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

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Swapping Students

AN essential component in the 'formation' of the engineer is the period of practical training and work experience either during his academic studies or after qualification. There is however a well-established alternative available for at least a part of the latter period of training which seems to be unfamiliar and therefore under-used, namely the world-wide programme for the exchange of students for course-related experience, IAESTE.

This acronym, of uncertain pronunciation, stands for 'International Association for the Exchange of Students for Technical Experience', and it was formed in 1948 through the initiative of Mr James Newby, Superintendent of Imperial College's Vacation Work Committee, who became IAESTE's first General Secretary. In the first year the ten founder-member countries arranged for 920 students to be exchanged to spend periods of training during their summer vacations in various industries abroad. Since 1968 about 187,000 students have been exchanged and the annual totals are currently in the order of 5000 with 47 member countries.

The British organization, IAESTE UK, is administered by the Department of Education's Central Bureau for Educational Visits and Exchanges and, in collaboration with the other national agencies, arranges placements of students with cooperating organizations and firms. In the UK the participating employers currently range from a handful of electronics organizations, through chemical manufacturers, to county councils; in previous years the number has been well over 100 but now it is down to about 60. On the other hand nearly every university and most polytechnics and similar bodies participate by sponsoring students to go abroad. While there is no rigidly applied 'one for one' basis for exchanges, the fact remains that in 1981 the United Kingdom sent out just 125 students and received 88. (The main study fields for outgoing students represented were mechanical engineering (13), chemical engineering (11), electrical engineering (10) and chemistry (9).) Consider, however, the corresponding exchange figures for other developed countries: Netherlands 335/360, France 134/167, Austria 297/127, Finland 227/230, Germany 726/916, Switzerland 219/256, USA 128/138. It was not always like this: in 1960 the UK sent out 883 students and received 909, and was one of the most active supporters of the scheme.

The steady reduction in UK participation in IAESTE cannot be blamed on recession: for some reason, perhaps the lack of proper publicity, British educational institutions and industry seem to have lost interest. This is surely short-sighted, for while the educationalists are failing to take advantage of means of setting their students more firmly in their chosen profession, in the crudest terms industry is neglecting to use a means of establishing good opinions abroad for its products and services which arguably compares favourably with overseas advertising.

Intuitively, most will take the view that exchanges of this kind are 'a good thing' and the evidence from those with direct experience fully bears this out. Students benefit enormously from the challenges of new environments, albeit for only a few months. Participating firms speak highly of the motivation of those coming to this country and their ability and enterprise, and regard the relatively low cost as a bargain!

What can be done to foster this wholly admirable scheme and enable it to rise and perhaps exceed the peak of more than 20 years ago? For a start it should be considered with some urgency by our industry associations in encouraging their members to provide places for overseas students and equally urgently by the Engineering Council as having an important role to play in the 'formation' process for our own students.

F.W.S.

Institution Announcements

Changes to Charter and Bye-Laws

The Institution has been advised by the Clerk to the Privy Council that, on 10th March 1982 at the Court at Buckingham Palace, Her Majesty allowed amendments to the Charter which had been the subject of Resolutions passed at the Annual General Meeting on 29th October 1981. At a meeting on the same day the Lords of the Privy Council allowed appropriate amendments to be made to the Bye-Laws.

The major changes to the Charter and Bye-Laws permit an Associate Member who is registered with the Engineers Registration Board to use after his name the designation 'T.Eng. (CEI), AMIERE'. This identifies the specialization of qualified and experienced Technician Engineers who are members of the IERE in a similar manner to that adopted by Corporate Members.

A further change to the Bye-Laws provides that a member whose annual subscription is unpaid after 30th June shall not be entitled to attend meetings, receive publications or exercise the rights and privileges of membership until his subscription is paid in full.

The full wording of the changes now approved was set out as a Schedule accompanying the Agenda of the Annual Meeting and published in the July/August 1981 issue of *The Radio and Electronic Engineer*.

CEI EXAMINATION—'The Engineer in Society'

As many readers will be aware, this Institution has for some years recommended a change in the rules relating to this compulsory subject in Part 2 of the CEI examination. We are pleased to report that the Chartered Engineer Section Board of CEI has accepted our recommendation that persons who have passed all five technical subjects in that examination should be permitted an unlimited number of attempts at 'The Engineer in Society'. Unfortunately, the Institution's manpower resources do not permit it to undertake the total search of its examination records which would be needed to ensure that every past examination candidate who might wish to take advantage of this change of rule is aware of it. May we therefore ask any reader who knows a CEI Part 2 examination candidate who has failed 'The Engineer in Society' only, to bring the change to his notice?

