed and three bits will have to be manipulated for every channel bit.

## 6.3 *Sequential Decoding*

Much longer code constraint lengths can be used in conjunction with sequential decoding, so that a more rapid rate of change of error probability is achieved with $E_B/N_0$. Sequential decoding is therefore more suitable than Viterbi decoding when low error rates ($< 10^{-5}$) are required. Forney and Bower[14] describe the performance of a complex sequential decoder capable of

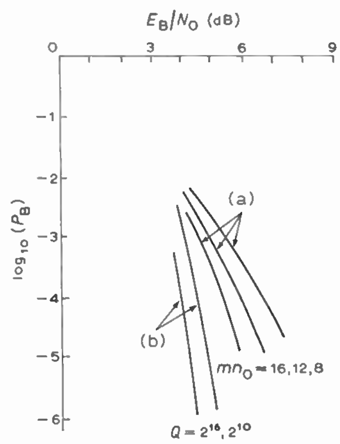



**Fig. 9.** Bit error rate for half ratio convolutional codes with hard-quantized received data.
(a) Viterbi decoding,
(b) sequential decoding, $mn_0 = 94$.

operating at MHz rates. The code used has a ratio of ½ and a constraint length of 94. The performance of the code depends upon the amount of back-search permitted, the number of branches stored in memory and the speed of computation. Because the computer is speed limited a degradation of performance is experienced with increasing data rate. Figure 9 shows bit error rate against $E_B/N_0$ at $R = 50$ kbits/s using a 13·3 MHz computation rate, a back-search limit of 240 and storage capacities, $Q$, of $2^{10}$ and $2^{16}$ branches. To obtain the best performance from the decoder about $2^{14}$ to $2^{16}$ branches need to be stored. At 50 kbit/s this implies a data delay of ¼ to 1 second.

Comparing the performance of the sequential and Viterbi decoded systems shows that the longer constraint length of the sequential decoder results in a more rapid reduction of error probability with $E_B/N_0$.

Comparing the sequentially decoded convolutional code of constraint length 94 bits with a BCH code with $n = 127$ shows a factor of 2·1 dB in favour of the convolutional code at a bit error rate of $10^{-6}$, as indicated by the larger reliability function of the convolutional code.

## 7 Other Systems

So far coding has been considered only in conjunction with binary p.r.k. signalling. By using an $M$-ary orthogonal modulation system with $M > 4$, improved efficiency can be obtained for a given channel bit error probability. If, for example an $M = 32$ system is employed, a gain of 2 to 3 dB in transmitter power can be affected.

Even better performance can be obtained by retaining more information from the receiver for use by the decoder, as in Jordan's system[15] for the Lincoln Experimental Terminal. The output of the receiver in this case consists of an ordered list of the $L$ most likely signals received. This is passed to a sequential decoder, allowing very efficient decoding.

Another possibility is to use concatenated codes; for instance sequential decoding of convolutional codes tends to produce errors in bursts. Such error bursts could be corrected using a suitable burst-correcting code.

This paper has been constrained to consideration of one-way communication channels. When a return channel is also available the possibility arises of using this to send information about the received signal back to the transmitter to modify its operation. This should offer some advantage over one-way systems, and will be the subject of further investigation by the author.

## 8 Summary

### 8.1 *Signal Quantization*

Using an average ensemble of codes, the constraint of binary signalling imposes a theoretical loss of about 3 dB for low ratio codes, compared with Shannon capacity. Hard quantization at the receiver imposes a further penalty of about 2 dB. As the code ratio increases a further small penalty is incurred, which is about 1 dB for a half-ratio code. The $E_B/N_0$ required to signal without error for such a code is 4½ dB.

### 8.2 *Block Codes*

The BCH family of error-correcting codes have a wide range of code length and code ratio and are capable of being implemented as cyclic codes. Analysis of the error probability of systems using these codes over an optimum phase-reverse-keyed binary channel agree well with the predictions of coding bounds and show that for long codes error probability decreases very rapidly with increasing $E_B/N_0$ above a threshold value of about +4½ dB.

### 8.3 *Convolutional Codes*

Convolutional codes have a more complex structure than block codes, and are consequently more difficult to decode, and more difficult to analyse. Coding bounds show that convolutional codes should perform better than block codes of similar code length. Comparing Heller's results of a simulation of a system employing convolutional coding and optimal Viterbi decoding with those of a comparable BCH code this is found to be the case. For short constraint length Viterbi decoding it is possible to quantize the received signal into 8 levels. This increases the decoder complexity, but it virtually eliminates the 2 dB loss due to using a hard-limiting detector. Such a system is competitive with practical $M$-ary coded systems for medium values of error probability. Unfortunately Viterbi decoding is not suitable for long codes, as the equipment complexity grows rapidly with code length. Sequential decoding is more suitable in this case, and for a given code constraint length can provide a better performance than BCH codes when low error probabilities are required.

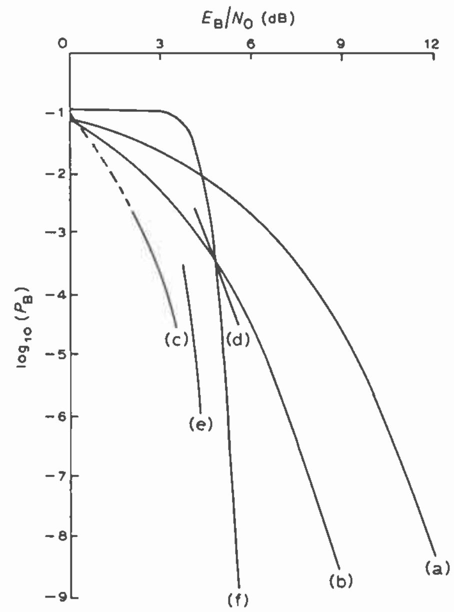



**Fig. 10.** Bit error probability v. $E_B/N_0$ for various modulation and coding schemes. (All coding schemes transmitting over a binary p.r.k. channel with coherent reception.)

(a) coherent p.r.k.,
(b) $M = 32$ orthogonal modulation,
(c) convolutional code, soft Viterbi decoding,
(d) convolutional code, hard Viterbi decoding,
(e) convolutional code, hard sequential decoding,
(f) BCH ½ ratio, $n = 1023$ code.

## 9 Conclusions

For transmitting over a system where the signal is corrupted by the addition of white Gaussian noise, the choice of modulation and coding to provide efficient use of transmitter power will depend upon the system bit error probability permitted. Figure 10 shows the performance of various modulation[1] and coding systems which have been chosen as being roughly representative of the limit of equipment complexity acceptable. If high system error probabilities ($P_B > 10^{-1}$) are permissible then no advantage is experienced in using either $M$-ary orthogonal modulation or any of the coding schemes discussed in the report. For intermediate values of error probability ($10^{-1} > P_B > 10^{-5}$) a short convolutional code in conjunction with soft limiting and Viterbi decoding offers the lowest error probability for a given amount of signal power. Orthogonal signalling is a possible second-best.

To achieve very low system error probabilities ($< 10^{-5}$) with efficient use of power it is necessary to use long codes. To make such systems practicable hard-limiting quantization is employed at the receiver. Despite the consequent loss of information such systems can still offer a better performance than soft-limiting systems of comparable complexity due to the large code length used. Of the two coding classes available, convolutional codes perform better than block codes of equivalent length.

It is very difficult to compare quantitatively the complexities of the different schemes considered other than on the ultimate basis of cost. Such an assessment could be done only for a specific requirement at a given instant in time. The boundaries given should be taken as a guide for design only. Nevertheless the results do give a feel for the systems which should be considered when a certain bit error probability is required.

Even greater efficiency in the use of transmitter power could be achieved at the expense of equipment complexity by interleaving the receiver functions further, for instance by using $M$-ary orthogonal modulation with binary coding, by passing 'list of $L$' information from demodulator to decoder, and/or using concatenated codes. Such systems will be very complex and will have limited application.

The codes which have been considered were chosen for their low bandwidth expansion ratio ($1/R_n$). It has been shown that both for BCH and for convolutional codes very little difference in performance is apparent between codes of ratio $\frac{1}{2}$ and $\frac{2}{3}$. For bandwidth constrained systems the $\frac{2}{3}$ ratio code would therefore seem a better choice.

One system configuration which has not been considered is that employing long $M$-ary orthogonal binary codes with a quantized receiver decision. Such a system could be of advantage when it is required to spread the signal energy over a very wide bandwidth, but is outside the scope of this report.

It should be stressed that the results are strictly applicable only to a white Gaussian noise channel. Satellite communication channels are the only ones which might legitimately be represented by this model. If communication is via a h.f. radio link then the effects of multi-path and fading will also influence the choice of system.[16] Data communication over telephone circuits will suffer disturbances from burst noise, which again requires special consideration.

Finally, it is worth pointing out that the present study has been confined to a one-way communication system. Where a return channel is available to the system then the use of feedback signals over this link may be used to indicate the status of received signals and to modify the transmitter's activity accordingly.

## 10 Acknowledgments

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# Noise and dynamic range in lossy delay lines with compensation amplification

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**SUMMARY**

It is shown that if the loss in a delay line is compensated by amplifying, the dynamic range of the system is maximized by amplifying continuously along the delay line so that the signal strength remains constant. This principle is particularly applicable to microsound delay lines and delay equalizers. Moreover, if the noise figure of the amplifier is sufficiently small, the dynamic range may be greater than in the line without amplification, although it will never exceed that of a lossless line.

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The problem with which this note is concerned is that of preserving dynamic range in a delay line which is lossy, the loss being compensated by amplification. The particular application in view is microsound delay lines and equalizers, although the theoretical result obtained is not dependent on any particular application. The practical significance of the result for microsound applications will be commented on later.

Suppose that we have a delay line with attenuation $1/G$, and that the loss is to be compensated for by means of an amplifier of gain $G$, the noise figure of the amplifier being $F$. The problem is how to distribute the delay and amplification so as to preserve the maximum dynamic range of the composite device. $F$ is defined as the ratio of signal power to noise power at the input divided by the ratio of the signal power to the noise power at the output, the noise power at the input being $kTB$, where $k$ is Boltzmann's constant, $T$ is standard temperature (290 K), and $B$ is the bandwidth of the device. The noise figure of the attenuator is $G$.

The dynamic range of the attenuator alone is limited on the one hand by the maximum signal strength that it can carry without non-linear behaviour, which in the case of a microsound device means either straining beyond the elastic limit or overloading the input transducer to destruction, and in the case of an electromagnetic device may mean excessive dissipative loss or dielectric breakdown, and on the other hand by the least acceptable signal/noise ratio at the output. We assume that the dynamic range of the amplifier is unlimited, so that the overall performance of the delay line and amplifier combined is not modified by non-linear behaviour of the amplifier. This assumption will not always be valid, but it will be seen to be valid for the result to be obtained in this note.

The amplification will be assumed to be of the travelling-wave type, (which in fact it is in the case of a microsound amplifier) so that the power in a wave increases as it progresses along a finite length of the amplifying structure. Thus the gain $G$ and noise figure $F$ apply to a length $l$ of the amplifier. Now suppose that the amplifier is divided up into $n$ identical segments, each of length $l/n$; the gain of each segment will be $G^{1/n}$. Suppose, too, that the delay line is divided up into $p$ identical segments; the gain of each segment will be $G^{-1/p}$ and its noise figure will be $G^{1/p}$. Let the noise figure of each amplifier segment be $f$. Then the overall noise figure $F$ is related to $f$ by

$$F-1 = f-1+\frac{f-1}{G^{1/n}}+\frac{f-1}{G^{2/n}}+\ldots+\frac{f-1}{G^{(n-1)/n}}$$

The series on the right-hand side is easily summed and we obtain

$$f-1 = (F-1)\left(\frac{G-G^{(n-1)/n}}{G-1}\right) \qquad (1)$$

The question now is how to arrange the $n$ amplifier segments and $p$ attenuator segments in cascade so as to maximize the dynamic range of the whole system. For a given configuration, the dynamic range can be calculated by calculating the signal and noise powers at each

**Table 1**

| Configuration | Dynamic range |
|---|---|
| delay line without amplifier | $K/G$ |
| complete amplifier followed by complete delay line | $\dfrac{K}{FG+G-1}$ |
| complete delay line followed by complete amplifier | $\dfrac{K}{FG-G+1}$ |
| two amplifier segments alternating with two delay line segments, amplifier first | $\dfrac{K}{\sqrt{G}+2(\sqrt{G}-1)\left[1+\dfrac{G(F-1)}{(G-1)}\right]}$ |
| two amplifier segments alternating with two delay line segments, delay line first | $\dfrac{K(\sqrt{G}+1)}{2GF+\sqrt{G}-1}$ |
| two delay line segments, with complete amplifier between them | $\dfrac{K}{GF+\sqrt{G}-1}$ |
| $n$ amplifier segments alternating with $n$ delay line segments, amplifier first | $\dfrac{K}{G^{1/n}+n(G^{1/n}-1)\left[1+\dfrac{G(F-1)}{(G-1)}\right]}$ |

junction, using the gains and noise figures of the various segments, and applying the limiting conditions stated above. The results for a few configurations are given in Table 1 where $K$ is a constant whose value depends on the limiting conditions.

It is apparent that the greatest dynamic range of the configurations listed is that for $n$ alternating amplifier and delay-line segments, and that this is the greater, the greater $n$. In practice there would be a limitation due to noise introduced at the terminals between the segments, so that the dynamic range indicated would not be obtained with a large finite number of segments. But if the limit is taken as $n$ approaches infinity, this difficulty is removed, for now we have a continuous system consisting of an active delay line with just enough gain to compensate for the dissipation.

An example of such a system is a microsound amplifier. This consists essentially of two parallel surfaces with a minute air-gap between. The lower surface carries a surface acoustic wave—probably a Rayleigh wave—and the material is piezoelectric. The upper surface is semiconducting and electrons flow along it under the action of a d.c. electric driving field. Electric fields from the lower surface, caused by the strains as the acoustic wave passes, interact with the electrons in the upper surface, causing bunching. The action is then like that in a microwave travelling-wave tube, and amplification occurs if the driving field is of a suitable magnitude. Because the acoustic waves must traverse a certain distance in order to be amplified a certain amount, and this takes a finite time, the amplifier is essentially also a delay line, and within wide limits the delay can be made dispersive by adding a thin layer of a different material to the lower surface, and the device may then be used as an equalizer. For large delays at low frequencies, and for relatively short delays at higher frequencies (greater than 1 GHz), losses in acoustic-surface-wave delay lines may become excessive, and could be compensated for by amplification. The limitation on length or frequency may then depend not so much on the materials available as on the tolerancing that can be achieved with photolithographic fabrication of transducers.

Such a system is continuous, so that there are no internal terminals, and the above difficulty about noise generated at such terminals no longer arises. Also, because the signal is maintained constant, the dynamic range of the amplifier does not enter into the question, which justifies the assumption made above for purposes of calculation that the dynamic range of the amplifier can be taken as unlimited.

A delay line with compensatory amplification has recently been reported by Coldren,† who does not, however, discuss the question of dynamic range.

The dynamic range of the continuous system with constant signal level is the limit of

$$\frac{K}{G^{1/n}+n(G^{1/n}-1)\left\{1+\dfrac{G(F-1)}{G-1}\right\}}$$

as $n$ approaches infinity. It can be shown that

$$\lim_{n\to\infty}\{n(G^{1/n}-1)\} = \ln G \tag{2}$$

and hence the required dynamic range is

$$\frac{K}{1+\ln G.\left\{1+\dfrac{G(F-1)}{G-1}\right\}} \tag{3}$$

The same result is obtained if the alternating amplifier and attenuator elements are taken as starting with an attenuator segment instead of an amplifier segment.

Expression (3) is greater than any of the expressions in the Table when there is amplification, from which we can conclude that for a delay line with amplification continuously along its length, as is convenient with microsound delay lines, the dynamic range is greater than with any other arrangement. Moreover, comparing expression (3) with the first tabulated value, it is seen that the dynamic range of the active delay line may be greater than that of the same delay without amplification if $F$ is small enough, that is, if

$$1+\ln G.\left\{1+\frac{G(F-1)}{G-1}\right\} < G \tag{4}$$

or

$$F < \frac{(G-1)^2}{G\ln G}+\frac{1}{G} \tag{5}$$

† Coldren, L. A., 'Optimizing loss-compensated long delay devices', *Inst. Elect. Electronics Engrs Trans. on Sonics and Ultrasonics*, SU-20, pp. 17–23, January 1973.

But although in principle the dynamic range of a lossy delay line can be improved by amplification, it can never be made to exceed that of a lossless line, because the left-hand side of the inequality (4) can never (with the restrictions $G \geqslant 1$, $F \geqslant 1$) be made less than 1. Expression (3) shows that dynamic range is greatest when $G$ and $F$ are both as near to 1 as possible.

To summarize, it has been shown that the dynamic range of a delay line with sufficient amplification to compensate for dissipation is maximized when the amplification process is sufficiently small, as given by the inequality (5). This result is particularly appropriate to microsound delay lines and delay equalizers, which can be designed in this manner.


### Acknowledgment

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## STANDARD FREQUENCY TRANSMISSIONS— August 1973

*(Communication from the National Physical Laboratory)*

| August 1973 | Deviation from nominal frequency in parts in $10^{10}$ (24-hour mean centred on 0300 UT) | | | Relative phase readings in microseconds NPL—Station (Readings at 1500 UT) | | August 1973 | Deviation from nominal frequency in parts in $10^{10}$ (24-hour mean centred on 0300 UT) | | | Relative phase readings in microseconds NPL—Station (Readings at 1500 UT) | |
|---|---|---|---|---|---|---|---|---|---|---|---|
| | GBR 16 kHz | MSF 60 kHz | Droitwich 200 kHz | *GBR 16 kHz | †MSF 60 kHz | | GBR 16 kHz | MSF 60 kHz | Droitwich 200 kHz | *GBR 16 kHz | †MSF 60 kHz |
| 1 | +0·1 | 0 | +0·1 | 720 | 628·9 | 17 | +0·1 | +0·1 | 0 | 705 | 617·8 |
| 2 | +0·1 | +0·1 | 0 | 719 | 627·9 | 18 | +0·1 | +0·1 | +0·1 | 704 | 616·6 |
| 3 | +0·2 | +0·1 | 0 | 717 | 626·7 | 19 | +0·1 | +0·1 | 0 | 703 | 615·6 |
| 4 | 0 | +0·1 | 0 | 717 | 626·1 | 20 | +0·1 | +0·1 | 0 | 702 | 614·6 |
| 5 | 0 | 0 | 0 | 717 | 625·8 | 21 | +0·1 | 0 | 0 | 701 | 614·8 |
| 6 | +0·1 | +0·1 | −0·1 | 716 | 625·0 | 22 | 0 | 0 | +0·1 | 701 | 614·5 |
| 7 | +0·1 | 0 | 0 | 715 | 624·8 | 23 | +0·1 | +0·1 | +0·1 | 700 | 614·4 |
| 8 | +0·1 | +0·1 | 0 | 714 | 624·0 | 24 | 0 | 0 | +0·1 | 700 | 614·4 |
| 9 | +0·1 | 0 | 0 | 713 | 623·9 | 25 | +0·1 | +0·1 | 0 | 699 | 613·9 |
| 10 | +0·1 | +0·1 | 0 | 712 | 623·0 | 26 | 0 | 0 | +0·1 | 699 | 612·6 |
| 11 | +0·1 | +0·1 | +0·1 | 711 | 622·0 | 27 | 0 | 0 | 0 | 699 | 612·4 |
| 12 | +0·1 | +0·1 | 0 | 710 | 621·2 | 28 | 0 | 0 | 0 | 699 | 611·5 |
| 13 | +0·1 | +0·1 | +0·1 | 709 | 620·4 | 29 | 0 | +0·1 | 0 | 699 | 611·0 |
| 14 | +0·1 | +0·1 | +0·1 | 708 | 620.2 | 30 | 0 | 0 | 0 | 699 | 610·8 |
| 15 | +0·1 | +0·1 | +0·1 | 707 | 619·5 | 31 | 0 | 0 | 0 | 699 | 611·0 |
| 16 | +0·1 | 0 | +0·1 | 706 | 619·1 | | | | | | |

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

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

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

UDC 621.375.4.029.64: 621.382.2.011.222

# Linear microwave amplification using impatt diodes

G. B. MORGAN, B.Sc., Ph.D.*

SUMMARY

Impatt diode amplifiers are candidates for use in super high frequency (s.h.f.) systems as broadband, intermediate power amplifiers, but their poor linearity makes them unattractive for some systems. The linearity of some X-band impatt diode amplifiers is considered in this paper. The amplifiers are of the circulator coupled, reflexion type, where $p^+nn^+$ silicon diodes of various doping levels are operated in the impatt mode. The impatt class of diodes shows a r.f. negative resistance that depends on the r.f. voltage, so that the design of linear amplifiers using passive circuitry is empirical. The structure of the diode chip determines the spatial distribution of the electric field which, in turn, determines the value of the negative resistance and its dependence on r.f. power. A range of theoretical ½ W mesa structures has been investigated to get a computed, optimum, linear amplifier at a frequency of 12 GHz, where the optimizing parameter is the choice of passive stepped impedance transformer. The computations show that somewhat better linearity can be obtained with lower-doped diodes than with the higher-doped diodes used in the design of impatt oscillators. The characteristics of experimental amplifiers using relatively highly-doped diodes are presented and show good qualitative agreement with the computed responses.