Some Recent Developments

Earth Satellite Link for BBC Overseas Service

The Foreign and Commonwealth Office, through which some of the BBC's Overseas Service programmes are disseminated to an estimated 40 million regular listeners, has just placed an order with Marconi for a specially-designed Standard B satellite Earth station worth more than half a million pounds. To be located on Masirah Island, off the east coast of Oman, the Earth station will comprise a receive-only terminal and a ten-metre dish antenna. It is due for completion, as a turn-key project, by the end of 1982.

The Earth station will receive the BBC's Overseas Service programmes from England at 4 GHz via an *Intelsat V* satellite located above the Indian Ocean, and will convert these to broadcasting frequency. Broadcast channel bandwidth will be achieved by employing a specialized version of single-channel-per-carrier equipment in which two 3·2 kHz speech channels are combined to give a 6·4 kHz broadcast channel. Six such channels will be available through this link.

The Overseas Service programmes will be transmitted to the *Intelsat V* satellite by one of three Marconi-built Standard A Earth-station terminals at Madley in Herefordshire. At Masirah, two Marconi 750 kW m.f. transmitters and a range of high-power h.f. curtain-array antennas will carry out the broadcast transmissions to designated areas.

Fission Track Radiography for Checking V.L.S.I. Circuits

Since the 'soft error' was diagnosed in 1978, very-large-scale integrated (v.l.s.i.) circuit memory manufacturers have sought a method of detecting the minute amount of naturally-occurring radioactive impurities which, if present in v.l.s.i. circuit materials, can disrupt circuit performances.

Now, however, Harwell has developed an extremely sensitive technique known as fission track autoradiography (FTA) which can detect the present of uranium in concentrations as small as 2 parts in 10^9. This provides manufacturers with a quality control enabling them to assess raw material, and components, thereby reducing the risk of component failure.

The 'soft error' effect is produced by alpha-particle emissions from radioactive impurities present in any part of the v.l.s.i. circuit assembly. The energy possessed by an alpha-particle can produce an electric charge which may change the content of a single memory location, giving rise to computational errors. Because of this, semiconductor manufacturers are now specifying alpha-particle emission rates of less than 0·001 particles/cm^2/hour for their memory device materials.

It is not possible to detect such emission levels directly. The Harwell FTA technique exploits uranium-235, the fissile isotope present as 0·72% of natural uranium; prepared specimens of semiconductor material are coated with a polyimide film solid state nuclear track detector (s.s.n.t.d.) and irradiated with thermal neutrons in Harwell's Materials Testing Reactor, DIDO. On irradiation, the U-235 undergoes fission and the resulting fission particles are registered as tracks on the s.s.n.t.d. Afterwards the polyimide film is chemically etched to develop the fission tracks which can then be examined by optical microscopy. From the information gained it is possible to determine precisely the amount of uranium present, down to 2 parts in 10^9 (or a surface distribution of 3×10^{-6} μg/cm^2) and thus to calculate alpha-emission rates of as little as 0·0002 particles/cm^2/hour.

FTA has been used successfully to evaluate many of the material used in semiconductor manufacture, including silicon, silicide, gold and other metallic foils and resins and ceramics, and Harwell is providing commercial services to leading semiconductor companies in the UK, the USA and Japan.

Improving Television Picture Quality

The quality of present-day television pictures is limited by the available transmission bandwidth and by the capabilities of currently available display tubes. There is no very immediate sign of any large bright higher definition display device to take over from the shadow-mask cathode-ray tube, but many workers are in the field and we can expect some development during the next few years.

Assuming, then, that a better display becomes available, what possibilities are there for wider bandwidth transmissions to match? Both satellite broadcasting and optical fibre cable distribution offer wider bandwidth and the BBC has been considering how these could best be exploited. A key factor in any new transmission system must be compatibility, whereby existing receivers could continue to work with new-standard signals, although new receivers would be necessary to derive full benefit. For at least the early years of satellite or optical fibre cable services it would be required that existing receivers continue to be usable, with appropriate converters. The introduction of any non-compatible system could require many years for international agreement and new receiver development and hence seriously delay the establishment of satellite broadcasting in this country.