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## 1 Introduction

At super high frequencies the impatt amplifier offers the radar or communication systems engineer a possible solid-state replacement for the intermediate power valve. Present-day microwave amplifiers requiring 10 W, or so, at X band and 1 dB bandwidths of about 5% are either travelling-wave tubes, or cross-field valves, and both these valves show good efficiency and linear characteristics. However 10 GHz impatt diodes have been reported[1] that are thermally limited to power outputs of 3 W c.w. per diode mesa with efficiencies, as oscillators, of greater than 10% for silicon chips and somewhat higher for gallium arsenide chips. Higher powers can be obtained by suitably combining the diode mesas, or chips, and the compatibility of the impatt diode with integrated circuitry thus makes it a fairly attractive proposition.

Impatt diodes exhibit r.f. negative resistance and analyses of the large signal behaviour of the impatt class of diodes show that the values of negative resistance depend on the r.f. voltage. A passive matching circuit can thus only be designed for a particular power level and a perfectly linear amplifier cannot be designed using passive circuits. The diode impedance is also a function of frequency, d.c. bias current, temperature, and the semiconductor material and doping profile. The impedance is further dependent on the immediate environment of the chip, through the parasitic effects of fringing fields, bonding leads and packaging.

In many microwave systems linear amplification is required, and this paper discusses some empirical methods of designing linear amplifiers and their limitations. The theory of reflexion amplifiers is given and then the design of theoretical amplifiers is first considered, before describing some experimental amplifiers.

## 2 Theory of Reflexion Amplifiers

Negative-resistance amplifiers are preferably of the circulator coupled reflexion type[2] and such an amplifier is shown schematically in Fig. 1. Assuming that the circulator is ideal, the network N is generally a passive circuit which provides a real impedance transformation and reactive equalization to the impatt diode. The transfer function and impedance of the network, $Z_L$, as seen from the diode of impedance $Z_D$, is chosen to give the required amplifier gain and response. If $Z_0$ is the impedance of the circulator and $Z_1$ the input impedance of the network N, the power gain of the amplifier is given by

$$G = \left|\frac{Z_0 - Z_1^*}{Z_0 + Z_1}\right|^2 = \left|\frac{Z_L - Z_D^*}{Z_L + Z_D}\right|^2$$

where $Z_D$ has a negative real part.

If the diode is resonated at the centre frequency of the amplifier, then $Z_D = -R_D$, and, choosing $Z_L$ to be purely resistive ($R_L$), the gain is given by

$$G_0 = \left|\frac{R_L + R_D}{R_L - R_D}\right|^2$$

Amplifier stability requires that $R_L > |R_D|$. Since $R_D$ is a function of the r.f. voltage then, if all the other parameters are passive, the gain will depend on the r.f. voltage.

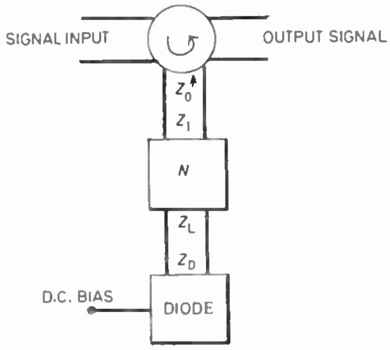


Fig. 1. Schematic diagram of a circulator coupled, negative resistance diode, reflexion amplifier.

To a first approximation (neglecting the change of reactance with r.f. power) the rate of change of gain with power is given by

$$\frac{dG}{dP} = 4G_0^{\frac{1}{2}} \frac{R_L}{(R_L - R_D)^2} \frac{dR_D}{dP}$$

where $P$ is the r.f. power level. $dG/dP$ thus depends on the required amplifier gain, the value of the negative resistance of the diode and the rate of change of this resistance with r.f. power.

## 3 Dependence of Impatt Negative Resistance on the Spatial Variation of the Electric Field

The large signal impedance values of impatt type diodes have been calculated by several workers and the effect of the spatial distribution of ionization on the diode impedance can be seen from following argument. For the common semiconductors, such as gallium arsenide, silicon and germanium, the ionization coefficients of both holes and electrons vary rapidly with the electric field and the electric field is determined in the first instance by the spatial distribution of the semiconductor doping. At d.c. breakdown the spatial distribution of the electric field and the mean ionization coefficient, $\langle\alpha\rangle$, for electrons in silicon is shown in Figs. 2(b) to (e) for some simple $p^+nn^+$ junction diode structures, which have heavily doped electrodes separated by 5 µm of uniformly doped n-type semiconductor (Fig. 2(a)). The effect of the doping concentration is to produce a triangular field distribution at the higher doping level of $5 \times 10^{15}/cm^3$ and a more constant field at the lower doping level of $1 \times 10^{15}/cm^3$. For the higher doped diode the depletion layer has just punched through, while for the other diode shown the field is not very much different from the constant field of a p-i-n diode with a breakdown voltage, $V_B$, of 140 V.

The conduction current is an exponential function of the ionization coefficient, which in silicon varies as the sixth power of the electric field. Consequently the higher-doped diode has a local avalanche region, whilst in the other diode impact ionization occurs throughout the n-region. The former gives rise to a high value of small signal negative resistance (because of the good phase relationship between the cyclic current pulse and the r.f. voltage applied), but this value changes rapidly with increasing r.f. voltage because of the larger effect of space charge. On the other hand, the lower-doped diode gives a low value of negative resistance, which is less dependent on the r.f. voltage.

The effect of a r.f. voltage corresponding to a power level of $\frac{1}{2}$ W is shown by the dotted lines. At a doping level of $5 \times 10^{15}/cm^3$ it is seen to modulate the width of the avalanche zone and also the width of the depletion region. This varying depletion layer gives rise to a reactance variation, an introduction of additional series resistance,[3] and back injection from the $n^+$ terminal. These effects do not occur with the other diode.

In the experimental investigations of impatt amplifiers the diode chips were encapsulated in S4 packages. To a first approximation[4] this is equivalent to shunting the chip with a low-pass $\pi$-filter network consisting of a series inductance of about 0·7 nH and two equal capacitors of 0·1 pF in parallel. More accurate representations of the equivalent circuit of the same package have recently been published[5] but for the present work the above representation should suffice. Neglecting the positive series resistance (due to the bulk semiconductor, depletion layer edge, contacts, and skin- and surface-effects), Grieling and Haddad's data[6] for the suspectance and conductance of $p^+nn^+$ silicon impatt diodes with n-regions 5 µm wide were used to calculate the impedance of a range of packaged diodes. The data herein correspond to chips with mesas of 125 µm diameter and d.c. bias currents of 65 mA. The theoretical impedances of diodes with $n = 0{\cdot}01$, $1{\cdot}0$, $3{\cdot}0$, $5{\cdot}0 \times 10^{15}/cm^3$ were calculated for frequencies from 8 to 16 GHz at r.f. powers of up to about $\frac{1}{2}$ W.

The values of negative resistance obtained for the various diodes at frequencies of 10, 12 and 14 GHz are given in Fig. 3. An examination of the data obtained at the two extreme levels of doping considered shows that

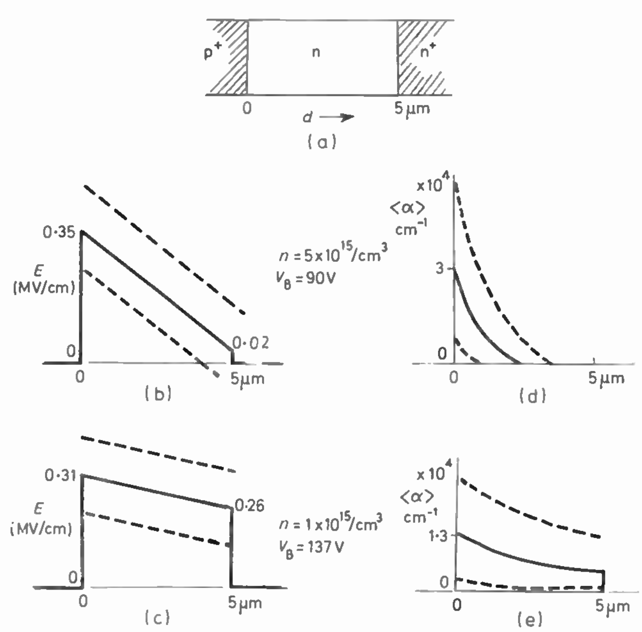


Fig. 2. Field profiles (b and c) and spatial distribution of the ionization coefficients (d and e) for a simple $p^+nn^+$ silicon diode structure (a), with $n = 5 \times 10^{15}$ and $1 \times 10^{15}/cm^3$ respectively.
solid lines—at d.c. breakdown
dotted lines—with 0·5 W r.f. in diode

a large value of small-signal negative resistance is accompanied by a rapid change of resistance with r.f. power, while a small change of negative resistance accompanies the low value of the negative resistance of lowly doped diodes. It is also seen that at high power levels the negative resistance of a diode with $n = 3 \times 10^{15}/\text{cm}^3$ is about the same as that with $n = 1 \times 10^{15}/\text{cm}^3$. For each diode the small-signal resistance decreases with increasing frequency over the range 10 to 14 GHz and at 10 GHz falls more rapidly with r.f. power than at the other two frequencies shown.

Similar data obtained for the rate of change of reactance with power showed that, as expected, the lowest doped diode showed the least change.

## 4 Theoretical Calculation of the Linearity of Amplifiers Using Various Silicon Diodes

The above data have been used to calculate theoretical amplifier responses. There are a number of circuits and technologies that may be used to make impatt diode amplifiers. But the convenience of coaxial circuits in experimental investigations and the fact that the S4 packaged diode is best characterized in a coaxial line suggested that the design should be coaxial. In the calculation of theoretical amplifier responses the design frequency was taken to be 12·0 GHz and the design gain for a three quarter-wave prototype transformer was taken to be 10 dB. This is typical of an amplifier output stage. The bias current for each diode was taken to be 65 mA and the diode was series resonated at about 200 mW. A computer was programmed to calculate the gain versus power at 12 GHz for a range of circuits. The lengths of the three steps of the transformer were varied from 0·09 mm in increments of 0·09 mm to 1·71 mm and, for each transformer, the gain versus output power curve was found. This was repeated for each of the four diodes in turn. The gain-power output curves at 12·0 GHz for various transformers for the diodes with $n = 1$ and $5 \times 10^{15}/\text{cm}^3$ are shown in Figs. 4(a) and (b). The curves plotted are for those transformers showing the 'most constant' gain, together with the response of the

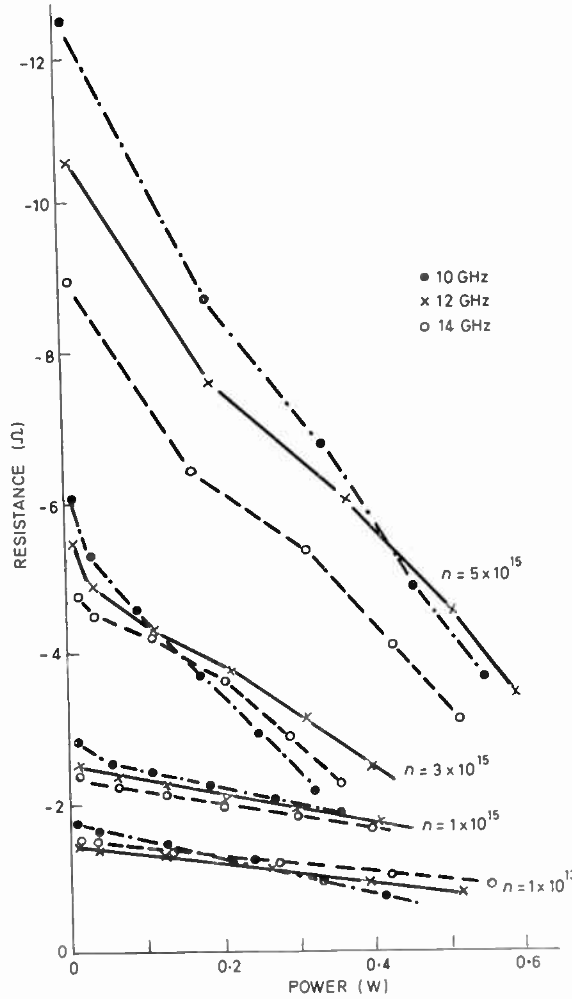


**Fig. 3.** Theoretical values of the negative resistance of packaged $p^+nn^+$ impatt diodes versus the r.f. diode power, for various doping levels and for various frequencies.

The chip data are taken from the work of Greiling and Haddad[6] and the diode is 125 μm in diameter; the n-region is of width 5 μm and the d.c. bias current is 65 mA. The package is assumed to be the S4 type.

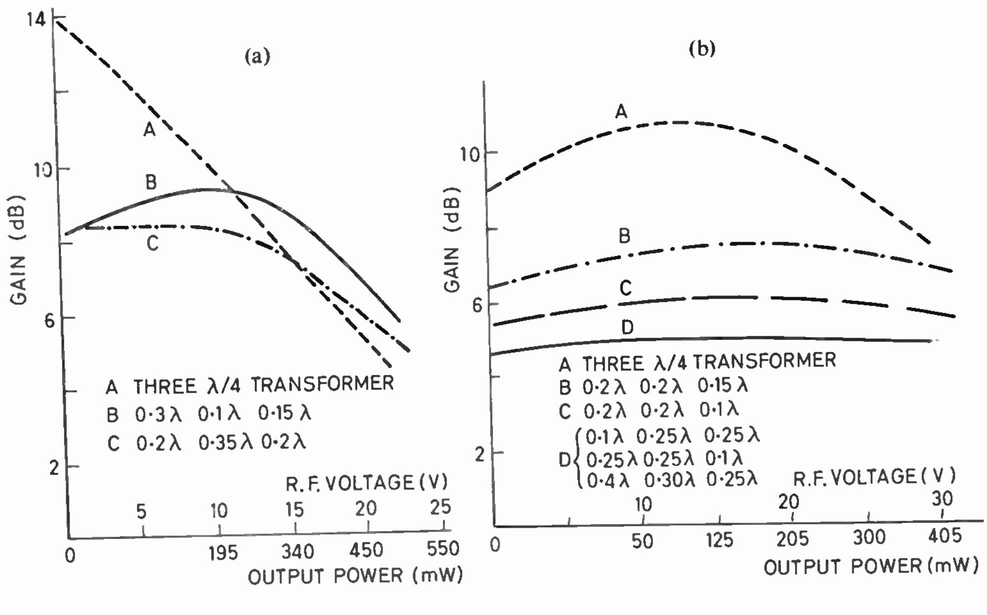


**Fig. 4.** (a) Gain versus output power curves for theoretical ½ W amplifiers at a frequency of 12 GHz using $p^+nn^+$ diodes with $n = 5 \times 10^{15}/\text{cm}^3$. The diode is resonated at ~200 mW and the data are presented for some of the transformers that gave the 'most constant' gain, together with data for a 10 dB gain, three λ/4 step transformer prototype for comparison.

(b) As for Fig. 4(a) but with $n = 1 \times 10^{15}/\text{cm}^3$.

three quarter-wave transformer for comparison. It can be seen that for $p^+nn^+$ diodes with $n = 5 \times 10^{15}/cm^3$ there is no transformer that gives a linear gain over the range 0 to 0·5 watt. For the diode with $n = 1 \times 10^{15}/cm^3$ the gain has a maximum value of 6 dB, and is constant to within $\pm 0{\cdot}5$ dB, for a transformer with 0·2 $\lambda$, 0·2 $\lambda$ and 0·1 $\lambda$ steps, where $\lambda$ is the wavelength at 12 GHz. Some other transformers also gave a similar linearity. Higher gain was accompanied by less linearity while better linearity could be obtained with lower gain. Calculations for the p-i-n diode also gave similar results, i.e. a best linearity to within $\pm 0{\cdot}5$ dB corresponding to a gain of 6 dB. For the diode with $n = 3 \times 10^{15}/cm^3$ linearity to within $\pm 0{\cdot}5$ dB at powers of up to 0·5 W could only be achieved at a gain of about 5 dB.

The transformers that gave the most constant gain up to 0·5 W were then used to calculate the amplifier gain versus frequency response, as a function of the r.m.s. r.f. voltage across the diode. Two examples of such plots are given in Figs. 5(a) and (b) and correspond respectively to amplifiers with $n = 5$ and $1 \times 10^{15}/cm^3$. At 12·0 GHz the better linearity of the lower-doped diode can be seen by comparing the shapes of the curves. However, if, instead of comparing the linearity at the frequency of 12 GHz, the comparison is made of linearity at a maximum gain of 6 dB, it can be seen that for the diode with $n = 5 \times 10^{15}/cm^3$, 6 dB gain is achieved at $\sim 11{\cdot}6$ GHz but is linear to only $\pm 0{\cdot}8$ dB and at a power corresponding to 20 V the gain is falling rapidly with increasing power. If the requirement is that the gain is constant to $\pm 0{\cdot}5$ dB then the maximum gain is lower at 4 dB, and, for the case shown, corresponds to a frequency of 11·0 GHz.

A similar, but more restricted, numerical study has recently been reported[7] wherein it is concluded that the effects of doping profile on amplifier performance is not appreciable. However, this work was not particularly concerned with the linearity of the amplifier, and did not optimize the circuits for linearity. This present investigation suggests that the use of a lowly-doped diode may be worthwhile if constant gain is required at the highest possible power levels.

## 5 Characteristics of Experimental Amplifiers

Brief details of some of our experimental amplifiers using $p^+nn^+$ diodes have already been given[8] so that this present Section considers only those properties that appertain directly to linear amplification. The diodes were AEI DC 1100-10 silicon 70 V diodes, which were power-limited to 100 mW (better, but less available, diodes were rated at over 0·5 W). These diodes have a 6 μm long n-region, doped to $7 \times 10^{15}/cm^3$, were about 120 μm in diameter and were packaged in S4 encapsulations. The small-signal frequency variation of the impedance of a typical diode is shown as a function of bias current in Fig. 6 and shows good qualitative agreement with small-signal theories. The operating frequencies of the diode shown was in the region 8 to 10 GHz; below this range the impedance variation was too large for the diodes to be useful in broadband applications. Above this range the negative resistance fell rapidly to zero. The negative $Q$-factor was about 12 and the series resistance of the diode at a voltage just below breakdown was 1·8 Ω.

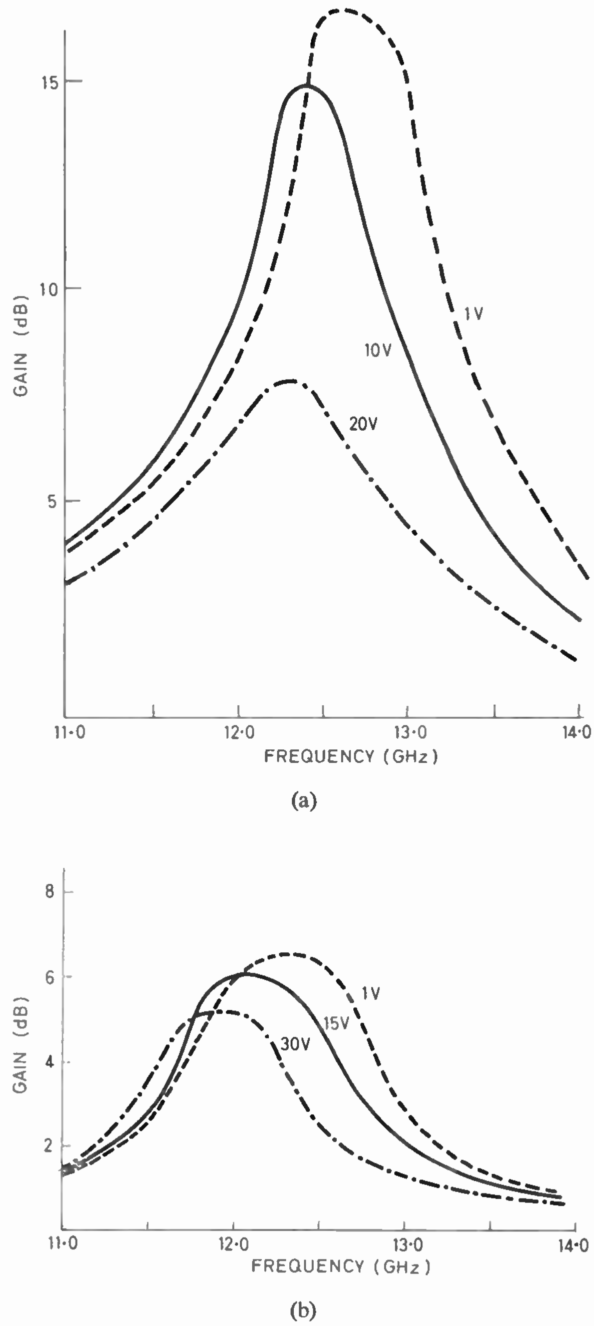


**Fig. 5.** (a) Theoretical amplitude response of impatt diode amplifier with $n = 5 \times 10^{15}/cm^3$. The transformer has three steps, 0·3 $\lambda$, 0·1 $\lambda$ and 0·15 $\lambda$ long at 12 GHz (see Fig. 4(a)).

(b) As for Fig. 5(a) but with $n = 1 \times 10^{15}/cm^3$. The transformer had 0·2 $\lambda$, 0·2 $\lambda$, 0·1 $\lambda$ steps.