The present-day UK 5·5 MHz video transmission bandwidth is adequate for 625-line monochrome pictures. The limitations become apparent when the colour signals must be squeezed in with the monochrome. Ingenious though the PAL coding system may be, it is impossible to avoid some mutal interference between monochrome (luminance) and colour components. These interferences show themselves as luminance appearing in chrominance channels (cross colour) giving rise to flashes of false colour on striped suits for example; and chrominance signals appearing in the luminance channel give spurious dot patterns. To reduce these effects to acceptable levels, signals in the region of the colour sub-carrier (4·43 MHz) are attenuated, usually resulting in loss of all signal frequencies from about 4 MHz up to the 5·5 MHz band limit. So the majority of colour receivers roll-off about 4 MHz and show little fine detail whilst still suffering from some degree of cross-colour aberrations.

A new proposal involves filtering-off high-frequency components above 3·5 MHz. This gives a very slight reduction in picture definition, scarcely noticable on present-day display-tubes, but virtually removes all possibility of interference between luminance and chrominance components so that colour effects disappear.

In a wider-bandwidth satellite or optical fibre channel there is room to transmit the filtered-off high-frequency luminance components separately. The high frequencies (3·5 MHz upwards) are shifted in frequency to a higher band (8 MHz upwards) and transmitted together with the original low frequencies and chrominance signals. The upper limit of the separated high frequencies could extend above the 5·5 MHz equivalent bandwidth of the present transmssion channel.

A new receiver, specially designed for this wide bandwidth transmission system, would shift the transmitted high frequencies back to their original values (3·5 MHz upwards) and hence display a much-enhanced degree of fine picture detail. The new receiver would of course also be free from cross colour effects, since the high frequencies would be re-inserted after colour decoding had taken place.

The BBC has demonstrated experimental coders and decoders working on this principle and has passed extended bandwidth signals, with associated digital sound channels, through an r.f. link simulating a satellite channel. Results were very satisfactory and showed also that the proposed system is entirely compatible with continuing use of present-day receivers.

Study of Modern Materials in British Industry

The Fellowship of Engineering has received a grant from the Department of Industry for a study on the engineering use of modern materials. By making a number of case studies, the Fellowship will review the factors which have inhibited the introduction of modern materials (including new materials processing techniques and materials treatments) into British manufactured products, and will isolate the main factors which have inhibited innovation. It is also intended to highlight those material processes and treatments which have already been introduced into other industrialized countries and which have not yet apparently made any significant contribution to British engineering design. This will lead to a review of likely future trends in new materials, new products and new methods of materials processing, and suggestions will be made on how these might be used to commercial advantage by British firms.

The study will be carried out by a group of Fellows under the chairmanship of Dr. A. Kelly, FEng, FRS, Vice-Chancellor of the University of Surrey. The other members are: Mr. John Collyear, Managing Director of Associated Engineering p.l.c.; Mr. R. J. Dain of Ford and Dain Partners; Sir St. John Elstub, formerly Chairman of IMI; Professor Cyril Hilsum of the Royal Signals and Radar Establishment; Dr. R. Nicholson, formerly Managing Director of INCO Europe Ltd, now Chief Scientist at the Cabinet Office; and Dr. D. S. Oliver, Research Director of Pilkington Brothers Ltd. The Fellowship has asked Michael Neale and Associates of Farnham, Surrey, to assist them with the study.

Standard Frequency and Time Service

(Communication from the National Physical Laboratory

Relative Phase Readings in Microseconds
NPL—Station
(Readings at 1500 UTC)

FEBRUARY 1982	MSF 60 kHz	GBR 16 kHz	Droitwich 200 kHz
1	−10·0	35·0	—
2	−9·8	35·1	71·6
3	−9·8	34·0	71·4
4	−10·0	35·5	71·2
5	−10·0	35·5	71·0
6	−10·0	35·3	70·8
7	−10·0	35·5	70·6
8	−10·0	36·0	70·4
9	−10·1	34·6	70·3
10	−10·1	35·5	70·1
11	−10·2	35·4	69·9
12	−10·1	35·7	69·7
13	−10·2	35·0	69·5
14	−10·2	33·5	69·3
15	−10·2	34·3	69·1
16	−10·2	34·5	69·0
17	−10·3	33·2	68·8
18	−10·4	31·3	68·7
19	−10·4	32·9	69·6
20	−10·4	33·0	68·4
21	−10·2	33·2	68·3
22	−10·4	34·0	68·2
23	−10·4	33·2	68·1
24	−10·6	33·3	68·0
25	−10·4	32·0	67·9
26	−10·7	31·4	67·7
27	−10·5	33·3	67·6
28	−10·4	32·8	67·3

Notes: (a) Relative to UTC scale (UTC_{NPL}–Station = +10 at 1500 UT, 1st January 1977.
(b) The convention followed is that a decrease in phase, reading represents an increase in frequency.
(c) 1 μs represents a frequency change of 1 part in 10^{11} per day.
(d) It may be assumed that the satellite stations on 200 kHz at Westerglen and Burghead will follow the day to day changes in these phase values.