The measured amplitude response of a 9 GHz, three $\lambda/4$ transformer amplifier is shown in Fig. 7, for various input drive powers. The small-signal response agreed well with that calculated from the experimental values of small-signal impedance. A comparison of the large-signal response of the experimental diode amplifier with the theoretical amplifier response is desirable. However, an extrapolation of the theoretical data of Greiling and

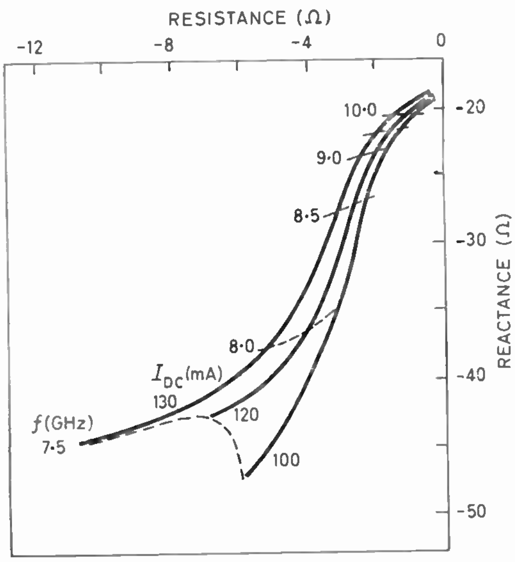


**Fig. 6.** Experimental small-signal impedance of an impatt diode, type DC 1100-10 as a function of frequency and bias current at a r.f. power input of 1 mW. The diode is a silicon $p^+nn^+$ diode with $n = 7\times10^{15}/cm^3$.

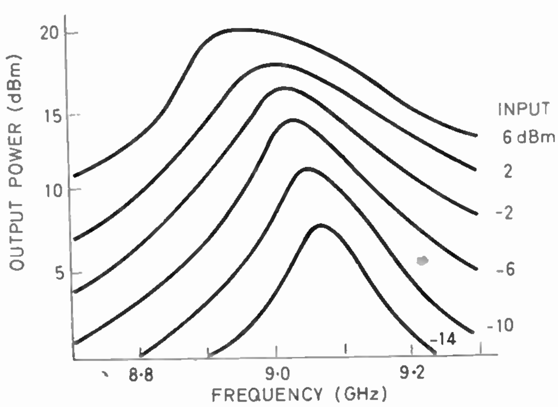


**Fig. 7.** Amplitude response of an experimental 9·0 GHz impatt diode amplifier using the DC 1100-10 diode of Fig. 6. $I_{DC} = 130$ mA, $V_{DC} = 75$ V and the transformer is composed of three $\lambda/4$ steps.

Haddad[6] to obtain conductance and susceptance data for diodes with $n = 7\times10^{15}/cm^3$ gave large errors, and since frequency scaling was also required (because of the differences in the widths of the $n$ regions of the experimental and theoretical diodes) a calculation of the theoretical amplifier response was not considered very meaningful. An examination of the data presented in Fig. 7 shows that at 9·0 GHz the gain compresses slowly with increasing drive. The requirement of linearity would thus limit the power and efficiency of the amplifier. However, at a frequency of 8·9 GHz there is linear gain up to the thermal power limit of the diode. An investigation of the third order intermodulation products of two equal tones at 9·00 and 9·03 GHz showed that at the 1 dB compression point the signal at 9·06 GHz was 18 dB below the single tone level, while the signal at 8·97 GHz was 13 dB down, the difference arising from the shift of the amplifier response with r.f. power.

The quarter-wave step transformer restricts the bandwidth of the amplifier and so the transformer was broadbanded by altering the lengths of the individual steps. The amplitude response of such an amplifier is shown in Fig. 8, while a schematic diagram of the amplifier is shown in Fig. 9. The 50 Ω coaxial line was 7 mm in diameter, with the diode placed in a 1·6 mm resonating recess at the end of the line. A 0·08 cm thick quarter-wave p.t.f.e. tuning slug was moved along the line to give the optimum amplifier response at an input power of 6 dBm. Linear gain ($\pm\frac{1}{2}$ dB) was obtained only over about 300 MHz of the total bandwidth, with distortion occurring at the band edges as the r.f. drive was decreased.

Preliminary experimental attempts have been made to demonstrate linear amplification using diodes with low doping levels. p-i-n and step recovery diodes are readily available, and the diodes investigated were 140 V, AEI type number DC 2109E. These diodes had intrinsic regions of ~ 5 μm and the doping level corresponded to values of $n \sim 0{\cdot}1\times10^{15}/cm^3$. The diode junction diameter was ~ 100 μm and the chips were mounted in S4 packages. However these diodes were not optimized for impatt operation and at 10 GHz were found to have a series resistance at breakdown of about 2·5 ohms. This value is slightly greater than the theoretical value of the negative resistance and attempts to determine the experimental value of the negative resistance, by designing an oscillator, failed. However it is worth noting that the technological improvements in diode fabrication over the last three years, or so, have resulted in the best conventional impatt diodes having series resistances as low as 0·4 ohm at breakdown.

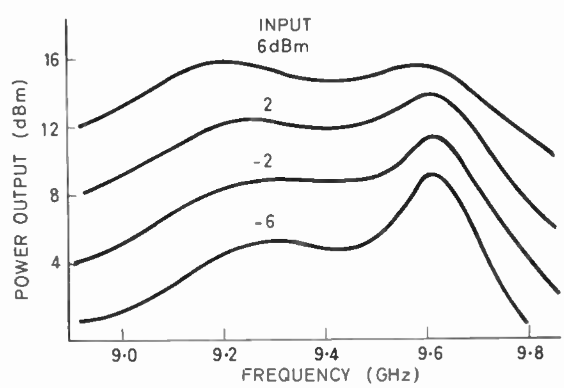


**Fig. 8.** Broadbanded amplitude response as a function of r.f. drive at a constant bias current of 120 mA at 73 V.

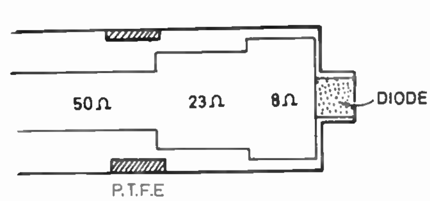


**Fig. 9.** Schematic diagram of broadbanded two-step transformer amplifier with a moveable p.t.f.e. tuning slug. The 8 Ω and 23 Ω steps are 0·16 λ and 0·24 λ long while the slug is 0·25 λ long where λ is the wavelength for 9·0 GHz.

## 6 Discussion and Conclusions

The use of passive circuitry satisfactorily to match an impedance that changes with r.f. power can be accomplished empirically only within certain restrictions. In the case of impatt amplifiers satisfactory matching for the design of linear amplifiers can be obtained by the use of a conventional oscillator diode (doped at $\sim 7 \times 10^{15}/\text{cm}^3$ at X-band), and working the amplifier at a lower frequency than that corresponding to the peak gain. Limited broadbanding can be achieved by using empirical circuits, such as transformers with staggered steps.

It is also shown numerically that, for 3-section stepped impedance transformers, amplifier linearity is best achieved with diodes that are heavily punched through at breakdown. However in the interpretation of the numerical results it must be noted that for all the diodes considered the d.c. bias current was taken as 65 mA. Now the diode with $n = 5 \times 10^{15}/\text{cm}^3$ was operated at 95 V, while the diode which has a value of $n = 1 \times 10^{15}/\text{cm}^3$ was operated at 138 V. If the former diode was operated at the same d.c. input power, namely 95 mA, the negative resistance would be larger, as would the rate of change of negative resistance with r.f. power. Approximate computations using data extrapolated from both our own amplifier work and from the work of Van Iperen and Tjassens,[3] and normalized to the data of Greiling and Haddad,[6] have shown that for the diode with $n = 5 \times 10^{15}/\text{cm}^3$ the linearity is somewhat worse at the increased bias currents of 95 mA.

## 7 Acknowledgments

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# Synthesis of wideband cascaded double-pole crystal lattice filters

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SUMMARY

A design procedure is given for crystal lattice filters with bandwidths for quartz of approximately 1 to 8% of the centre frequency; the method results in simple formulae in terms of the elements of a low-pass ladder whose order is a multiple of four.

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## List of Principal Symbols

| | |
|---|---|
| $a$ | inverse fractional bandwidth of a band-pass network ($a = \omega_0/\Delta\omega$) |
| $D$ | ratio of input to output decrements |
| $d$ | decrement, or inverse $Q$-factor, for a low-pass reactance |
| $g$ | element in normalized low-pass ladder network |
| $\kappa$ | source resistance for normalized low-pass ladder network |
| $n$ | order of network, which equals twice the number of crystal resonators (Jaumann lattice) |
| $\mathrm{j}R$ | frequency independent reactance |
| $R_\mathrm{T}$, $R_\pi$ | values of $R$ resulting from an impedance inverter realized in T and $\pi$ forms |
| $\omega_c$ | band-pass cut-off frequency |
| $\omega_0$ | band-pass centre frequency |
| $\Delta\omega$ | bandwidth |

## 1 Introduction

The earliest band-pass filters[1] were composed of inductors and capacitors, and consisted of ladder structures image matched at the centre frequency.† Since inductors are bulky and lossy, various substitute elements and design methods have been used in an attempt to reduce their number or avoid these elements completely, including replacement with piezoelectric and mechanical resonators, and elimination using active and digital techniques.

Piezoelectric resonators near resonance have an equivalent circuit (in terms of lumped elements) consisting of a static capacitor in parallel with a motional reactance consisting of an inductor and capacitor in series. Quartz is especially favoured because of its very low internal dissipation. The reactance of resonators is positive within a narrow frequency band defined by the series and parallel resonant frequencies, and is negative elsewhere. This property makes them ideal elements for use in a lattice, whose image impedance is given by the geometric mean of the arm impedances.‡ A considerable time elapsed before rigorous analysis and synthesis techniques were applied,[3,4] and more recent image parameter methods generally result in ladder networks[5-7] whose characteristics are more readily distorted by resistive losses.

† Network sections are image matched if the input impedances are equal in the forward and return directions at each termination (considered separately); if, in addition, all the image impedances are real, i.e. resistive, maximum power transfer occurs. It is usually not possible to satisfy this criterion exactly at more than one frequency.

‡ A pass-band is obtained when the image impedance for the lattice is resistive, i.e. one arm is inductive and the other capacitive. A design for two crystals is easily obtained using this principle.[2] Let $\omega_{-c}$, $\omega_0$, $\omega_c$ denote the required lower cut-off, centre, and upper cut-off frequencies, respectively; the crystals are then adjusted so that the series resonant, parallel resonant frequencies occur at $\omega_{-c}$, $\omega_0$, and $\omega_0$, $\omega_c$ for the two crystals respectively. Below $\omega_{-c}$, both crystals are capacitive, between $\omega_{-c}$ and $\omega_c$ one is inductive and the other capacitive, and above $\omega_c$, both are inductive.

Designs using the insertion loss method are preferable, since algebraic expressions for optimum charactetistics may be utilized. The problem now is to generate the corresponding network, and this may be formidable if, as in our case, the resulting structure is limited to configurations restricted in form and element values. Several design methods for a single lattice are available,[8–13] but these are long and tedious computationally; a more direct method is given by Bown.[14]

A simpler and more versatile alternative, as considered here, is to synthesize a low-pass ladder network and apply frequency and impedance transformations until the resulting network is in a form suitable for accepting quartz crystals. The first stage can usually be eliminated since element values of low-pass ladders are readily available.[15–17] A wideband design† using this approach has been given by Bown,[18] but is restricted to a symmetric ladder of order seven, and uses four crystals and three inductors. O'Meara's wideband design[19] of order eight requires an antimetric ladder and uses four crystals and four inductors.

The procedure given here is a modification and extension of O'Meara's method for any network whose order is a multiple of four, and the restriction on element values no longer applies; the resulting solution is for crystals of equal motional capacitance (or inductance). Also considered are limitations in the lumped-element approximation to the motional reactance of a piezoelectric resonator.

## 2 Design Method

### 2.1 *Development of Low-pass Analogue*

A low-pass network normalized with respect to load termination $R_L$ and cut-off frequency $\omega_L$ will be first considered. Using the symbolic representation shown in Fig. 1, an impedance inverter (see Appendix 1) is inserted after each pair of elements as shown in Fig. 2.† $n/2-1$ inverters are required where $n$ is a multiple of 4. To set the inductors equal we require

$$g_2 = g_3R_{23}^2 = g_6\left(\frac{R_{45}}{R_{23}}\right)^2 = g_7\left(\frac{R_{23}R_{67}}{R_{45}}\right)^2$$
$$= g_{10}\left(\frac{R_{45}R_{89}}{R_{23}R_{67}}\right)^2 = g_{11}\left(\frac{R_{23}R_{67}R_{10,11}}{R_{45}R_{89}}\right)^2 = \ldots$$
$$= g_{n-2}\left(\frac{R_{45}R_{89}\ldots R_{n-4,n-3}}{R_{23}R_{67}\ldots R_{n-6,n-5}}\right)^2$$
$$= g_{n-1}\left(\frac{R_{23}R_{67}\ldots R_{n-2,n-1}}{R_{45}R_{89}\ldots R_{n-4,n-3}}\right)^2.$$

Therefore,

$$R_{23} = \left(\frac{g_2}{g_3}\right)^{\frac{1}{2}} = \frac{g_2}{(g_2g_3)^{\frac{1}{2}}}$$
$$R_{45} = R_{23}\left(\frac{g_2}{g_6}\right)^{\frac{1}{2}} = \frac{g_2}{(g_3g_6)^{\frac{1}{2}}}$$
$$R_{67} = \frac{R_{45}}{R_{23}}\left(\frac{g_2}{g_7}\right)^{\frac{1}{2}} = \frac{g_2}{(g_6g_7)^{\frac{1}{2}}}$$
$$R_{89} = \frac{R_{23}R_{67}}{R_{45}}\left(\frac{g_2}{g_{10}}\right)^{\frac{1}{2}} = \frac{g_2}{(g_7g_{10})^{\frac{1}{2}}}.$$

The inverters are realized in alternate T and $\pi$ forms, indicated by a superscript. Hence

$$R_{23}^{T} = \frac{g_2}{(g_2g_3)^{\frac{1}{2}}} \qquad R_{45}^{\pi} = \frac{g_2}{(g_3g_6)^{\frac{1}{2}}}$$
$$R_{67}^{T} = \frac{g_2}{(g_6g_7)^{\frac{1}{2}}} \qquad R_{89}^{\pi} = \frac{g_2}{(g_7g_{10})^{\frac{1}{2}}}$$
$$\vdots$$
$$R_{n-2,n-1}^{T} = \frac{g_2}{(g_{n-2,n-1})^{\frac{1}{2}}} \qquad R_{n-4,n-3}^{\pi} = \frac{g_2}{(g_{n-5}, g_{n-2})^{\frac{1}{2}}}$$

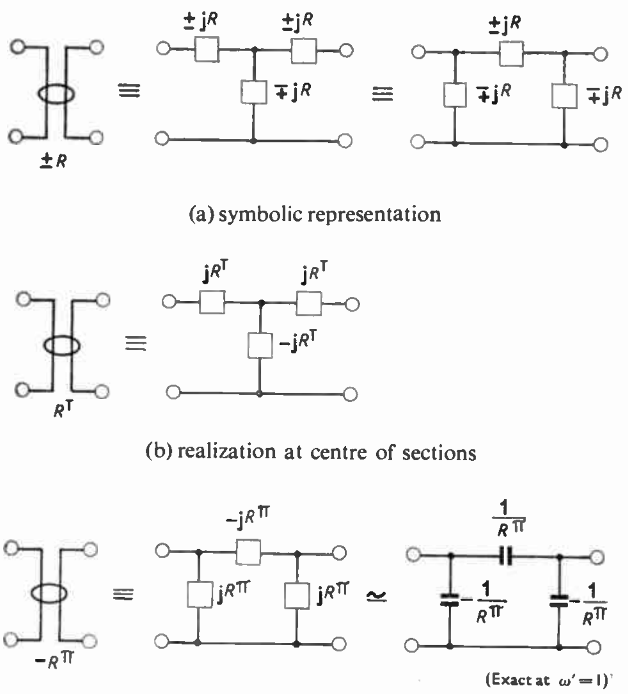
The T-inverter at the centre of the $v$th four-crystal section (from the load), and the $\pi$-inverter which separates the $v$th and $(v+1)$th sections are as follows,

$$\left.\begin{aligned} R_{4v-2,4v-1}^{T} &= \frac{g_2}{(g_{4v-2}g_{4v-1})^{\frac{1}{2}}} \\ R_{4v,4v+1}^{\pi} &= \frac{g_2}{(g_{4v-1}g_{4v+2})^{\frac{1}{2}}} \end{aligned}\right\} \qquad (1)$$

The terminating impedance $\kappa_D$ is given by

$$\kappa_D = \frac{1}{\kappa}\left(\frac{R_{23}R_{67}\ldots R_{n-2,n-1}}{R_{45}R_{89}\ldots R_{n-4,n-3}}\right)^2 = \frac{1}{\kappa}\frac{g_2}{g_{n-1}}$$

The input and output decrements (inverse $Q$-factors) are



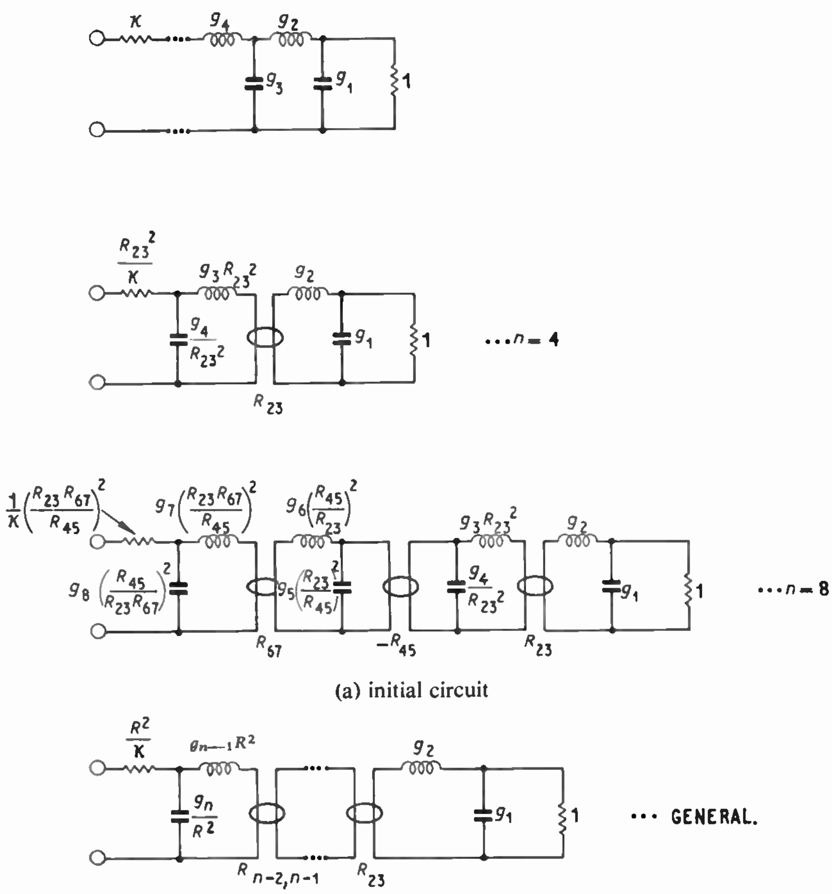

(a) symbolic representation

(b) realization at centre of sections

(c) realization for separating sections.

**Fig. 1.** Impedance inverters.

† The static (or parallel) capacitance associated with resonators severely restricts the bandwidth, which for crystals alone can rarely exceed $\omega_0/r$, where $\omega_0$ is the centre frequency and $r$ the static to motional capacitance ratio; for fundamental mode quartz, $r \simeq 250$, and this sets an upper bandwidth limit of 0·4% of the centre frequency. The bandwidth may be extended using inductors, and such filters are usually referred to as wide band.

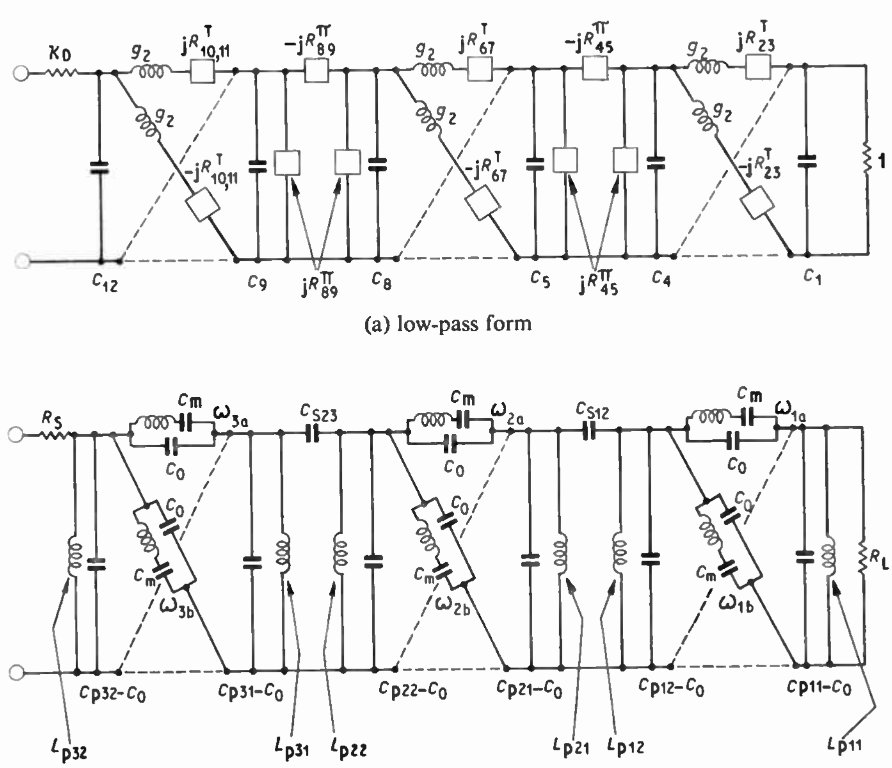



(a) initial circuit

$$R = \frac{R_{23}\,R_{67}\cdots R_{n-2,\,n-1}}{R_{45}\,R_{89}\cdots R_{n-4,\,n-3}}$$

(b) addition of inverters.

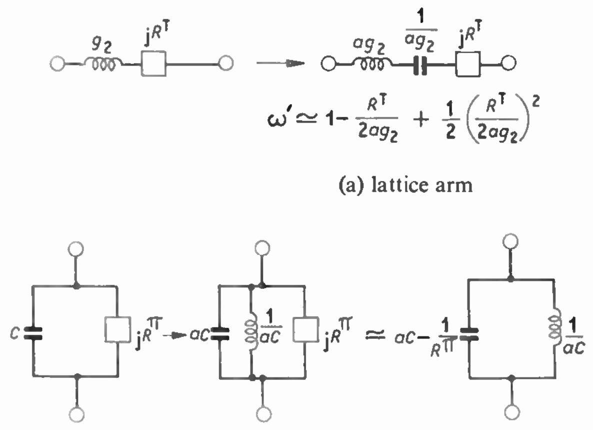
Fig. 2. Low-pass ladder.



(a) low-pass form

(b) band-pass form

Fig. 3. Lattice representation of network ($n = 12$).




(a) lattice arm

(b) parallel arm

**Fig. 4.** Frequency transformations (normalized notation).

given as follows,

$$d_n = \frac{\kappa}{g_n}, \quad d_1 = \frac{1}{g_1}$$

With $D(=d_n/d_1)$ as the decrement ratio we obtain

$$\kappa_D = \frac{1}{D}\cdot\frac{g_1 g_2}{g_{n-1} g_n} \qquad (2)$$

From Fig. 2(b) the parallel capacitors for the first two sections are given as follows,

$$C_1 = g_1, \quad C_4 = \frac{g_4}{R_{23}^2} = \frac{g_3 g_4}{g_2} \qquad \text{section 1}$$

$$C_5 = \left(\frac{R_{23}}{R_{45}}\right)^2 g_5 = \frac{g_5 g_6}{g_2},$$

$$C_8 = \left(\frac{R_{45}}{R_{23}R_{67}}\right)^2 g_8 = \frac{g_7 g_8}{g_2} \qquad \text{section 2}$$

For the $v$th section

$$C_{4v-3} = \frac{g_{4v-3}g_{4v-2}}{g_2}, \quad C_{4v} = \frac{g_{4v-1}g_{4v}}{g_2} \qquad (3)$$

with capacitor subscripts relating to the corresponding elements in the original ladder. After a ladder-to-lattice transformation performed using Bartlett's bisection theorem,† and a T-inverter realization as shown in Fig. 1(b), the circuit shown in Fig. 3(a) is obtained.

### 2.2 *Application of Frequency Transformation*

The low-pass to band-pass transformation is given by

$$\{\omega'\}_{\mathrm{LP}} \quad \left[a\left(\omega' - \frac{1}{\omega'}\right)\right]_{\mathrm{BP}}$$

and elements are transformed as shown in Fig. 4. With normalized frequencies indicated by a prime, normalization is now with respect to centre frequency, $\omega_0$, instead of $\omega_{\mathrm{L}}$, and $a$ is defined as follows,

$$a = \frac{\omega_0}{\Delta\omega} = \frac{\omega_0}{\omega_c - \omega_{-c}}$$

where $\omega_{-c}$, $\omega_c$ denote the lower and upper cut-off frequencies. The modified parallel capacitors at each end of the $v$th section are now as follows,

$$C_{4v-3}^{\mathrm{BP}} = aC_{4v-3} - \frac{1}{R_{4v-4,\,4v-3}^{\pi}}$$

$$C_{4v}^{\mathrm{BP}} = aC_{4v} - \frac{1}{R_{4v,\,4v+1}^{\pi}}$$

where

$$R_{01}^{\pi} = R_{n,\,n+1}^{\pi} = 0.$$

From eqns. (1) and (3)

$$\left.\begin{aligned} C_{4v-3}^{\mathrm{BP}} &= \frac{ag_{4v-3}g_{4v-2} - (g_{4v-5}g_{4v-2})^{\frac{1}{2}}}{g_2} \\ C_{4v}^{\mathrm{BP}} &= \frac{ag_{4v-1}g_{4v} - (g_{4v-1}g_{4v+2})^{\frac{1}{2}}}{g_2} \end{aligned}\right\} \qquad (4)$$

where $g_{-1} = g_{n+2} = 0$. The adjacent inductors are as follows:

$$\left.\begin{aligned} L_{4v-3}^{\mathrm{BP}} &= \frac{1}{aC_{4v-3}} = \frac{g_2}{ag_{4v-3}g_{4v-2}} \\ L_{4v}^{\mathrm{BP}} &= \frac{1}{aC_{4v}} = \frac{g_2}{ag_{4v-1}g_{4v}} \end{aligned}\right\} \qquad (5)$$

The series capacitor separating the $v$th and $(v+1)$th sections is given from eqn. (1) and Figs. 1(c) and 3(a) as follows:

$$C_{4v,\,4v+1}^{\mathrm{BP}} = \frac{(g_{4v-1}g_{4v+2})^{\frac{1}{2}}}{g_2} \qquad (6)$$

### 2.3 *Crystal Resonant Frequencies*

For the circuit in Fig. 4(a), the normalized series resonant frequency, $\omega'$, is given by

$$jag_2\left(\omega' - \frac{1}{\omega'}\right) + jR^{\mathrm{T}} = 0$$

Hence‡

$$\begin{aligned} \omega' &= -\frac{R^{\mathrm{T}}}{2ag_2} + \left[1 + \left(\frac{R^{\mathrm{T}}}{2ag_2}\right)^2\right]^{\frac{1}{2}} \\ &\simeq 1 - \frac{R^{\mathrm{T}}}{2ag_2} + \tfrac{1}{2}\left(\frac{R^{\mathrm{T}}}{2ag_2}\right)^2 \end{aligned} \qquad (7)$$

### 2.4 *Design Formulae*

With normalized values denoted by a prime, de-normalization is performed as follows,

$$R = R_{\mathrm{L}}.R', \quad L = \frac{R_{\mathrm{L}}}{\omega_0}.L', \quad C = \frac{1}{\omega_0 R_{\mathrm{L}}}.C', \quad \omega = \omega_0.\omega'$$

The final circuit is given in Fig. 3(b).

† Bartlett's bisection theorem[20] enables any symmetrical network to be converted into lattice form. With the network divided at the centre of symmetry, the lattice series and shunt arms are given by the short-circuit and open-circuit input impedances of the half-section, respectively.

‡ We have considered the motional reactance of a piezo-electric resonator in terms of lumped elements; if the more accurate distributed transmission-line model is used, (as described in Appendix 2), the squared term in eqn. (7) is eliminated.

Using eqns. (4) and (5), the parallel capacitor and inductor at each end of the $v$th section are given as follows:

$$\left.\begin{aligned} C_{pv1} &= \frac{1}{\omega_0 R_L}\cdot\frac{ag_{4v-3}g_{4v-2}-(g_{4v-5}g_{4v-2})^{\frac{1}{2}}}{g_2} \\ L_{pv1} &= \frac{R_L}{\omega_0}\cdot\frac{g_2}{ag_{4v-3}g_{4v-2}} \end{aligned}\right\}\begin{array}{l}\text{towards}\\ \text{load}\end{array} \quad (8)$$

$$\left.\begin{aligned} C_{pv2} &= \frac{1}{\omega_0 R_L}\cdot\frac{ag_{4v-1}g_{4v}-(g_{4v-1}g_{4v+2})^{\frac{1}{2}}}{g_2} \\ L_{pv2} &= \frac{R_L}{\omega_0}\cdot\frac{g_2}{ag_{4v-1}g_{4v}} \end{aligned}\right\}\begin{array}{l}\text{towards}\\ \text{source}\end{array} \quad (9)$$

where

$$g_{-1} = g_{n+2} = 0, \quad a = \frac{\omega_0}{\Delta\omega}$$

The series capacitor separating the $v$th and $(v+1)$th sections is given by eqn. (6) as

$$C_{s(v,\,v+1)} = \frac{1}{\omega_0 R_L}\cdot\frac{(g_{4v-1}g_{4v+2})^{\frac{1}{2}}}{g_2}$$

The motional capacitance is given by Fig. 4(a) as follows,

$$C_m = \frac{1}{\omega_0 R_L}\cdot\frac{1}{ag_2} \quad (10)$$

If motional inductance, $L_m$, is a design parameter, $C_m$ is replaced by

$$L_m = \frac{1}{\omega_0^2 C_m} = \frac{R_L}{\omega_0}\cdot ag_2$$

(see Appendix 3).

The frequencies of the crystals in the $v$th section are given by eqns. (1) and (7).

$$\omega_{va} = \omega_0\left[1 - \frac{1}{2a(g_{4v-2}g_{4v-1})^{\frac{1}{2}}}\right]$$

$$\omega_{vb} = \omega_0\left[1 + \frac{1}{2a(g_{4v-2}g_{4v-1})^{\frac{1}{2}}}\right]$$

Finally, we have the source resistance given by eqn. (2) as follows:

$$R_s = R_L\cdot\frac{1}{D}\frac{g_1 g_2}{g_{n-1}g_n}$$

### 2.5 *Bandwidth*

The maximum bandwidth is determined by the parallel capacitors whose minimum value is zero. From eqns. (8) and (9) and Fig. 3(b)

$$C_0 \leqslant \frac{1}{\omega_0 R_L}\cdot\frac{az}{g_2}$$

where

$$z = \left[g_{4v-3}g_{4v-2} - \frac{1}{a}(g_{4v-5}g_{4v-2})^{\frac{1}{2}},\; g_{4v-1}g_{4v} - \frac{1}{a}(g_{4v-1}g_{4v+2})^{\frac{1}{2}}, \dots\right]_{\min} \quad 1 \leqslant v \leqslant \frac{n}{4}$$

With $C_m$ given by eqn. (10) the fractional bandwidth is limited as follows.

$$\frac{\Delta\omega}{\omega} = \frac{1}{a} \leqslant \left(\frac{z}{r}\right)^{\frac{1}{2}}$$

where $r(=C_0/C_m)$ is the crystal capacitance ratio. For quartz crystals operating in a fundamental mode, a typical value of $r$ is 250. Hence

$$\frac{\Delta\omega}{\omega} \leqslant 6{\cdot}32z^{\frac{1}{2}}\% \simeq 8\%$$

For Butterworth and Chebyshev responses with $D = 1$, $z = g_1 g_2$.

The minimum bandwidth must usually exceed a multiple of the inverse $Q$-factor of the coils which sets a lower limit of about 1%.

## 3 Conclusion

The simple formulae given here enable element values of a useful class of wideband crystal lattice filters to be directly obtained without recourse to circuit theory.

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## 6 Appendix 1: Impedance Inverters

Impedance inverters were introduced by Cohn,[22] who represented the equivalent circuit for a quarter wavelength of transmission line of characteristic impedance $R$ in symmetrical T or $\pi$ form, as shown in Fig. 1(a). The lumped elements are affected by impedance but not frequency transformations; j$R$ is therefore a convenient symbol, with $R$ treated as a conventional resistor in transformations. A practical representation in terms of lumped reactive elements is only valid at a particular frequency, but the variation is usually small except for applications with large bandwidths, e.g. $>10\%$ of centre frequency.

The transmission matrix for an inverter, as shown in Fig. 1(a), is as follows:

$$T=\begin{pmatrix}0 & \mathrm{j}R\\ \mathrm{j}/R & 0\end{pmatrix},\quad T^{-1}=\begin{pmatrix}0 & -\mathrm{j}R\\ -\mathrm{j}/R & 0\end{pmatrix}$$

The inverse matrix defines an inverter of the same characteristic impedance with series and shunt elements $-\mathrm{j}R$, $\mathrm{j}R$, respectively. A two-port network placed between such a pair is equivalent to a two-port network which is its dual, i.e.

$$\begin{pmatrix}0 & -\mathrm{j}R\\ -\mathrm{j}/R & 0\end{pmatrix}\begin{pmatrix}A & B\\ C & D\end{pmatrix}\begin{pmatrix}0 & \mathrm{j}R\\ \mathrm{j}/R & 0\end{pmatrix}=\begin{pmatrix}D & R^2C\\ B/R^2 & A\end{pmatrix}$$

It is not necessary for inverters to be used in pairs. If a two-port network is divided at any point and an inverter inserted, all the elements on one side (but not both) are converted into their duals. Attenuation and group delay are unaffected and the phase shift is modified by $\pm 90°$ which, being constant, is of no consequence. Use is made of this property in the given design procedure, and inverters are added individually as shown in Fig. 2. Impedance inversion converts a series inductor $L$ into a parallel capacitor $L/R^2$ and a parallel capacitor $C$ into a series inductor $CR^2$, where $R$ is the characteristic impedance. This is easily verified as follows:

$$\mathrm{j}\omega L\rightarrow\frac{R^2}{\mathrm{j}\omega L}=\frac{1}{\mathrm{j}\omega(L/R^2)}\ \ldots\text{Inductor } L\rightarrow\text{Capacitor } L/R^2$$

$$\frac{1}{\mathrm{j}\omega C}\rightarrow\frac{R^2}{1/\mathrm{j}\omega C}=\mathrm{j}\omega(CR^2)\ldots\text{Capacitor } C\rightarrow\text{Inductor } CR^2$$

## 7 Appendix 2: Equivalent Circuit of a Piezoelectric Resonator

The motional reactance of a piezoelectric resonator (crystal or polarized ceramic) near resonance is given from the distributed transmission model[21, 23–25] as follows,

$$X_{\mathrm{D}}=-\frac{4\omega_v L_{\mathrm{m}}}{\pi}\cot\frac{\pi\omega}{2\omega_v}=\frac{4\omega_v L_{\mathrm{m}}}{\pi}\tan\frac{\pi(\omega-\omega_v)}{2\omega_v}\tag{11}$$

where $\omega_v$ is the resonant frequency. This expression is usually approximated with lumped elements using an inductor $L_{\mathrm{m}}$ and capacitor $C_{\mathrm{m}}$ in series, resonating at $\omega_v$. We then have

$$X_{\mathrm{LC}}=\omega L_{\mathrm{m}}-\frac{1}{\omega C_{\mathrm{m}}}=\omega_v L_{\mathrm{m}}\left(\frac{\omega}{\omega_v}-\frac{\omega_v}{\omega}\right)\tag{12}$$

A more accurate expression is as follows:

$$X_R=2L_{\mathrm{m}}(\omega-\omega_v)\tag{13}$$

For wide bandwidths, the error in eqn. (12) becomes significant and it is preferable, if possible, to use eqns. (11) or (13) in place of eqn. (12) in an analysis program; an alternative is to modify the resonant frequency of a lumped element simulation to equate the reactances at the crystal centre frequency, as described in Appendix 3. For a frequency shift of 4% of the resonant frequency, for example, the errors in the approximations to the motional reactance resulting from eqns. (12) and (13) are 2%, 0·1% respectively.

Equations (11), (12) and (13) are most easily compared by expanding in polynomial form, i.e.

$$X_{\mathrm{D}}=\frac{4\omega_v L_{\mathrm{m}}}{\pi}\cdot\frac{\pi(\omega-\omega_v)}{2\omega_v}\left[1+\frac{\pi^2}{12}\left(\frac{\omega-\omega_v}{\omega_v}\right)^2+\frac{\pi^4}{120}\left(\frac{\omega-\omega_v}{\omega_v}\right)^4+\ldots\right]$$

$$=2L_{\mathrm{m}}(\omega-\omega_v)\left[1+\frac{\pi^2}{12}\left(\frac{\omega-\omega_v}{\omega_v}\right)^2+\ldots\right]$$

$$X_{\mathrm{LC}}=2L_{\mathrm{m}}(\omega-\omega_v)\left[1-\tfrac{1}{2}\left(\frac{\omega-\omega_v}{\omega_v}\right)+\tfrac{1}{2}\left(\frac{\omega-\omega_v}{\omega_v}\right)^2-\ldots\right]$$

Hence,

$$\frac{X_{\mathrm{LC}}-X_{\mathrm{D}}}{X_{\mathrm{D}}}\simeq\frac{\left[1-\tfrac{1}{2}\left(\frac{\omega-\omega_v}{\omega_v}\right)+\tfrac{1}{2}\left(\frac{\omega-\omega_v}{\omega_v}\right)^2\right]-\left[1+\frac{\pi^2}{12}\left(\frac{\omega-\omega_v}{\omega_v}\right)^2\right]}{1+\frac{\pi^2}{12}\left(\frac{\omega-\omega_v}{\omega_v}\right)^2}$$

$$\simeq-\tfrac{1}{2}\left(\frac{\omega-\omega_v}{\omega_v}\right)$$

$$\frac{X_R-X_{\mathrm{D}}}{X_{\mathrm{D}}}\simeq\frac{1-\left[1+\frac{\pi^2}{12}\left(\frac{\omega-\omega_v}{\omega_v}\right)^2\right]}{1+\frac{\pi^2}{12}\left(\frac{\omega-\omega_v}{\omega_v}\right)^2}\simeq-\frac{\pi^2}{12}\left(\frac{\omega-\omega_v}{\omega_v}\right)^2$$

Note that eqn. (11) applies to a resonator with the wave propagation perpendicular to the electric field, such as in length-compressional modes of bars or radial modes of disks. For wave propagation parallel to the electric field, such as in thickness shear or thickness compressional modes of plates, the reactance is modified by a series negative capacitor, but the difference in $X_D$ in the above example for fundamental mode quartz is less than 0·1 %, which is negligible.

## 8 Appendix 3: Motional Reactances of Crystals

It is usual to specify crystals of equal motional capacitance in preference to equal motional inductance. However, the latter enables impedance inverters to be realized entirely with elements of the same type (capacitors), but the improvement is usually negligible for bandwidths below about 4 % of the centre frequency. For larger bandwidths it is preferable to specify equal inductances and, for a reliable network analysis, the distributed transmission-line model should be used to represent the motional reactance of a piezoelectric resonator.[21] For an 8th-order Chebyshev filter of 1 dB ripple with a 4 % bandwidth and 3 dB attenuation at the cut-off frequencies, the maximum deviation in values of pass-band attenuation between the lumped and distributed models is 0·30 dB. To save using transcendental functions in an analysis program, the lumped element simulation of a crystal resonator is improved by modifying the resonant frequency to provide the same reactance as the resonator at the filter centre frequency. Ignoring third and higher order terms in $a$, we then obtain

$$\left.\begin{aligned} X_m &= \frac{4\omega_v L_m}{\pi}\tan\frac{\pi(\omega-\omega_v)}{2\omega_v} \\ \omega_v &= \omega_0\left[1 \mp \frac{1}{2a(g_{4v-2}g_{4v-1})^{\frac{1}{2}}}\right] \end{aligned}\right\}\text{motional reactance of piezoelectric resonator}$$

$$\left.\begin{aligned} X_{m1} &= \omega_{v1} L_m\left(\frac{\omega}{\omega_{v1}}-\frac{\omega_{v1}}{\omega}\right) \\ \omega_{v1} &= \omega_0\left[1 \mp \frac{1}{2a(g_{4v-2}g_{4v-1})^{\frac{1}{2}}} - \frac{1}{8a^2(g_{4v-2}g_{4v-1})}\right] \end{aligned}\right\}\text{lumped-element simulation}$$

where

$$X_{m|\omega=\omega_0} = X_{m1}\big|_{\omega=\omega_0} = \frac{\pm\,\omega_0}{a(g_{4v-2}g_{4v-1})^{\frac{1}{2}}} L_m$$

By using $\omega_{v1}$ in place of $\omega_v$ to simulate a resonator, the maximum deviation in the values of pass-band attenuation for the Chebyshev example is reduced to 0·07 dB.

*Manuscript first received by the Institution on 30th April 1973 and in final form on 21st June 1973. (Paper No. 1549/CC 182.)*

# IERE News and Commentary

## Dinner of Council and Committees

Members are reminded that the 15th Dinner of the Council and its Committee will take place at the Savoy Hotel, London, on the evening of Wednesday, 21st November.

There are still a few tickets available, and any member wishing to attend, who has not yet completed and returned the form which was enclosed in the September 1973 issue, should send it to the Institution without delay.

## IERE Christmas Cards 1973

Orders are now being received for IERE Christmas cards, details of which were given in the September issue of the *Journal* (pages (iv) and (v)). The final date for receipt of orders from members in the U.K. is 30th November.

Members in Europe whose orders can be received by the printers (Colmore Press) at a date intermediate between this and the 'overseas closing date' of 30th October, should not be deterred from ordering cards, as all practicable priority will be given to them.

## Role of New Technician Education Council

The role of the new Technician Education Council (TEC), which will be principally concerned with setting standards, awarding qualifications and promoting advances in technician education, has been outlined in a preliminary statement.

Set up in March, under the Chairmanship of Mr. A. L. Burton, for the purpose of developing policies of education for men and women at all levels of technician occupations in industry in England, Wales and Northern Ireland, the Council will be engaged in producing a more rational and simplified range of courses. These will lead to TEC qualifications of Certificate, Higher Certificate, Diploma and Higher Diploma, details of which are due to be published in Spring 1974.