Conference Report

Industrial Applications of Learning Curves and Progress Functions*

Held in London on 4th December 1981.

Despite the publication of so many books and papers on learning curves and progress functions, it is surprising that this was the first known conference exclusively devoted to these important topics. This shortcoming is even more remarkable when it is remembered how many different professionals, including industrial engineers, cost accountants, personnel managers, and production controllers have a 'need-to-know'. Perhaps the explanation is that the diversity of backgrounds has prevented an integrated theme emerging around which a Conference can be based. This widespread application can best be understood by categorization by industry and reason for application of the technique.

Reported company users include those in the following industries:

Printing press operation and typesetting	Electronics	Cigar Making
Process plant	Housing	Aircraft Assembly
Machine tools	Steel Mills	Clerical Operations
Heavy electrical	Paper Mills	Watchmaking
	Electromechanical Assembly	Automobiles

Most users apply learning curves and progress functions for more than one reason, sometimes for as many as five of the following purposes:

Work standards and allowances	Forecasting delivery dates
Incentive schemes	Operator selection
Manpower planning	Equipment scheduling
Costing	Batch size determination
Measuring training schemes effectiveness	Setting WIP levels
	Line balancing

The Conference Organizers, Professor D. R. Towill and Mr R. A. Harvey therefore sought to bring together as many user papers as possible so as to maximize cross-fertilization between various industries and researchers. Although each author makes an individual contribution, there are many cross-threads between papers which give us considerable hope for the future of the emergence of a discipline in its own right.

E. A. Hackett (British Telecom) was generally concerned with learning curve models, although his curve fitting algorithms are more widely applicable. Eighteen models were initially considered, of which nine stayed the course up to the detailed analysis stage. The test data used to compare the applicability of models were obtained from a wide variety of sources. This included experiments on novice telephonists, since the author's brief covered the assessment of the effectiveness of training schemes devised for the Telecommunication industry.

The paper gives us confidence in the applicability of particular models, and computer algorithms for parameter estimation should we wish to move away from graphical techniques. Perhaps it is most valuable for indicating which models might let us down if we delegate all our thinking to the computer. Far better to use a model which works most of the time with reasonable accuracy than to opt for one which fits accurately occasionally, but which leads to disaster through computational instability!

Although there is much evidence that learning takes place in industry (as evidenced by global productivity figures) and can be studied in the laboratory on an individual task, there has been surprisingly little work done to link the two. This is the contribution to knowledge of F. W. Bevis (South Glamorgan Institute of Higher Education, Llandaff) who has been able to secure the confidence and co-operation of shop floor operators in reliably measuring long-term improvement in the performance of short cycle tasks. These included electronic assemblies, which were also simulated under idealized conditions. His experience suggests that the ultimate performance which can be achieved by individual operators is often beyond the understanding of experienced work study observers making judgements in the early phases of a production run. The improvement in performance he has recorded does help explain apparently conflicting attitudes towards methods-time measurement (m.t.m.) times. Early during a production run, the union-agreed standard time is sometimes as high as twice the m.t.m. value. Sometimes, depending on the task, the experienced operator can achieve a performance level corresponding to a cycle time of one-half the m.t.m. value. This highlights the problems facing work study observers in a 'learning' situation, which are worsened if union agreements permit timing over a small number of cycles only, thus making the trend much more difficult to establish. It is also observed that the methods adopted by the operator, and the elemental times recorded are affected by the payment-by-results scheme adopted.

Understanding the mechanism and constituents of the learning process can be tackled at the following levels

(i) total industry learning
(ii) total product learning
(iii) laboratory studies on sub-products and individual elements.