Although it will be concerned with arrangements for assessing students, the Council does not see itself primarily as an examining body. It intends that maximum flexibility will be a feature of its schemes covering entry qualifications, methods of assessment and course structures.

TEC also intends to participate in curriculum development and encourage development on the part of others to meet the need for new schemes to be introduced. Colleges, in co-operation with industry, will be encouraged to plan and operate their own programmes within the Council's guidelines, for the Council believes that any form of education is more likely to be successful if the teacher has a sufficient measure of control over what he teaches and is involved in its assessment. There will, however, be a place for external examinations where required.

Copies of the Council's full statement may be obtained from: Francis Hanrott, Chief Officer, Technician Education Council, 76 Portland Place, London W1N 4AA. (Tel: 01-580 3050.)

## Commonwealth Engineers Meet in London

The Commonwealth Engineering Conference, which was attended by 40 delegates from 13 countries, ended its three-day meeting in London on Thursday, 13th September. One of the major issues discussed and agreed in principle was that of a credential card or 'passport' which can be made available for Commonwealth engineers and which would serve as an introduction of them into engineering circles in countries other than their own. It is not the intention that the card would provide for reciprocity or acceptance of qualification standards in the country or countries being visited, nor will it be a licence to practice professionally in them.

The delegates felt that the title Commonwealth Engineering Conference does not now reflect the role and purpose of the organization. At the time of the first meeting in 1946 it was indeed a 'conference' as it was not then known how it would develop. It now meets regularly, has a constitution, and is developing a purposeful role among Commonwealth countries. It was agreed that consideration be given to changing its title from 'Conference' to something more appropriate, and this will be formalized in the near future.

It was reported at the Conference that two highly successful meetings of regional organizations within the CEC had been held in the last year, one in Jamaica and the other in India. Their main concern had been with education and training and of co-operation of members within their respective regions. The main Conference in London endorsed a proposal that there should be further meetings of these two regional groupings. It was also agreed that further groupings might be possible if there was a call for them.

Delegates to the Conference were officially welcomed by Sir Angus Paton, Chairman of CEI and later attended a Government reception at Lancaster House at which their host was Mr. Cranley Onslow, Parliamentary Under-Secretary of State for Aerospace and Shipping, Department of Trade and Industry.

## IEEE Travelling Scholarship

An IEEE travelling scholarship of £300 is offered for visits to foreign electrical or electronic research or manufacturing establishments by a postgraduate student. The purpose is to promote an exchange of research and technological ideas and to foster a closer relationship between young engineers in different countries. Similar scholarships were awarded in 1972 and 1973.

Candidates must submit programmes for their visits by 31st December 1973, and the award will be made to the candidate whose programme is judged most likely to promote the objects of the scholarship.

The scholarship is financed by the UK and Republic of Ireland Section of the Institute of Electrical and Electronics Engineers, which is acting in collaboration with the Institution of Electrical Engineers and the Institution of Electronic and Radio Engineers. Entrants must be Student or Graduate Members of one of these three institutions.

Further information and entry forms may be obtained from: Prof. C. W. Turner, Dept. of Electrical & Electronic Engineering, King's College, Strand, London WC2R 2LS.

## Royal Aeronautical Society Appoints New Secretary

Dr. A. M. Ballantyne, O.B.E., T.D., Ph.D., C.Eng., F.R.Ae.S., has retired after 22 years as Secretary of the Royal Aeronautical Society. He is succeeded by Mr. E. M. J. Schaffter, M.A., C.Eng., A.F.R.Ae.S., who has been with the R.Ae.S. as Deputy Secretary since 1970.

## The 1974 Graham Clark Lecture

Sir Kingsley Dunham, F.R.S., Director of the Institute of Geological Sciences and a Past President of the Institution of Mining and Metallurgy, is to give the next Graham Clark Lecture. The Lecture and CEI Dinner will take place on Wednesday, 9th January, 1974, at the Institution of Civil Engineers.

## Pilot Scheme on Individual Design Consultants

The CEI, in conjunction with the Design Council, announced last April that it was proposing to conduct a pilot scheme on the use of individual design consultants, to see whether there was a need for those experienced in design, especially older or retired engineers, to contribute to the solution of industrial problems and to help raise standards of engineering design.

As a result of the announcement of the scheme in several of the journals of the constituent members of the CEI, a large number of replies was received indicating a great deal of interest in the proposal both from established consultants and from those who wished to participate.

By arrangement with the Design Council, many have been interviewed and their interests and experience catalogued. Details of all applicants have been passed to the Design Council who will include the information in the schedules and lists of expertise which they are setting up. Already some half dozen firms have expressed interest in using individual consultants and arrangements have or are being made for one of those who applied to assist. It seems likely, therefore, that by the time administrative arrangements are complete within the Design Council there will be no difficulty in carrying out the pilot scheme in the use of Design Consultants as the CEI and Design Council originally contemplated.

## Royal Society Mullard Award

The President and Council of the Royal Society have bestowed the Royal Society Mullard Award for 1973 on Professor C. W. Oatley, O.B.E., F.R.S., emeritus professor of electrical engineering in the University of Cambridge, in recognition of his outstanding contribution over a long period to the design and development of the scanning electron microscope which has been sold in substantial numbers to laboratories all over the world during the last eight years and has made a significant contribution, both direct and indirect, to national prosperity. The instrument has been widely used in many branches of science, both physical and biological, and in a large number of industries. It produces images of surface structure (which can only be done with the conventional transmission electron microscope via thin film replicas); it has been applied to the study of natural and synthetic fibres, of semiconductor devices, of magnetic domain structures, of metallurgical specimens; and one of its early uses was in botanical research, for the classifying of pollen grains from their surface morphology. The instrument is highly versatile, because of the many different modes of operation, and the ease with which images can be electronically processed and manipulated.

The Award which was instituted in 1967 will be presented by the President at the Anniversary Meeting of the Royal Society on 30th November 1973. The Award consists of a gold medal and a prize of £1000 provided by a gift being made by the board of directors of Mullard Ltd. It is made annually for an outstanding contribution to the advance of science, engineering or technology that in the preceding ten years has led directly to national prosperity in the United Kingdom; with emphasis on the application of science and technology to the whole range of manufacturing industries, including those based on agricultural and other biological processes of production.

## European Science Foundation

Representatives of academies and research councils from Austria, Belgium, Denmark, France, German Federal Republic, Greece, Norway, Netherlands, Portugal, Spain, Sweden, Switzerland, United Kingdom and Yugoslavia agreed in principle at a meeting on 25th September 1973 at Gif-sur-Yvette, France to set up a European Science Foundation.

The objects of the Foundation would be:

to advance cooperation in fundamental research;

to promote mobility of research workers between participating bodies;

to assist the free flow of ideas and information;

to facilitate the harmonization of the fundamental research activities of member bodies;

make grants for the support of concerted actions and collaborative schemes;

facilitate cooperation in the use of existing facilities;

facilitate cooperation in assessing and executing major projects; and

facilitate cooperation in the provision of expensive specialized services.

The Italian representatives expressed their support in principle for the creation of a European Science Foundation but were not yet able to make a firm decision.

The representatives agreed to set up a Preparatory commission to:

(a) seek offers of a seat for this Foundation, to assess those offers, and to make proposals for the location of the seat of the Foundation to a plenary meeting of representatives of prospective members to be held not later than May 1974 in Stockholm; and

(b) to make proposals to the same meeting for the organization, method of work, staffing and administrative arrangements of this Foundation;

in order to enable the Foundation to be set up as soon as possible thereafter.

Professor H. Curien (France) was appointed Chairman of the Preparatory Commission and the following as the other members:

J. H. Bannier (Netherlands); H. Bloch (Switzerland); A. P. Masia (Spain); F. Schneider (German Federal Republic); H. G. Schuster (EEC); T. Segerstedt (Sweden); R. St. J. Walker (U.K.). (Mr. Walker is Secretary of the Science Research Council).

## CNAA Reports a Year of Expansion

Continued expansion in the non-university sector of further education is reflected in the 1971/72 Annual Report of the Council for National Academic Awards published recently.

In the 1972/73 session, the Report records, 34,189 students were engaged on courses leading to CNAA first degrees—an increase of 5,666 over the previous year. Of these 21,229 were taking science and technology subjects, 12,843 arts and social studies and 117 education courses. The total number of 444 first-degree courses in progress during 1972/73 represents a 22% increase over 1971/72.

Significant trends noted in the Report are that the proportion of sandwich courses has declined from 53% to 49%, while the proportion of part-time courses has increased from 10% to 13% of the total.

The proportion of sandwich course students fell from 60% in 1971/72 to 54% during the period under review, while the proportion of full-time students increased from 34% to 40%. The proportion of part-time students remains unchanged at 6% but in science and technology the number of part-time students has increased by 49%.

The number of part-time degree courses (56) represents an increase of 51% and constitutes 13% of the total number of first degree courses in operation. Notable areas of expansion in this category have been physics and electrical engineering.

In 1971/72 the CNAA approved 173 first-degree courses out of the total of 293 submitted by polytechnics and colleges. A further 44 were still under consideration at the end of September 1972.

There are currently 47 postgraduate courses leading to the award of the Council's Master's Degree, with a total of 583 enrolments. The present session is the first year in which courses leading to the Council's Diploma have been in operation and there are 96 enrolments for seven courses. A total of 1,350 candidates were registered for the CNAA M.Phil. and Ph.D. research degrees at the end of 1972; 440 such awards having been made during the year.

## First General Assembly of the ERB

It is expected that the first General Assembly of the Engineers' Registration Board will be held early next year and that it will be fully representative of more than 40 member bodies of the Engineers' Registration Board in all three sections: Chartered Engineer, Technician Engineer and Technician. This will be a very real milestone in the progress of the Board.

The Engineers' Registration Board is the nationally recognized authority for engineers and technicians across all engineering disciplines and at all levels of qualifications. Since its formation in 1971, the Board has been largely preoccupied with laying the foundations, and dealing with the main problems inherent in providing the structure for a register. It is now clear that there exists an excellent basis for progress in the years ahead toward its main objectives.

The last two years have seen continuous progress in creating the basic structure of the ERB. The 43 independent Institutions which are now members have given a remarkable example of the co-operation and constructive action without which no federal organisation can progress. It is now a stable and influential body which is able to express views based upon common standards and interest. Work on the integration of the existing Chartered Engineers' register, together with its supporting functions, is well under way, and will soon be complete.

The complete Engineers' and Technicians' register is being built up rapidly. Over 150,000 Chartered Engineers are already registered and the registers for Technician Engineers and Technicians are growing fast.

## Conference on Gas Discharges

The third International Conference on 'Gas Discharges' will be held at the Institution of Electrical Engineers, London from 9th–12th September 1974. The 'call for papers' states that the aim of the Conference is to emphasize the engineering applications of gas discharges. The subjects covered will include: arc furnaces; discharges for space vehicle propulsion; display devices; gas-filled valves; gas lasers; high-voltage technology; lamps; switchgear; welding, cutting and machining; and other gas-discharge applications.

In addition, it is intended to cover the fundamental processes connected with gas discharges and these will include: corona; electrode-less and radio frequency discharges; electrode phenomena; glow discharges; high- and low-pressure sparks; laser-produced plasma; lightning; long sparks; plasma diagnostics and measurement techniques; pre-breakdown phenomena; radiation from discharges; transient and steady-state phenomena and high- and low-pressure arcs; and waves in plasma. (Contributions on discharge chemistry, m.h.d. generation and fusion are not invited.)

The conference is being organized by the Science, Education and Management Division of the Institution of Electrical Engineers in association with the IERE, the Institute of Electrical and Electronics Engineers (United Kingdom and Republic of Ireland Section), the Institute of Mathematics and its Applications, the Institute of Physics, and the Welding Institute.

The Organizing Committee invites offers of contributions not exceeding 2,500 words (that is, a maximum of five A4 pages including typescript and illustrations) for consideration for inclusion in the conference programme. Those intending to offer a contribution should submit a synopsis of approximately 250 words to the IEE Conference Department by 19th November 1973. The full typescripts will be required for assessment by 15th March 1974.

Registration forms and further programme details will be available a few months before the date of the conference from the IEE Conference Department, Savoy Place, London WC2R 0BL.

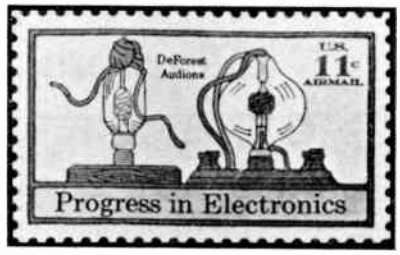
## Trading with Japan

The announcement that UK exports to Japan in the first seven months of 1973 were £152M, an increase of 57·9% over the same period of 1972, lends particular point to a new booklet, 'Japanese Government and Business', which has just been published by the British Overseas Trade Board. The booklet provides British firms entering the expanding Japanese market with a background to the close system of co-operation between the Japanese Government and business community.

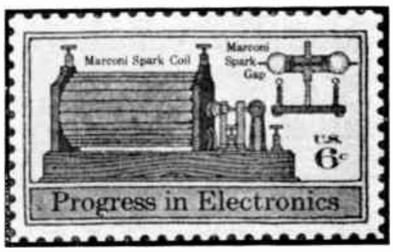
'Japanese Government and Business' shows that the unique and long-established pattern of industry, trading and finance has been little affected by Western business methods. The booklet explains how the system developed and now operates in Japan. The role of the large Japanese trading conglomerates is examined together with company finance and public ownership. Export incentive schemes available to Japanese firms, the Japanese employment system and labour relations are also described.

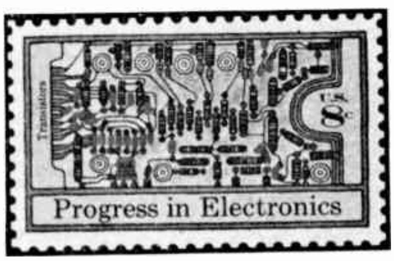
Copies of 'Japanese Government and Business' are obtainable from the British Overseas Trade Board's Exports to Japan Unit, Hillgate House, Old Bailey, London EC4M 7HU. (Telephone: 01-248 5757, extn 7162 or 613).

# Electronics progress shown on stamps








The four stamps were designed by Walter and Naiad Einsel and the devices shown are based mainly on models in the Smithsonian Institution.

Last year the 50th anniversary of the British Broadcasting Company (later Corporation) and the 75th anniversary of Marconi's first successful tests of the transmission of radio signals over water were commemorated by a series of British postage stamps—described and illustrated in the October 1972 issue of *The Radio and Electronic Engineer*. On July 10th last, the US Postal Service issued four stamps—on the 10th anniversary of the launching of the *Telstar* communications satellite—which illustrated some of the most significant electronic devices to be introduced since the start of the century.

On the 6 cents stamp are shown a spark coil and gap of the kind used by Guglielmo Marconi to transmit wireless signals across the Atlantic in 1901. Both devices are typical of the equipment used by experimenters in the 1900–1918 period.

The 8 cents stamp typifies the 'semiconductor revolution' with a representation of a typical early transistor circuit board.

Coincidentally, 1973 is the centenary of the birth of Lee de Forest, inventor of the audion or triode, two versions of which are shown on the 11 cents stamp. The first commercial audions, like the one pictured on the left side of the stamp, were made for de Forest by a Christmas-tree lamp manufacturer—H. W. M. McCandless of New York. These had a flat plate (anode), usually platinum and about 10 by 15 mm in size, placed about 3 mm from a carbon or metal filament in a roughly cylindrical bulb. A simple zig-zag platinum wire grid was fixed between the anode and filament. Both anode and grid were supported on wires sealed through the glass at the top of the bulb, and held in place by pitch or tar, while the filament was connected to the base like any ordinary incandescent lamp.

About 1908, audions were changed to the bulb-shape shown on the stamp at the suggestion of McCandless who felt this shape would permit easier assembly. To increase the life of the devices, an extra filament was added and connected to a wire coming out just above the base. The bulb was used until the first filament burned out. Then the projecting wire was wrapped around the base and contact was maintained by a heavy rubber band supplied by the manufacturer. This brought the second filament into action. Average filament life was 35–100 hours, despite de Forest's advertised claim of 800–1000 hours. A September 1909 advertisement shows audions priced at \$3·50 to \$7·50 and complete de Forest detector apparatus at \$18·00 to \$25·00.

Dr. Lee de Forest, who died just over twelve years ago, was a colourful figure with a great propensity for litigation (usually successful), and one celebrated case concerning the use of the audion as an oscillator was described by Professor D. G. Tucker in a paper in *The Radio and Electronic Engineer* for February 1972.

At the left side of the 15 cents stamp is a carbon microphone of the type used on the NBC radio network in the 1920s. On the right is a horn-type speaker, a design that was in widespread use about 1924. The radio tube is a 1912 design and the television camera tube is a dissector-multiplier type developed by Philo T. Farnsworth in the 30s.

The information on this set of United States stamps has been based on an article in *IEEE Spectrum* for July 1973.

# Members' Appointments

## CORPORATE MEMBERS

**Col. E. H. Davies** (Fellow 1973), for the past three years on the staff of the Signals Officer-in-Chief, Ministry of Defence (Army), has been appointed Marketing Manager, Army Systems, with Sperry Gyroscope, Bracknell.

**Mr. L. A. Smulian** (Fellow 1963, Member 1953) who has been with the Plessey Company since 1966, latterly as Managing Director of the Microsystems Division, has been appointed Head of the Telecommunications Division in the Directorate-General for Industrial and Technological Affairs of the European Economic Commission. Mr. Smulian joined the Institution's Papers Committee at the beginning of this year and a fuller note on his career was published in the Journal for April 1973.

**Sqn. Ldr. T. Briggs, RAF** (Member 1968, Graduate 1962) has been posted to RAF Wattisham as Officer Commanding Electrical Engineering Squadron. He was previously in the Directorate of Electrical Engineering of the Ministry of Defence.

**Mr. W. R. Cameron** (Member 1966) is now Production Manager with Fabri-tek Computer Components, Blantyre. He was previously a Design Engineer with Honeywell Controls Limited, Motherwell.

**Mr. P. A. Clarke, B.Sc.,** (Member 1969) who was a Lecturer in Automatic Data Processing Systems at the Defence ADP Training Centre, Blandford Camp, has moved to the School of Signals, Blandford, to be Senior Lecturer in computer systems.

**Wing Cdr. W. R. F. Cooney, M.B.E., RAF** (Member 1951) has retired from the Royal Air Force. His final appointment was Staff Officer Engineer at the British Embassy in Washington. Wing Cdr. Cooney served for some years on the Education and Training Committee and more recently on the Management Techniques Group Committee.

**Lt. Col. R.A. Garrad, REME,** (Member 1964, Graduate 1960) has taken up the appointment of Assistant Director responsible for the reliability engineering branch of the headquarters of the Director of Electrical and Mechanical Engineering with the Ministry of Defence (Army) in London. He was Officer Commanding the Electronic Engineering Training Wing, REME Electronic School, Arborfield. Lt. Col. Garrad is a member of the Computer Group Committee.

**Flt. Lt. P. A. G. Leach, RAF** (Member 1971), until recently Officer Commanding Air Electrical Engineering Flight, RAF Wattisham, has been posted to HQ Defence Communications Network, RAF Medmenham, as Signals Officer (Engineering Terminal) Systems.

**Mr. D. H. Morley** (Member 1973, Graduate 1970) has been promoted to Professional and Technical Officer 1 in charge of a section dealing with the appraisal and trials of digital scientific systems in the Central Computer Agency of the Department of Trade and Industry. Mr. Morley was previously in the Computer Environmental Section of the Technical Services Division of the Central Computer Agency.

**Mr. R. A. Nye** (Member 1968, Graduate 1965) who was Communications Technical Superintendent, British European Airways, is now Manager Radio Systems and Services, British Airways Group at London Airport.

**Inst. Capt. P. J. Poll, M.Sc., RN** (Member 1961) will shortly take up the appointment of Dean of the Royal Naval Engineering College, Manadon, (HMS *Thunderer*). Captain Poll who was previously Assistant Director of Naval Education Services and Director of Naval Training Support, served as a member of the Council of the Institution from 1969-72 and on the Education and Training Committee and on the Examinations Committee from 1961–66.

**Mr. A. G. Povall** (Member 1963) who was manager of EASAMS Limited at Camberley, is now Trials Manager for the company at its Avionics Test Facility at the British Aircraft Corporation's factory, Warton, Lancashire.

**Mr. K. S. Rajagopalan** (Member 1952) has been appointed Director of the International Airports Authority of India at Madras Airport. Mr. Rajagopalan joined the Civil Aviation Department of the Government of India in 1946 following service with the Indian Air Force.

**Group Captain H. R. Riley** (Member 1973) has been posted to the Ministry of Defence as Deputy Director, Electrical Engineering (Air). He was previously Station Commander at RAF Sealand.

**Mr. P. A. Robertson** (Member 1963, Graduate 1949) has been appointed Director of Engineering of the English Services Division of the Canadian Broadcasting Corporation. Mr. Robertson joined the CBC in 1961; before going to Canada in 1959 he was with Marconi's Wireless Telegraph Company and with Associated Rediffusion Limited.

**Mr. B. G. Smee** (Member 1971, Graduate 1966) is in charge of sales of miniature lamps and various types of gas discharge light sources with Valiant Electrical Company Limited. He was previously for a number of years with Hivac Limited.

**Mr. D. Smith, B.Sc., M.Eng.** (Member 1962) has been appointed head of Department of Electrical Engineering at Doncaster College of Technology.

**Mr. M. H. Taylor, M.Sc.,** (Member 1972) who was until earlier this year Senior Lecturer in Electronics at the RAF College, Cranwell, with the rank of Squadron leader, has taken up an Education Staff (Ground Training) post at RAF Training Command, Brampton.

**Mr. H. J. Terry, B.A., Ph.D.,** (Member 1971, Graduate 1966) who has been a Lecturer in the Department of Medical Electronics at the Medical College of St. Bartholomew, London, has joined the Medical Physics Unit of St. James's Hospital, Leeds, as a Principal Physicist.

**Mr. D. W. Thomasson** (Member 1949) has joined Marconi Space and Defence Systems Limited, Stanmore as a Professional Principal Engineer. Mr. Thomasson, who has contributed several papers to the Institution's Journal, has previously held senior appointments with a number of companies concerned with avionic equipment and with computers.

**Dr. Thong Saw Pak** (Member 1961) has been appointed Principal of the Tunku Abdul Rahman College in Kuala Lumpar. Dr. Thong had been Professor of Physics at the University of Malaya since 1959.

**Mr. R. V. Todhunter,** (Member 1969, Graduate 1966) who has been with Oldham College of Technology since 1967, has been promoted to Lecturer II in the Department of Electronic Engineering.

**Mr. L. J. Townsend,** (Member 1966, Graduate 1963), formerly with Shell U.K. Limited as an Instrument Engineer, has joined Allied Breweries (Production) Limited as Area Engineer – Brewing and Fermentation.

**Mr. P. B. Walker,** (Member 1968, Graduate 1966) is now Assistant Manager, Technical Administration, with Rank Xerox Limited, Mitcheldean, Gloucestershire.

**Mr. P. R. Walwyn, M.A.** (Member 1969) who retired from the Royal Navy last year in the rank of Lieutenant Commander, has recently taken up an appointment in the Avionics System Department of Hawker Siddeley Aviation Ltd., Kingston-upon-Thames.

**Mr. J. E. Wicks** (Member 1971, Graduate 1955) has joined ITT Creed, Brighton, as Mechanical Engineering Manager. He was previously Chief Mechanical Designer with ENM Limited.

**Mr. G. A. Williams** (Member 1971, Associate 1968) has been appointed Technical Manager, Barber Weston Limited, Weston-Super-Mare, Somerset. He was previously with Quick Maid Vending Company Limited as Research and Development Manager.

## NON-CORPORATE MEMBERS

**Mr. R. Chapman** (Graduate 1967) has been appointed Lecturer in Electrical Engineering at Paisley College of Technology.

**Mr. I. F. Gordon** (Graduate 1971) is now in the Systems and Technical Support Department of ICL (Australia) Pty Limited with responsibility for data capture systems and remote terminal systems. He was previously with EMI Electronics (Australia) Pty Limited.

**Mr. Q. Hlaing** (Associate 1970) is now an Exchange Engineer with the Rangoon Telephone System, Posts and Telecommunications Corporation, Burma.

**Mr. P. G. Howard** (Graduate 1968) is now an Electronic Engineer with Hewlett Packard Limited, Colorado Springs. Before going to the United States earlier this year Mr. Howard was with Fielden Research, Llangefni, Anglesey.

**Mr. R. J. Jones** (Graduate 1970) has been appointed a Senior Reliability Engineer with Rank Xerox Limited, Mitcheldean, Gloucestershire.

**Mr. R. D. Lawrance** (Graduate 1964) has been appointed Senior Tutor at the Post Office Telecommunications Management College.

**Mr. C. L. Lawson** (Associate 1972) has joined Philips Electrologia GmbH at Eiserfeld, West Germany as a Computer Development Engineer following 13 years service in the Royal Air Force, latterly as Sergeant Technician concerned with mobile ground radar equipment.

**Mr. S. F. Monk** (Graduate 1971) is now an Electronics Test Engineer with Stibbe Electronics, Leicester. He was previously employed in a similar capacity with Marconi Radar Systems Limited.

**Mr. S. E. Osime** (Graduate 1971) has been transferred from the Nigerian Post Office to the Posts and Telecommunication Training Centre, Lagos, as a Lecturer.

**Mr. F. B. Norman** (Graduate 1969) who was formerly with the Plessey Company at Beeston, Nottingham, has recently joined Stromberg-Carlson Corporation, Rochester, New York, as an Equipment Engineer.

**Mr. J. E. Pearson, M.Sc., Ph.D.** (Graduate 1967), formerly Manager, Research Service, Kings College, London, has joined Roche Products Limited, Welwyn Garden City, as an Electronics Engineer.

**Mr. M. J. Webber** (Graduate 1967) who is with the BBC, has been transferred to the HF Unit, TCPD; he was formerly with the Aerial Unit.

**Mr. R. G. White** (Graduate 1970) who is with the British Gas Corporation is now responsible for Communications in the North Area of Wales Gas Board.

**Mr. W. S. Willis** (Graduate 1968) has taken up the appointment of engineering manager with IDC (Northern) Company Limited, Wrexham. Mr. Willis was previously with the UK Atomic Energy Authority.

# Obituary

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

**Laurence Duncan Line** (Member 1969) died on 12th August at the age of 54 years. He leaves a widow and a son and daughter.

Educated at Ealing Grammar School Mr. Line was commissioned in the Royal Corps of Signals in 1943. He saw war service in the Middle East as a regimental Signals Officer, and later held appointments as a Signalmaster at the War Office, as a Senior Wireless Officer and, from 1949–50, as a Radio Staff Officer, GHQ, Far East. He passed the Technical Staff Officers' course at Shrivenham in 1952 and subsequently served in the Ministry of Supply Inspectorate of Armaments, in command of a Radio Squadron in BAOR, and again in the Far East. He was seconded to the Ministry of Aviation in 1963 and was promoted Lieutenant Colonel in 1964. In 1965 Col. Line became associated with Project Mallard intially as Military Deputy to the Head of the U.K. delegation and in 1966 as Deputy to the U.K. Programme Manager. He retired from the Army in 1968 and was appointed Engineer II in the Ministry of Technology. He was promoted to Principal Professional and Technology Officer in 1971 and a month before his death had been appointed to manage a project on airborne communications installations within the Ministry of Defence Procurement Executive.

**Lord Renwick of Coombe, KBE,** (Fellow 1947) died on 31st August aged 68. He leaves a widow, and a son and three daughters from his first marriage.

Formerly Sir Robert Renwick, Bt., Lord Renwick was created a baron in 1964. He was educated at Eton and Trinity College, Oxford and from 1939 was Chairman of the County of London Electric Supply Company until its nationalization in 1948. From 1942 to 1945 he was Controller of Communications Equipment in the Ministry of Aircraft Production and he played a leading part in the development of radar and, as Chairman of the Airborne Forces Committee, directed the fitting out of airborne expeditions into Europe. He was appointed K.B.E. in 1946.

Lord Renwick was on the Boards of many companies either as Chairman or a Director, including Associated Television, Unidare, British Insulated Callender's Cables, East African Power & Lighting Co., Nigerian Electricity Supply Corporation and the Kenya Power Co. He was also a partner in the stockbroking firm of W. Greenwell & Co., and had been Chairman of the Council of the Institute of Directors and President of the Radar Association.

**See Keng Tan** (Student 1971) died on 26th April last as a result of a road accident in Sarawak. He was 23 years of age.

After primary and secondary education in Kuching Mr. Tan attended Kuala Lumpur Technical College from 1968 to 1971 where he obtained his Diploma in Electrical Engineering. He then returned to Kuching where he was employed as a Broadcasting Technician with Radio Malaysia from July 1971 to the time of his death.

# An Army Communications System for the 1980s

About fifteen years ago the British Army recognized that a new communication system would be required to support a modern army in the field. It was established that speed of response, tactical mobility, survivability, communication availability and an ability to provide the improved service without increasing military manpower were primary features in determining the nature of system design.

Improved service facilities which were determined included:

The ability to provide secure communication in four modes: speech, facsimile, telegraph and data;

Direct dialled communication between subscribers on the battlefield, including Headquarter Staffs and mobile commanders of units engaged in battle;

The provision of a survivable and highly mobile automatic trunk network using stored program controlled switches;

The capability to attach data processing terminals to the communication system which would itself act as a data bearer;

The enhancement of service by providing such features as conference, broadcast, call transfer, call hold and priority service.

Study and project definition proceeded on this concept originally known as *Hobart* and subsequently as *Mallard*. The next stage in the implementation of the project now begins with the engineering development of the *Ptarmigan* system and its equipments.

*Ptarmigan* differs from previous military communication systems in that it provides greater communications capability for the user, places fewer constraints in operational planning, reduces the size of headquarters and improves tactical mobility. It is also less vulnerable to interruption through equipment failure or by enemy action.

The system concept was derived from the outset by considering the service features required by military command and control in the tactical environment. The use of stored program controlled (SPC) digital circuit switches was fundamental to the original plan and these are included in Project *Ptarmigan*.

*Ptarmigan* will provide the military equivalent of an STD type of person-to-person telephone service between units—either fixed or mobile—anywhere within the combat zone. Transmission channels are carried almost entirely by means of u.h.f. and v.h.f. radio links instead of by cables so that the necessary mobility is obtained.

## The Trunk Network

The trunk network is set up by deploying the SPC circuit switches at suitable geographical points in the tactical area over which the military force is deployed. The circuit switches are interconnected by multi-channel radio relay trunk links which carry the information channels in time division multiplex streams. Each information channel may be employed by any user for any of the four modes (speech, telegraph, facsimile or data) provided. Interconnexion of the trunk circuit switches provides a grid network over the tactical area.

Damage to a radio relay, or to a circuit switch, does not break the communication system as was the case in the 'chain of command' systems which *Ptarmigan* replaces. Survivability of the *Ptarmigan* area system is therefore greater than previous systems.

The trunk nodes are not co-located with any command headquarters and any headquarters may obtain service from any trunk node to which it can gain access. In this way the number of communication vehicles within a headquarters will be reduced, thus increasing mobility and survivability.

## The Local Distribution Methods

Subscribers to a military communications system are located in the command headquarters of Army, Corps, Division, or Brigade, in individual units, or are mobile over the battlefield. Service is given to the subscribers within a command headquarters by providing radio relay interconnexion between the headquarters site and any trunk node.

The heart of the SPC switches for *Ptarmigan* will be the Plessey PP250 processor—shown above—developed by Plessey Telecommunications for its SPC programme.

Within the headquarters the channels of the radio relay are terminated on an SPC switch of the same type as that within the trunk node and individual subscribers are connected to it by four-wire loops or loop groups. The heart of the SPC switches will be the Plessey PP250 processor which was developed by Plessey Telecommunication for its SPC programme.

In the case of smaller headquarters or of other groups of co-located subscribers it is feasible to dispense with the local switch and distribute circuits from a multiplexer termination at the radio relay, treating the group as a loop group at the trunk node.

Subscribers who require communication whilst mobile, or are remote from other users, are provided the full service features of *Ptarmigan* by a 'radio telephone' subsystem known as Single Channel Radio Access (SCRA). A feature of this subsystem is that it does not require the user to have any special operator training. This particular subsystem provides a service to the military user which has not been possible in earlier military switched networks.

Any of the trunk switches can support groups of mobile subscribers. A radio central station provides a number of communication channels which are available as common user channels to the group of mobile terminals which are affiliated to it. As in the rest of the *Ptarmigan* system, secure duplex communication is provided to the mobile subscriber. The distribution methods described are shown in the system outline above.

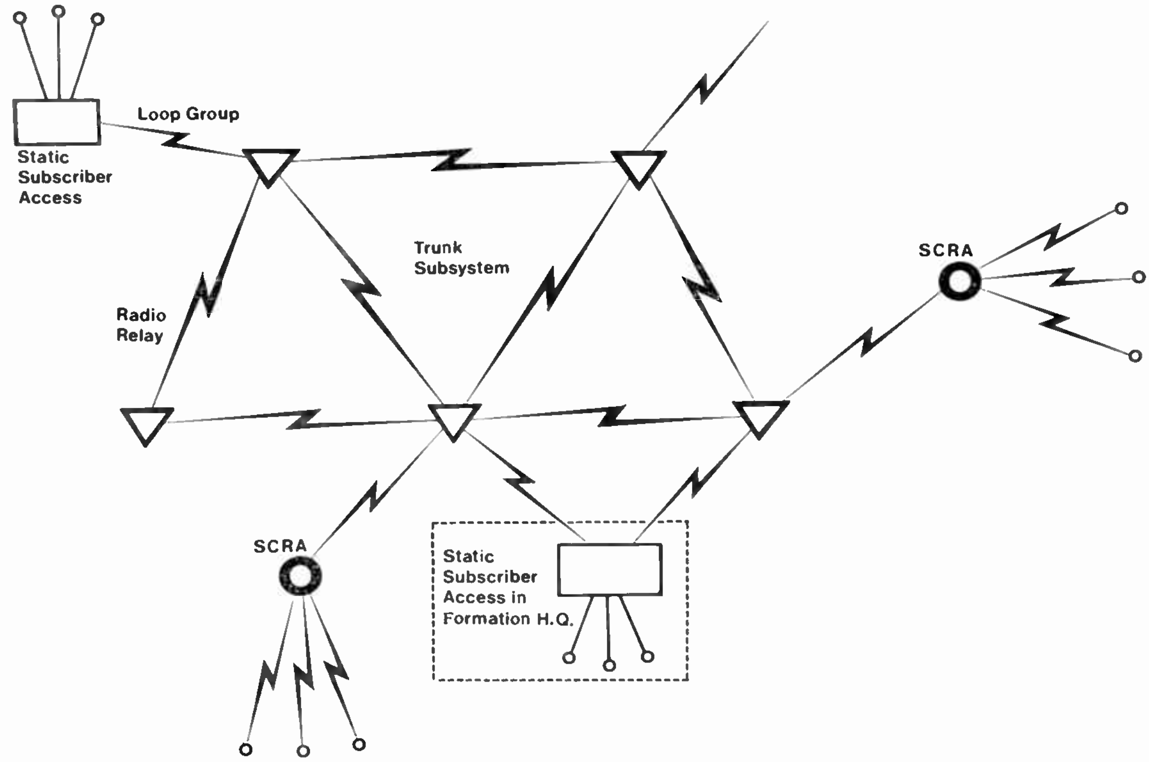
## System Management and other Features

The *Ptarmigan* system is designed to operate and provide service under the hostile conditions of the battlefield. Since the battle is mobile, the communication system must be capable of rapid adjustment to tactical movement whilst still providing the required grade of service to the user. The system requirements in this respect are therefore quite different from those experienced in civil environment. Unique features that are incorporated within *Ptarmigan* include:

- A unique never-changing directory number for each subscriber;
- Automatic alternative routing;
- Broadcast and conference facilities;
- Call hold and call transfer;
- Priority service for nominated subscribers;
- Automatic management aids for system control and planning.

Plessey has been selected as Design Authority and prime contractor for the development and has systems management responsibility for the project. The first phase of development will be carried out over a four-year period at a cost of about £17M. The total programme is expected to lead to production valued in excess of £100M.

The General Electric Company and Standard Telephone & Cables have been associated with Plessey in the definition phase, and, under the new programme, have been designated sub-contractors to Plessey. As prime contractor, Plessey is responsible for project management, system design and evaluation of the total system to prove integrity and performance prior to the commencement of user trials. Plessey will also develop the major switch, system control facilities, the single-channel radio access sub-system and all the system software. Production is expected to commence early in the 1980s.




*Ptarmigan* system outline. The system is overlaid throughout by the system control and support subsystems.

# Forthcoming Institution Meetings

## London Meetings

*Monday, 5th November*

JOINT IEE/IERE COMPUTER GROUP

**Colloquium on DISPLAY TECHNOLOGY**

IEE, Savoy Place, London WC2R 0BL, 10.30 a.m.

Advance Registration necessary. Apply to The Secretary (LS(MA)), Institution of Electrical Engineers, Savoy Place, London WC2R 0BL.

**The electrophoretic image display**

By J. C. Lewis (*Plessey*)

**The d.c. electroluminescent panel as a flat computer display**

By Dr. C. Hilsum (*RRE*)

**Liquid crystal technology and its possible impact on flat computer displays**

By Dr. J. Kirton (*RRE*)

**The a.c. plasma panel**

By J. G. Titchmarsh (*STL*)

**D.c. gas discharge matrix displays**

By F. Walters (*Ferranti*)

**Light emitting diode display techniques**

By Dr. A. R. Peaker (*Ferranti*)

**High brightness—high definition displays in yellow and green light emitting gallium phosphide**

By T. F. Knibb (*Plessey*)

**Cathodochromic cathode-ray tubes**

By Dr. H. C. A. Hankins (*UMIST*)

*Wednesday, 14th November*

COMPONENTS AND CIRCUITS GROUP

**Colloquium on DOMESTIC EQUIPMENT CONTROL SYSTEMS**

IERE Lecture Room, 10.30 a.m. *Please note change of time.*

Further details and registration forms from Meetings Secretary, IERE.

**Washing machines—problems and possibilities**

By T. Jacobs (*Mullard*)

**Developments in domestic controls with particular reference to electric cookers**

By D. V. Martin (*Diamond H Controls*)

**Development of controls for electric heating systems**

By A. M. Umpelby (*London Electricity Board*)

**Domestic dimming circuits**

By A. Isaacs (*Thorn Lighting*)

**Speed controls—motor controls: review of past, present and future**

By A. Collie (*Kenwood*)

**International standards applicable to electronic controls for domestic electrical equipment**

By C. Lee (*Texas Instruments*)

**Generation and propagation of r.f. noise**

By G. A. Jackson (*Electrical Research Association*)

**Interference suppression in the home**

By A. Tennant (*Plessey Capacitors*)

*Tuesday, 20th November*

AEROSPACE, MARITIME AND MILITARY GROUP MEETING

**Developments in Position Measurement Techniques**

By D. J. Phipps (*Decca Survey*)

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

During the past decade, we have seen a large expansion in the use of electronics for position measurement on the surface of the earth. This presentation deals with one widely accepted method of operation, namely, time shared, differential phase measurement of radio frequencies. In particular the design philosophy, problems and engineering details that led to the development of the recently announced HIFIX/6 system.

*Wednesday, 21st November*

**Dinner of Council and Committees**
Savoy Hotel, London.

*Wednesday, 28th November*

AUTOMATION AND CONTROL SYSTEMS GROUP

**Design and Application of Active Compensation Circuits for Servo Control Systems**

By Dr. D. R. Wilson, D. R. Corrall and Dr. D. F. Neale (*Polytechnic of Central London*)

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

The paper utilizes the phase variable form of the state equation describing a given linear network. The networks that are considered are those most commonly used, namely, phase advance/phase retard and notch networks, to stabilize a servo system. Practical results are given which illustrate the simplicity of tuning the characteristics and which demonstrate that implementation of these filters is a significant improvement in the 'state of the art'.

*Wednesday, 5th December*

AEROSPACE, MARITIME AND MILITARY SYSTEMS GROUP

**Use of Split PPI Techniques in Clutter and other investigations**

By P. D. L. Williams (*Decca Radar*)
IERE Lecture Room, 6 p.m. (Tea 5.30 p.m.)

The accurate comparison between two radars having only one or two chosen differences in parameters is often upset by having to view two independent displays driven from independents scanners. Croney and White have demonstrated split image displays, but the two pictures are separated in time by half an aerial rotation period.

A selection of p.p.i. photographs are to be presented delayed only by a progressive 1 millisecond delay. The chosen parameters investigated are wavelength, horizontal aerial aperture and receiver type (log or linear).

The coastal siting of the equipment enabled land and sea targets to be examined in the presence of varying amounts of sea and rain clutter as well as thermal noise.

*Thursday, 6th December*

EDUCATION AND TRAINING GROUP

**TEC, ERB and the Technician Engineer**

By A. J. Kenward (*SERT*)

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

The identification, definition, education and training of Technician Engineers and Technicians have been important topics for debate over the last few years. The success of the work of the Engineers' Registration Board in forming the composite register and producing and obtaining acceptance of common standards for registration among forty autonomous engineering bodies has helped to clarify ideas on identification and definition.

The education, and to some extent the training, of these two categories of Technicians and Technician Engineers is now under consideration in depth by the Technician Education Council. The first consultative report of the Council will be published immediately before this meeting and will provide an opportunity for discussion of the aims and objectives and of proposals for technician education.

*Wednesday, 12th December*

COMPONENTS AND CIRCUITS GROUP COLLOQUIUM

**IMPACT OF MICOELECTRONICS ON INSTRUMENT DESIGN**

IERE Lecture Room, 2.30 p.m.
Further details and registration forms from Meetings Secretary, IERE.

## Kent Section

*Thursday, 1st November*

**Electronics in the Commercial Vehicle Industry**

By G. H. Leonard (*CAV*)

Medway and Maidstone College of Technology, Chatham, 7 p.m.

The electrical and mechanical environment encountered on modern commercial vehicles presents the equipment designer with some unusual problems. The ways in which these are overcome will be discussed and the mode of operation of some examples of commercial vehicle electronic equipment, such as automatic transmission controls for public service vehicles will be described.

*Wednesday, 5th December*

**Electronic Systems for the Space Environment**

By A. J. Price (*Marconi Space and Defence Systems*)
Medway and Maidstone College of Technology, Chatham, Kent, 7 p.m.

The paper considers the influence of the space environment on the design of electronic circuits. Specific systems are described and their operation in weather satellites and sounding rockets.

## East Anglian Section

*Wednesday, 7th November*

ANNUAL GENERAL MEETING at 6 p.m.

Followed by a lecture and film on

**The RAF 'Red Arrows' Aerobatic Team** (at 6.30 p.m.)

Havering Technical College, Ardleigh Green Road, Hornchurch.

## Thames Valley Section

*Thursday, 8th November*

**Stereophonic and Ambisonic Reproduction of Sound**

By Professor P. B. Fellgett (*University of Reading*)

The J. J. Thomson Physical Laboratory, University of Reading, Whiteknights Park, Reading, 7.30 p.m.

Monophonic reproduction gave no directional information. Stereophonic reproduction adds this information, but in a way having artificiality and limitations which will be discussed. Ambisonic reproduction is defined as capable of surrounding the listener with sound and providing ambience information. After comment on the four-channel fallacy, an outline will be given of the ways in which ambisonic reproduction can be achieved using a minimum of two channels. Subjective and aesthetic aims will be discussed, and speculations presented based on experience of the way in which live and reproduced sound is localized. Reasons will be given for doubting if commercially proposed so-called 'matrix' systems can be adequate even for 'pop' purposes.

*Thursday, 29th November*

**Digital Filters**

By A. R. Owen (*University College of North Wales*)

The J. J. Thomson Physical Laboratory, University of Reading, Whiteknights Park, Reading, 7.30 p.m.

## Southern Section

*Wednesday, 7th November*

**Solid State Microwave Sources**

By J. J. Finlay (*Plessey*)

University of Surrey, Guildford, 6.30 p.m.

This paper relates to the design of microwave oscillators using IMPATT and TRAPATT devices. A discussion of circuit aspects will demonstrate how reliable sources can be obtained. A design of c.w. IMPATT oscillators with output powers up to 1 W and pulsed TRAPATT oscillators with peak powers up to 100 W will be discussed. Typical applications will indicate how these devices are gaining acceptance in practical systems.

*Friday, 9th November*

**Colour Television**

By A. C. Maine (*I.O.W. Technical College*)
Isle of Wight Technical College, 7 p.m.

This lecture will review the progress of colour television from the early Baird experiments to the present day quality transmissions. After an outline of the principles involved, the advances made in receiver design will be examined, with particular reference to the various types of display tube and electronic devices available. The lecture will be illustrated by slides and demonstrations.

*Wednesday, 14th November*

**What's New in Multilayer Printed Wiring Board Manufacture**

By G. C. Wilson (*Ferranti*)

Portsmouth Polytechnic, 6.30 p.m.

The speaker will discuss problems which arise in the manufacture of multilayer printed wiring boards, and methods of overcoming them by the use of buried through-plated holes. Laminate shrinkage, resin viscosity testing, drilling and chemical etching of interconnecting holes together with the environmental testing of the finished product, will be discussed.

*Tuesday, 20th November*

**Solid State Microwave Sources**

By H. J. Finlay (*Plessey*)

Bournemouth Technical College, 7 p.m.

See under Wednesday, 7th November.

*Wednesday, 28th November*

**Cash, Credit and Electronics**

By B. W. Parker and M. J. Davies

Portsmouth Polytechnic, 6.30 p.m.

*Tuesday, 4th December*

**Future Telecommunications Projects in Space**

By W. M. Lovell (*Marconi Space and Defence Systems*)
Brighton Technical College, 6.30 p.m. (Refreshments available from 5.45 p.m. in Refectory)

*Wednesday, 5th December*

**Inertial Navigation**

By G. U. Rands (*Marconi-Elliott Avionic Systems*)

H.M.S. *Daedalus*, 6.30 p.m.

The ability to determine the present position of an aircraft to an accuracy of 1 mile or better without using any ground aids or airborne radar is provided by inertial navigation techniques. The lecture deals with principles, techniques and hardware, and briefly traces the history of inertial navigation, describes present day systems and takes a look into the future.

*Wednesday, 12th December*

**Stored Program Control of Telephone Exchanges**

By B. L. Nuttal (*Post Office*)

University of Southampton, 6.30 p.m. (Tea served in Senior Common Room from 5.45 p.m.)

Within the last few decades we have seen enormous advances in the technologies available for logic and circuit design. In particular the advent of transistors and integrated circuits has enabled computers to be manufactured that are larger, faster and more powerful than could ever have been conceived 30 years ago. These computers have had many applications within industry but it is only comparatively recently that the use of a computer in the control of a telephone exchange has been considered. Nowadays, many telephone administrations are turning to stored program control of exchanges as a possible alternative to existing equipment. However, the introduction of stored program control poses probably as many new problems as it solves old ones so is this faith in computers for telephone purposes justified?

## Yorkshire Section

*Wednesday, 28th November*

JOINT MEETING WITH IEE

**World Wide Communication**

By R. T. Mayne

University of Sheffield, 6.30 p.m. (Tea 6 p.m.)

*Tuesday, 4th December*

JOINT MEETING WITH IEE

**Faraday Lecture**

By Dr. A. J. Churchman
City Hall, Sheffield, 7 p.m.

## East Midland Section

*Tuesday, 13th November*

JOINT MEETING WITH IEE

**Fourier Analysis of Video Telephone Systems**

By Dr. D. E. Pearson (*Essex University*)

Edward Herbert Building, Loughborough University of Technology, 7 p.m. (Tea 6.30 p.m.)

*Tuesday, 4th December*

**The Impact of Advances in Electronics in Electrical Heating Processes**

By J. E. Harry (*Loughborough University*)

Edward Herbert Building, Loughborough University of Technology, 7 p.m. (Tea 6.30 p.m.)

Recent advances in electronics are finding extensive applications in electroheat. Examples of their application are for temperature and power control and in the generation of medium and high frequency power. Some of these developments will be described together with their applications.

## West Midland Section

*Wednesday, 21st November*

**Pin-Wheels to Pulses: Electronics—Servant of Postal Sorting**

By S. W. Godfrey (*Midland Postal Region*) City of Birmingham Polytechnic, Franchise Street, Perry Barr, 7 p.m.

The British Post Office has led the world in the development of postal mechanization. The lecture will describe how the postal service has used electronics to aid letter and parcel sorting. A description of the machinery and systems now in use will be given.

## South Midland Section

*Thursday, 18th October*

**Provision of Communications for Remote Clustered Visual Display Units**

By F. B. Sanders (*West Midlands Gas Board*)

Gloucester College of Technology, 7.30 p.m.

West Midlands Gas today makes extensive use of visual display units (v.d.u.) and typical applications include information retrieval and data input to assist in the administration of accounts and service work in an area containing $1\frac{1}{4}$ million gas consumers (domestic and commercial). The v.d.u.s are connected into the central computer hardware via a communications network shared by other systems. The paper includes a description of the computer and communications system and details the evolution of v.d.u. systems from the design to post-design stages. Storage media, c.p.u. message transmission and multiplexing, microwave network, u.h.f. scanning techniques, P.O. lines, modems, and polling techniques are discussed for the hardware aspects while network planning, implementation and support including project co-ordination, system acceptance, commissioning and maintenance are also covered. Mention will also be made of future developments.

*Tuesday, 20th November*

JOINT MEETING WITH IEE

**R & D on Passive Components in the UK**

By Professor D. S. Campbell (*Loughborough University*)

G.C.H.Q. Oakley, Cheltenham, 7.30 p.m.

The paper aims to give a brief survey of present R & D activity in the field of passive components. In order, however, to limit the field, the latest statistics on the electronics industry are discussed and the most important components are identified. Some of these are then discussed in terms of present R & D activity. The importance of the development in the related field of thick film hybrid circuit technology is also examined.

## South Western Section

*Wednesday, 14th November*

**The Planning of Maplin Airport**

By D. W. Turner (*British Airports Authority*)

No. 1 Lecture Theatre, School of Chemistry, University of Bristol, 7 p.m. (Tea 6.45 p.m.)

*Thursday, 15th November*

JOINT MEETING WITH IEE

**Development in Digital Transmission Systems**

By G. H. Bennett

Main Hall, Plymouth Polytechnic, 7 p.m. (Tea 6.45 p.m.)

*Wednesday, 28th November*

JOINT MEETING WITH IEE

**Video Recording**

By J. Jeffrey (*Bell and Howell*)

Queen's Building, University of Bristol, 6 p.m. (Tea 5.45 p.m.)

*Tuesday, 4th December*

JOINT MEETING WITH IEE

**Space Technology and the Future**

By G. K. C. Pardoe (*General Technology Systems Ltd*)

The College, Swindon, 6.15 p.m. (Tea 5.30 p.m.)

The lecture will establish the current situation in space technology and examine how this will evolve in both short term and long term (to year 2000) periods. Technical and cost benefit aspects will be discussed, so too will be the fundamental problem of how to organize space projects. Mr. Pardoe is well known in this field and in addition to his present position of Managing Director of General Technology Systems Ltd., is Vice-President of Eurospace, Paris, and British Director of Eurosat SA, Geneva.

*Wednesday, 5th December*

**Liquid Crystals**

By G. Elliott (*Marconi*)

No. 4 Lecture Theatre, School of Chemistry, University of Bristol, 7 p.m. (Tea 6.45 p.m.)

The speaker will discuss the following: the nature of liquid crystals and their special and optical properties; the behaviour of thin layers and their response to electrical fields; the construction of an electro-optic device and examples of electro-optic effects; and their development and applications.

## North Eastern Section

*Wednesday, 14th November*

**Codes and Coding**

By J. T. Kennair (*University of Newcastle upon Tyne*)

Main Lecture Theatre, Ellison Building, Newcastle upon Tyne Polytechnic, Ellison Place, 6 p.m. (Refreshments in Staff Refectory from 5.30 p.m.

*Wednesday, 12th December*

**Computer Controlled Telephone Exchanges**

By Dr. M. T. Hills (*University of Essex*)

Main Lecture Theatre, Ellison Building, Newcastle upon Tyne Polytechnic, Ellison Place, 6 p.m. (Refreshments in Staff Refectory from 5 30 p.m.)

## North Western Section

*Thursday, 8th November*

**Automatic Electrical Inspection**

By W. J. Stickland (*Marconi Instruments*)

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

Cost reduction in functional testing can be achieved by the use of automatic testing techniques. To optimize these reductions it is important that a very rigorous form of inspection is adopted in order to reduce the product fault rate at the test stage. The various types of faults and methods of location, by using an automatic electrical inspection technique, will be discussed.

*Thursday, 13th December*

**The Application of Electronics in Telephone Exchange Switching**

By F. W. Croft (*Post Office Telecommunications*)

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

The lecturer will give an outline of the Post Office electronic telephone exchange widely used in public service, and touch briefly on a new system for large electronic exchanges. Also covered will be electronic equipment systems used to steer calls over the electro-mechanically switched network, including reference to Stored Program Control processors.

## Merseyside Section

*Wednesday, 14th November*

**The Role of Electronics in the Movement of Shipping**

By K. D. Jones, Extra Master Mariner

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

The paper will examine the two fields of ship movement, that in the open sea and that within the port confines, to describe how traditional methods of making a safe passage have been changed by modern marine developments and electronic equipment. By the successful application of electronic solutions to old questions, many changes have simplified the mariner's problem, but others have increased the dangers through presenting incomplete information.

The future safety and economic operation of ships will lean heavily on the electronics industry. The introduction of on-board computers and their special sensors is an active field of study. The paper will look to the future to suggest the outline requirements on an automatic ship working into a computer controlled port.

*Wednesday, 12th December*

**R.F. Sputtering of Thin Films**

By E. F. Lever (*Liverpool Polytechnic*)

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

The introduction will outline some applications of dielectric thin films, and will describe the problems of their fabrication. R.f. sputtering will be introduced as a solution to this problem, and the general idea of the process will be described.

The characteristics of the r.f. glow discharge will be described, with emphasis on ion sheath formation and the effects of magnetic fields. Practical sputtering systems will then be considered, with detailed attention to the target system, the substrate, the vacuum plant, and the problem of impedance matching. R.f. power requirements will be discussed, with mention of the relative merits of free-running and fixed frequency operation, and the problems of power losses. Finally, the performance of practical systems will be summarized.

## Northern Ireland Section

*Tuesday, 13th November*

JOINT MEETING WITH IEE

**Forum on Designing for Reliability**

Several Speakers

The Main Theatre, Ashby Institute, Queen's University, Stranmillis Road, Belfast, 6.30 p.m.

*Saturday, 8th December*

**Dinner/Dance**

Culloden Hotel, 8 p.m.

*Wednesday, 12th December*

**Wine and Cheese Party**

90 Belmont Road, Belfast

## South Wales Section

*Wednesday, 14th November*

**Solid State Microwave Power Amplifiers**

By G. B. Morgan (*UWIST*)

Department of Applied Physics, UWIST, Cardiff, 6.30 p.m. (Tea in College Refectory from 5.30–6 p.m.)

A brief review of power saturation mechanisms in various types of microwave solid state amplifiers will be given and the competition offered to valves discussed. The power-impedance product limitations of the more important power amplifiers will be considered.

*Wednesday, 12th December*

JOINT MEETING WITH IEE

**Developments in Data Communications**

By M. B. Williams (*PO Telecommunications HQ*)

Department of Applied Physics, UWIST, Cardiff, 6.30 p.m.

(Tea in College Refectory from 5.30–6 p.m.)

Most data terminals in the forseeable future are expected to operate at speeds within the capacity of telephone circuits. Development will continue therefore, of modems and other techniques of adapting the telephone network to handle data. Many countries and organizations are studying the prospects for providing new forms of data communication services. These include more versatile switching systems and the evolution of digital networks capable of carrying a range of new services.

## Scottish Section

*Tuesday, 6th November*

JOINT MEETING WITH IEE

Robert Gordon's Institute of Technology, St. Andrew's Street, Aberdeen, 7 p.m.

*Wednesday, 7th November*

JOINT MEETING WITH IEE

Napier College of Science and Technology, Colinton Road, Edinburgh, 7 p.m.

*Thursday, 8th November*

JOINT MEETING WITH IEE

Glasgow College of Technology, Hanover Street, Glasgow, 7 p.m.

**Medical and Industrial Electronics—from text book to shop floor**

By J. G. Mitchell (*University of Aberdeen*)

*Monday, 10th December*

JOINT MEETING WITH IEE

Room 406, James Weir Buildings, University of Strathclyde, Glasgow, 6 p.m.

*Tuesday, 11th December*

JOINT MEETING WITH IEE

South of Scotland Electricity Board Showrooms, 130 George Street, Edinburgh, 6 p.m.

**Electronic Aids for Medical and Biological Studies**

By Dr. E. T. Powner (*UMIST*)

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### Membership Approval List No. 165 (*cont.*)

OVERSEAS

CORPORATE MEMBERS

**Transfer from Member to Fellow**

SCHER, Joseph. *Ramat-Gan, Israel.*

**Transfer from Graduate to Member**

ANSDELL, Arthur Anthony. *Burlington, Ontario, Canada.*
RAZ, Zeev. *Kiryat Ono, Israel.*
THOMAS, Alwyn, M.Sc. *Tehran, Iran.*
WILLIAMS, Roy Franklin. *B.F.P.O. 161.*
WILLOUGHBY, Anthony Roy. *Lutherville, Maryland, U.S.A.*

**Direct Election to Member**

CHAU, Lam Ko, B.E., M.E. *Ottawa, Canada.*
DEB GUPTA, Ranjit, B.Sc., (Eng.). *Berlin.*
MITCHELL, James. *Regina, Saskatchewan, Canada.*

NON-CORPORATE MEMBERS

**Transfer from Student to Graduate**

GUPTA, Ramesh Chand. *New Delhi 1, India.*
PONNLE, Ayantunde. *Lagos, Nigeria.*

**Direct Election to Graduate**

BARNETT, Francis Paul, B.Sc., Captain, REME. *B.F.P.O. 22.*
CHARALAMBIDES, Christos. *Nicosia, Cyprus.*
LOWING, John Edward. *Glenbrook, N.S.W., Australia.*

**Direct Election to Associate**

DENYER, Maurice Richard. *Plaisance, Mauritius.*

**STUDENTS REGISTERED**

MHATRE, Prakash Kashinath. *Camp Pooni, Maharashtra, India.*
OGUNTADE, Olaniran Adegboye Andrew. *Ibadan, Nigeria.*
QUEK, Chin Swee. *Singapore 8.*
SNG, Yeow Koon. *Singapore 14.*
TAN, Bin Chye. *Singapore 12.*
TAN, Teck Chye. *Singapore 19.*
TAY, Siak Chwee. *Singapore 8.*

**Notice is hereby given that the elections and transfers shown on List 161 have now been confirmed by the Council.**

# INSTITUTION OF ELECTRONIC AND RADIO ENGINEERS

## Applicants for Election and Transfer

THE MEMBERSHIP COMMITTEE at its meetings on 4th and 18th September 1973 recommended to the Council the election and transfer of 142 candidates to Corporate Membership of the Institution and the election and transfer of 24 candidates to Graduateship and Associateship. In accordance with Bye-law 21, the Council has directed that the names of the following candidates shall be published under the grade of membership to which election or transfer is proposed by the Council. Any communications from Corporate Members concerning these proposed elections must be addressed by letter to the Secretary within twenty-eight days after the publication of these details.

### *Meeting: 4th September 1973 (Membership Approval List No. 164)*

#### GREAT BRITAIN AND IRELAND

CORPORATE MEMBERS

Transfer from Member to Fellow

HACKING, John Bellamy. *Teddington, Middlesex.*
MACKELLAR, John Campbell. *Copthorne, Sussex.*

Direct Election to Fellow

COLLINS, Jeffrey Hamilton, Professor, B.Sc., M.Sc., *Balerno, Midlothian.*

Transfer from Graduate to Member

ABBOTT, David Michael. *Ickenham, Middlesex.*
ALTHORP, Derek Peter. *Gosport, Hampshire*
ANDREWS, Ian. *Thornton Heath, Surrey.*
APPS, Terence Alfred. *Chatham, Kent.*
BALDWIN, Richard John, B.Sc. *Cinderford, Gloucestershire.*
BALLANCE, George Henry. *Wilmington, Kent.*
BEST, Dereck Roy. *Crewe, Cheshire.*
BROMLEY, Fred Arthur. *Sandiacre, Notts.*
DEACON, Peter Richard. *Yateley, Surrey.*
FACEY, Edward William. *Fareham, Hampshire.*
FARROW, Ronald. *Reading, Berkshire.*
FAULKNER, David Hamilton. *Whitstable, Kent.*
FOORD, Alexander. *Malvern, Worcs.*
GODBOLD, Alec Frank. *Chelmsford, Essex.*
MIZINIAK, Julian Stanley. *London, SE15.*
OATS, Roger, B.Sc. *Newbury, Berkshire.*
OGILBY, David Anthony. *Markfield, Leicester.*
OWEN, Gwilym Morris. *Penicuik, Midlothian.*
PREES, Michael Ronald. *Milton, Cambridge.*
PREST, Peter Harry. *London, W.4.*
RING, Terence. *Lowton, Lancashire.*
ROBERTS, John William. *Leicester.*
ROBERTSON, Archibald Alistair. *Falkirk, Stirlingshire.*
ROBINSON, Barry Donald. *Southampton, Hampshire.*
SHIEL, Christopher Bennett. *Hartford, Cheshire.*
SMITH, Colin Charles. *Capel St. Mary, Suffolk.*
SMITH, Donald. *Killearn, Glasgow.*
SMITH, Peter John. *Worsley, Manchester.*
SPALDING, James Melville Ferguson. *London, S.W.14.*
STOKOE, Cecil Walter. *St. Albans, Hertfordshire.*
STOREY, Michael Henry. *Harrow, Middlesex.*
SUDUL, Keith Murray. *Bedfont, Middlesex.*
SULWAY, Anthony Brian. *Westfield, Hastings, Sussex.*
SYLVESTER, Brian Edward. *Dartford, Kent.*
TAYLOR, John Henry. *Axford, Hampshire.*
TAYLOR, Raymond John. *Wallington, Surrey.*
TEMPEST, Gary Neal. *Cold Norton, Essex.*
TOMS, Keith Ronald. *Epsom Downs, Surrey.*
UNSWORTH, Charles Frederick. *Nether Stowey, Somerset.*
WARD, David Laing, M.A., Ph.D. *Frodsham, Cheshire.*
WARRINGTON, Roger Anthony. *Hemel Hempstead, Hertfordshire.*
WATERIDGE, Graham Edward. *Dibden Purlieu, Hampshire.*
WILSON, Richard Emile. *Reading, Berkshire.*

Transfer from Associate to Member

BURTON, Thomas Jeffrey. *Tunbridge Wells, Kent.*

Direct Election to Member

ABBOTT, Sandra Margaret, (Mrs.). *Croydon, Surrey.*
BURT, Michael, B.Sc., (Eng.). *Edinburgh.*
CROWLEY, Timothy Patrick. *Dun Laoghaire, Eire.*
DUTTON, Ralph. *Kenilworth, Warwickshire.*
MITSON, Ronald Bert. *Lowestoft, Suffolk.*
STONES, Peter John, Lieutenant Colonel, R.A. *West Byfleet, Surrey.*
WINDLE, Kevin Joseph. *Dublin 14.*

NON CORPORATE MEMBERS

Direct Election to Companion

ZIANI de FERRANTI, Sebastian Basil Joseph. *Hollinwood, Lancashire.*

Transfer from Student to Graduate

KERR, Brian James Dorman, B.Sc. *London, N.4.*
PICKEN, William David, B.Sc. *Johnstone, Renfrewshire.*

Direct Election to Graduate

McCORMICK, Cyril. *Flint, Flintshire.*

Direct Election to Associate

CARR, Christopher. *Northallerton, Yorkshire.*
HALLIDAY, David Clifford. *Little Rissington, Cheltenham, Glos.*
PACKHAM, John Harold. *Brede, Rye, Sussex.*
ROBERTSON, Brian Andrew. *Oakham, Rutland.*

#### OVERSEAS

CORPORATE MEMBERS

Transfer from Member to Fellow

SIPAHIMALANI, Bhagwan Kishinchand. *Hong Kong.*

Transfer from Graduate to Member

BHATT, S. M. Venkatarama, M.Sc. *Chembur, Bombay, India.*
BILIMORIA, Rusy Piroshaw. *Tehran, Iran.*
NARAYANA, Ghattu. *Thana, Bombay, India.*
NORTHROP, Donald. *Munich, West Germany.*
SERGHIOU, Demetriou Costas. *Nicosia, Cyprus.*
TAN, Chew Kim. *Singapore 2.*
TAY, Kai Miang. *Singapore 12.*

Transfer from Student to Member

KINI, P. Ramaraya, M.E. *Bangalore 1A, India.*

Direct Election to Member

AZZOPARDI CAFFARI, Ronald. *Sliema, Malta.*
FADAMI, Ahmad. *Tehran, Iran.*
SCIBERRAS, Victor. *Birkirkara, Malta.*

NON-CORPORATE MEMBERS

Transfer from Student to Associate

FRAZER, Samuel Babashola. *Freetown, Sierra Leone.*

Direct Election to Associate

IRIMISOSE, Peter Oberuomo. *Kaduna, Nigeria.*
MOORE, John Herbert, Warrant Officer, REME. *B.F.P.O.41.*
POON, Sui Kee. *Singapore.*
POTTER, Michael John. *R.A.F. Episkopi, B.F.P.O. 53.*

### *Meeting: 18th September 1973 (Membership Approval List No. 165)*

#### GREAT BRITAIN AND IRELAND

CORPORATE MEMBERS

Transfer from Graduate to Member

ANDERSON, Michael Christopher. *London, E.18.*
ANDREW, Peter Thomas. *Penketh, Lancs.*
BAGWADIA, Feroze. *Chandlers Ford, Hampshire.*
BOWLES, Michael. *Hinckley, Leics.*
BRIDGES, Raymond Russell. *Harrow, Middlesex.*
BROOMFIELD, Christopher Frank. *Hildenborough, Kent.*
BROWN, Michael Anthony. *Taplow, Bucks.*
BUFFIN, John David. *Hove, Sussex.*
BURKE, Lawrence Vincent. *Holywood, County Down.*
CARROLL, Anthony Frederick. *Hindley Green, Lancs.*
CHALTON, Gordon Alfred. *Eastham, Cheshire.*
CROOK, Derek Frederick. *Maidenhead, Berkshire.*
DANN, Robert James. *Gillingham, Kent.*
DIXON, John. *Ryton, County Durham.*
DYER, David John. *Blunham, Bedford.*
EVANS, Peter John. *Chandlers Ford, Hampshire.*
FOX, Stuart Andrew. *Bishops Stortford, Hertfordshire.*
FRAMPTON, John Michael. *Wimborne, Dorset.*
FRENCH, Desmond Roy. *Alexandria, Dunbartonshire.*
GILCHRIST, Thomas Harold. *Westbury-on-Trym, Bristol.*
GILL, Michael Arthur Ford. *Kings Lynn, Norfolk.*
HOGG, Peter Michael. *London, E11.*
HOOD, Christopher Robin. *Oulton Broad, Suffolk.*
HOOKER, Denis Stanley. *East Farleigh, Kent.*
HUNTINGTON, Edwin Arthur. *Attenborough, Notts.*
HUTTON, George Neil. *Bourne End, Buckinghamshire.*
LEACH, Andrew John, Flight Lieutenant. *Aberdeen, Scotland.*
McCABE, Derek William. *Newcastle upon Tyne 3.*
MARRIAGE, David John. *Fareham, Hampshire.*
PEARCE, Leonard. *Harold Wood, Essex.*
PETTY, Colin John. *Stratford-on-Avon, Warwickshire.*
POND, Peter William. *Eastleigh, Hampshire.*
PRESKETT, Ian William. *Ashford, Middlesex.*
RAMSAY, Frank. *Edinburgh.*
REPTON, Charles Samuel, B.Sc., Ph.D. *Wooburn Green, Buckinghamshire.*
ROBINSON, Leonard Stanley Frank. *Rayleigh, Essex.*
ROGERS, Michael James. *Peterborough.*
RUDLING, Kevan. *Coventry, Warwickshire.*
RYCROFT, David Haigh, Flight Lieutenant. *Woodhall Spa, Lincolnshire.*
SHINWELL, Alan. *Putnoe, Bedfordshire.*
SMITH, Donald. *London, N.13.*
SMITH, Graham Arthur. *Ilford, Essex.*
STEVENS, David Charles. *Bagshot, Surrey.*
STREETER, Raymond John. *Crawley, Sussex.*
STRINGER, David John. *Stapleford, Notts.*
THATCHER, Jeffery Albert. *Ilford, Essex.*
TINKLER, Alan. *Middlesbrough, Teesside.*
VENISON, John Frederick. *Chatham, Kent.*
WALKER, Robert Anthony. *Selsdon, Surrey.*
WALKER, William. *Marple, Cheshire.*
WALLACE, James Raymond. *Winnersh, Berkshire.*
WALMSLEY, Kenneth. *Brixham, Devon.*
WEEDON, Peter Charles. *Romford, Essex.*
WILKINS, Alfred George. *Woking, Surrey.*
WILLIAMS, Melvyn John, Flight Lieutenant. *Lyneham, Wiltshire.*
WILLIAMSON, William David John. *Portsmouth, Hampshire.*
WINSPUR, Colin Charles. *Birmingham.*
WOODWARD, Arthur Frederick, Lieutenant R.N. *Andover, Hampshire.*

Transfer from Student to Member

WASPE, Peter Aubrey. *Burwell, Cambs.*

Direct Election to Member

BEDFORD, Keith Lewis, B.Sc. *Welwyn Garden City, Hertfordshire.*
BROWN, Anthony. *Malvern Link, Worcestershire.*
BUTTON, John Michael. *London E17.*
FERRER, Vyvyan John, B.Sc. *Llantarnam, Mon.*
HENK, Anthony John. *Emsworth, Hampshire.*
KNOPP, John William. *Olney, Buckinghamshire.*
PHARAOH, Peter Leonard George. *Kirkintilloch, Glasgow.*
SANDERS, Roy Herbert. *Quorn, Leicestershire.*

NON-CORPORATE MEMBERS

Direct Election to Graduate

BOWERS, David John, B.A. *Theydon Bois, Epping, Essex.*
CORNEY, Maximillian Edward. *Market Harborough, Leicestershire.*
POULSON, John Vincent. *Birstall, Leicestershire.*
PERERA, Weddippuli Arachchige Karunasena. *Bolton, Lancashire.*

Direct Election to Associate

LAU, Ivan Cheng Sen. *London, S.W.20.*

*(continued on page 649)*

# New Books Received

***All the books which are described below are available in the Library and may be borrowed by members in the United Kingdom. A postal loan service is available for those who are unable to call personally at the Library***

**Electronic Integrated Systems Design**

HANS R. CAMENZIND. Van Nostrand Reinhold, New York 1972. 23 × 16 cm. 332 pp. £7·50.

CONTENTS: I The Fundamentals: Interface. IC processes reviewed. The question of economy. II The Tools: Integrated Components: Resistors. Capacitors. n-p-n transistors. p-n-p transistors. Diodes, Zener diodes. Schottky-barrier diodes. Junction field-effect transistors. MOS transistors. Special devices. Maximum voltage, current, power and frequency. III Integrating The System: Partitioning a system or instrument. Digital elements, circuits, and subsystems. Linear elements, circuits, and subsystems. IV A Guide Through The Design Stages. Breadboarding. Layout. The role of computer-aided design. Testing. V The IC In Its Environment: Packaging and hybrid systems. Reliability.

Mr. Camenzind has produced a very useful and interesting book; however, it is possible that its impact may not be fully felt since its title does not accurately reflect its subject matter. Only about four of some 300 plus pages are concerned with system design, the remainder being dedicated to the design and manufacture of integrated circuits.

The book is essentially a very good guide for electronic engineers who wish to familiarize themselves with the principles and techniques used in integrated circuit manufacture. This branch of technology is very fast moving and it is refreshing to find such an up-to-date publication.

It is not unusual for books as concise as this one to completely miss their place in that they are too complex for the casual reader and too trivial for the specialist. This is not true of this book which, although it is mainly aimed at the newcomer to integrated circuits, contains much of interest to the experienced worker in the field.

I particularly appreciated the chapters on digital and linear circuits, which for ease of reference are arranged in alphabetical order, and the way that the text was generously illustrated. The book would be of immense value to anyone wishing to learn the details of integrated circuit design.

J. S. BROTHERS

(*Mr. Camenzind is founder and president of Interdesign Inc. and a faculty member of the University of Santa Clara.*)

**Thyristors and Their Applications**

P. ATKINSON. Mills and Boon, London 1972. 21 × 13·5 cm. 128 pp. £2·30.*

CONTENTS: Principles of thyristors. Control of voltage or current using thyristors. Frequency control. Gating a thyristor. Thyristor applications. Other applications of the thyristor. The triac.

Developed in the U.S.A. by General Electric, the thyristor has now been marketed for over a decade. Other than the excellent manufacturers' handbooks, few books have been devoted to the principles and applications of these devices. The purpose of this book is to summarize some of the fundamental aspects of thyristor technology and it is intended as an aid to engineers who have a sound working knowledge of electronics. Unfortunately however, the book does not meet the requirements as well as do some of the manufacturers' handbooks.

The detailed physics of the device has been omitted and the characteristics are presented in a way which the intended user should find simple and appropriate. Thyristor applications are dealt with under three main headings, Electrical Power Conversion, Power Regulation and Protection Technology. Applications not falling in the main categories are treated separately.

The data on the special applications is minimal and ignores some of the more effective work carried out in the various areas. Information on the use of thyristors in d.c. traction is extremely limited, yet this is one of the main areas where the device is employed. The section on latest applications is also limited in its scope, in fact the complete document would have been enhanced by featuring a number of examples with calculations to guide the user to use these versatile components to solve some of his problems.

The author lists as his sources of information one or two of the manufacturers' handbooks and a limited number of papers, mainly from the IEE Conference on Power Thyristors held in 1969.

To its credit the book is easy to read and is not cluttered with useless information, also the price is extremely reasonable.

J. HAMBLETON

(*Mr. Atkinson (Member 1962) is a Senior Lecturer in the Department of Applied Physical Sciences, University of Reading.*)

**MOS/LSI Design and Application**

WILLIAM N. CARR and JACK P. MIZE. McGraw-Hill, New York 1972. 25·5 × 17·5 cm. 331 pp. £9·05.*

CONTENTS: MOS device physics. The MOS technology arsenal. Reliability aspects of MOS integrated circuits. Inverters, static logic, and flip-flops. Shift registers for data delay, logic, and memory. The MOS/bipolar interface. Memory applications. Programable logic arrays. MOS/analog circuitry. The economics of MOS/LSI.

One drawback of a unified design for a dust jacket is that although a book is identified with its parent series it tends to be so identified that its title and substance do not seem to be new. This may be so for some of the McGraw Hill-Texas Instruments Electronics Series. However, 'MOS/LSI Design and Application' is new in that it represents 1972 Texas state-of-the-art—even the references at chapter ends are that recent. Although there is reference to analogue techniques, such as v.h.f. tuners and i.f. amplifiers for television, naturally the emphasis is principally on digital circuitry, with stress being laid more on design principle rather than 1972 achievements. For example, the section on memory applications discusses two read-only memories when engineers generally must be aware that their potential (and 1973 achievement) is distinctly in excess of this.

The book ends on a high note of achievement however, the one-chip calculator, an illustration of the economics of MOS/LSI. Some terms may sound strange to British ears, such as 'silicon real estate' but the implications for yield and surface area costs are clear enough, particularly for a chip of this size and complexity. The publishers claim that this book '. . . belongs on the desk of every systems engineer . . . and fabrication technologist interested in cutting costs and meeting competition'. With such a threat and at over £9 a copy could one possibly recommend it only for the library?

K. J. DEAN

(*The authors are respectively Professor and Visiting Industrial Professor of Electronics Sciences and Electrical Engineering at Southern Methodist University, Dallas, Texas.*)

**MOS Integrated Circuit Design**

Edited by E. WOLFENDALE. Butterworths, London 1973. 21·5 × 13·5 cm. 120 pp. £4·00.*

CONTENTS: The MOST physics and technology. MOS logic circuits. Logic design to layout. Computer aids for the designer. A simple MOS design example.

Electronic equipment designers who are involved in the use of custom-designed m.o.s. circuits will find this book a useful practical guide. For many years large-scale integration was a technique which could only be used by designers associated with component manufacturers, but the advances in m.o.s. techniques have now made it

## Book Supply Service

As a service to members, the Institution can supply copies of most of the books reviewed in the *Journal* at list price, plus a uniform charge of 25p to cover postage and packing.
Orders for these books, which are denoted by an asterisk (*) after the price, should be sent to the Publications Department at Bedford Square and **must** be accompanied by the appropriate remittance.

possible to design l.s.i. circuits with a well-defined set of design rules. This book describes the m.o.s. transistor with the basic design equations derived, physical effects explained and typical values given. An example of the design of a small chip shows how all the techniques described are used to solve a practical problem.

(*Mr. Wolfendale (Fellow 1968) is Managing Director of REDAC Software Ltd.*)

**Transistor Circuit Design**

LAURENCE G. COWLES. Prentice-Hall, Englewood Cliffs, N.J. 1972. 22·5 × 15 cm. 344 pp. £7·10.*

CONTENTS: The design method. Static characteristics and temperature problems. Equivalent circuits and gain-impedance relations. Amplified design principles. Common-emitter amplifiers. Transistor pairs. Transformer-coupled class-A amplifiers. Class-B power amplifiers. Feedback—a design tool. Feedback-amplifier stability. Video amplifiers. Radio-frequency tuned amplifiers. Transistor limitations—noise, breakdown, and temperature. Diodes and microwaves. Integrated-circuit amplifiers. Field-effect transistor amplifiers.

This reference manual of practical transistor circuits with design procedures and formulae covers d.c. to microwaves, and small-signal to high-power circuits, related to discrete components and integrated circuits.

(*Mr. Cowles is Senior Electronic Engineer at The Superior Oil Company, Houston, Texas.*)

**Electronic Display and Data Systems—Constructional Practice**

C. J. RICHARDS (Ed.). McGraw-Hill, Maidenhead, Berkshire 1973. 23 × 15 cm. 448 pp. £9·00.*

CONTENTS: Generation of electronic data. Transmission and refinement of electronic data (P. D. Goodenough). Display of electronic data (J. H. Rowe). Elementary aspects of c.r.t. design and usage. Mechanical design parameters associated with c.r.t. displays (C. J. Richards). Interconnection techniques (R. P. Selby). Manufacturing processes associated with printed circuit boards (C. A. Markwick). Packaging techniques associated with microelectronic circuits (L. A. Dyer). Cabinet and console design practice (W. J. Dimond). Electronic equipment shelters (R. P. Selby). Electro-mechanical aspects of radio interference suppression and systems earthing methods (C. J. Richards).

The rapid technological changes which occur in this specialized field of electronics make the compilation of a 'designers' guide to good constructional practice' a difficult and challenging task. Seven engineers have drawn on experience gained in designing and manufacturing defence equipment and have collaborated here to produce a very readable and informative book on the 'hardware' aspects of display and data systems.

It comprises eleven self-contained but related chapters, each with its own references and selected bibliography. Although aimed primarily at electronic, mechanical and installation engineers, production engineers who encounter problems at the design/production interface will find the subject-matter helpful although a little biased towards one company's experiences and practices. A notable omission is a chapter specifically dealing with engineering for production, including management of the transition from design to manufacture.

There are a few editorial errors, and in places the direct substitution of metric dimensions for Imperial gives the reader the impression that a high degree of precision in measurement is required. For the most part this is quite unjustified and it would have been better to have 'rounded up' (or down) in the conversion process. Much useful design data is presented in tabular form but it is felt that some readers would prefer to find this material in an Appendix rather than in the main body of the text. Welcome features are the encouragement given to engineers to adopt a modular approach to design, to set up data banks of 'preferred' components and sub-assemblies, and to use standard codes of practice wherever possible.

In summary, the shortcomings of the book are well compensated for by its virtues, and engineers involved in any way with design or manufacture of electronic hardware will undoubtedly derive benefit from the advice it has to offer.

J. M. PETERS

(*The editor and his six fellow contributors are senior engineers with Plessey Radar or, in one case, Plessey Telecommunications Research.*)

**Compatibility and Testing of Electronic Components**

C. E. JOWETT. Butterworths, London 1972. 21·5 × 13·5 cm. 345 pp. £6·00.*

CONTENTS: Integrated circuit compatibility. Survey of semiconductor joining techniques. Commercial application of thick film hybrids. Thin film hybrid approach to integrated circuits. Factors affecting thick film devices. Adhesion of platinum-gold glaze conductors. Thin inlays for electronic applications. Humidity corrosion of metallic resistors. The interface between glaze resistors. Confusion in multilayer systems. Dielectric formulations for screened ceramic. Microcircuit substrates. Failure modes in thin film circuits. Specifying resistance temperature stability. Poly-para-xylylene in thin film applications. Thin film $Al\text{-}Al_2O_3\text{-}Al$ capacitors. Compatible fabrication of tantalum thin film. Integrated circuits. Silicon oxide micromodule capacitors. Chip capacitors in hybrid microelectronics. Design construction and testing of miniature relays. Compatibility of flexible film wiring. The purpose of testing. Reliability screening using infra-red radiation. Environmental and life testing of magnetic components. Searching for incompatibility in integrated circuits. Encapsulated component stress testing.

In its short life time microelectronics has seen a tremendous growth and there are now a number of competing fabrication techniques. This book which covers a range of thin and thick film topics should be found useful by the process engineer who is often hard put to choose from the various materials and processes.

(*Mr. Jowett is an electronics consultant.*)

**Modern Control Theories: Nonlinear, Optical and Adaptive Systems**

FRIGYES CSÁKI Akademiai Kiado, Budapest 1972. 23·5 × 17 cm. 1096 pp. £16·00.

CONTENTS include: Characteristic features of nonlinear system. Methods of linearization. Transient processes. Characteristic features of nonlinear systems. Methods of linearization. Transient processes. The state-plane and phase-plane method. Piecewise linear systems. On-off control systems. Stability of nonlinear systems. Determination of the Lyapunov functions. Sampled-data systems. Absolute stability of nonlinear discrete-data systems. Optimal systems. Pontryagin's principle. Dynamic programming. Functional analysis in the solution of optimal control problems. Adaptive control systems. Optimizing methods.

This comprehensive survey is intended for engineers, and instead of aiming for strict mathematical proofs, practical methods of using the theories are presented, supported by numerous solved numerical examples and many problems. There are references in profusion widely drawn and up to date. The translation is of a standard that one is not aware of any such barrier between author and reader.

(*Professor Csaki is at Budapest Technical University.*)

**110 Thyristor Projects using SCRs and TRIACS**

Edited by R. M. MARSTON. Butterworths, London 1972. 21·5 × 13·5 cm. 138 pp. £2·40 (cloth), £1·40 (limp).*

CONTENTS: Basic principles and projects. 15 a.c. power-switching projects. 20 electronic alarm projects. 15 time-delay projects. 25 lamp-control projects. 15 heater-control projects. 15 universal-motor control projects. 5 miscellaneous projects.

Intended for students and electronics amateurs, this book describes projects which make use of thyristor devices able to handle mains voltage and high currents. These devices can be used in applications such as control of electric lamps, motors, heaters, alarm systems. All projects described are designed around internationally available components.

(*Mr. Marston is a consultant and a technical author.*)

**Introductory Topics in Electronics and Telecommunication.**

F. R. CONNOR.

**2. Wave Transmission** 99 pp.
**3. Antennas** 107 pp.

Edward Arnold (Publishers) Ltd, London 1972. 21·5 × 13·2 cm. £1·00 each.

These two books in a series of introductory topics in electronics and telecommunication will be found useful by students preparing for CNAA, CEI, HNC and other qualifications. 'Wave Transmission' provides classic definitions and partial differential solutions to open-wire and coaxial transmission line problems. Electromagnetic field theory is then dealt with and multimode field propagation in waveguides is treated.

'Antennas' treats the subject from elementary principles dealing with radiation, antenna arrays ending in quasi-optical solutions to large area.

(*Dr. Connor (Member 1965) was until recently a lecturer at the Polytechnic of Central London.*)

**Solid State Electronic Devices**

D. V. MORGAN and M. J. HOWES. Wykeham Publications, London 1972. 21·5 × 13·5 cm. 190 pp. £2·25 (paperback).

CONTENTS: The solid state. The electron. Electrons in solids. Charge carriers in semiconductors. Passive circuit elements. Bipolar junction transistors and field effect transistors. High frequency solid-state energy sources. Amorphous solid switches. Semiconductor technology. The device as a circuit element.

This book intended as an introduction to solid state devices is written for first and second year students at a university.

(*The authors are lecturers in the Department of Electrical and Electronic Engineering, University of Leeds*).