R. A. Harvey (British Aerospace Ltd., Weybridge) was concerned mainly with a contribution to knowledge on (ii). For total product learning, the source breakdown between organizational learning and operator learning is important to establish, since this will, in general, vary from industry to industry, and from new product to new product. Harvey argued that management can only decide on the most effective way to reduce learning costs if this breakdown is known. He was particularly concerned with the interaction of incentives with learning. A learning curve is necessary in order to make the appropriate bonus scheme allowances (covering both new tasks and new operators). It is plausible that some observed operator learning is at least partially due to financial incentives, although there would not be general support for the view that this would entirely explain the phenomenon of learning. His paper contains a multiple regression analysis applied to manufacture of aircraft sub-assemblies. A conventional progress function of 78% slope was found to be appropriate. However, the regression analysis showed that an 84% slope would be achieved if there were no individual operator learning, and a 75% slope if operators were allowed to work on the line without interruption. Numerical examples of the likely effects on productivity of operator transfer and rate-fixing initiatives are included.

The paper by G. S. Smith (Procurement Executive, M.o.D.) was concerned with progress function models, with examples

* IERE Conference Proceedings No. 52 includes the nine papers presented at the Conference, together with a historical bibliography of over 120 papers and books on the subject. (Price £20)

chosen from aeroengine and missile production. Procurement agency experience suggests that the 'straight line unit' model is more appropriate to his data, which come from an extensive 'learner library', commenced in 1955, and enlarged over the years. Computerized records are now available, and systematic analysis of data has been in progress since 1970 and organized into product groups such as aeroengines, missiles, electronic equipment, etc. The agency emphasizes the need for the estimator to establish a 'pitch point' on the learning curve, which they feel is best done by considering projection over a significant production run, say 250 items.

A number of discontinuities have been observed and are explained in the paper. Typical reasons for a change in unit costs are the transfer of a proportion of production to another site, and a reduction in raw material costs. Difficulties facing the Procurement agency when cost data are only provided on the basis of large batch orders are a salutary reminder to all of the need for the manufacturing organization to install proper cost monitoring procedures based on an adequate sampling rate of floor quantities and cycle times.

There are few behavioural science theories on which learning curve models can be based. In this respect the 'hardware' engineer has a considerable advantage in that the laws of physics can be utilized to derive a mathematical description of his device. If this does not fit the facts, then the engineer relaxes his assumptions and creates more complicated descriptions of the device until he is satisfied with the prediction obtained. Model complexity therefore increases in a planned manner. When the equations of motion have no known analytic solution, the engineer can resort to computer-based simulation to match the experimental data. With learning curves and progress functions, the model chosen usually depends on the experience of the analyst. For example, E. A. Hackett's paper considers nine such models in detail, but the majority are in no way related to each other; there is no common thread which allows us to move from one model to another.

Such a progression is, however, possible with one family of models, which was the subject of the paper of D. R. Towill (UWIST, Cardiff). This is devoted to learning curves (as distinct from progress functions). The best known and most useful member of the family is the time-constant model. This has been found to have wide applicability; in this Conference, for example, the industrial studies of F. W. Bevis and J. E. Cherrington are readily modelled in this way. His defined family of models does have a definite physical analogue in that the curves used are the step responses of linear systems of sequentially increasing complexity such as overcoming the effect of inertia in a spring-mass system. An advantage of the time-constant model is the easy interpretation of the model parameters, and these are also readily estimated via graphical or computer-based techniques.

In setting performance standards for heavy manual multi-operator tasks, J. E. Cherrington (Birmingham Polytechnic) found the time-constant model appropriate. The task requires planning and co-ordination between crew members, in addition to the carrying out of individual work functions. Considerable differences between performance standards achieved by the three crews manning the plant on a continental shift system have been recorded and explained via filmed sequences of the operation. A number of events can contribute to 'start-up' phenomena, with such a process including installation of new equipment and planned changes in work methods. The importance of a proper data monitoring system to account for management and worker induced discontinuities is emphasized. Otherwise, the true learning curve is distorted and the parameters wrongly estimated. Future predictions are then less accurate than need be, a point which also strongly emerges in G. S. Smith's paper. An analysis of work inefficiencies observed for the worst crew shows that the major factors, in order of importance, are (1) crane positioning losses, (2) synchronization of operations between worker losses, and (3) lack of integration between worker losses. His recommended solution includes closer management awareness and monitoring of start-up problems, even if these are apparently 'minor' events.

Ross Henderson (University of Manitoba) was concerned with management of new plant so that realistic process start-up performance, as measured by the progress function, is achieved. His experience is mainly in continuous casting steel plant start-up, in which he quotes one notable example of start-up commencing about a year late and then needing five years to reach full capacity. He points out that executives commonly expect new plant to surge suddenly to production at full capacity within a few months and make announcements accordingly. The reality is that the management of 'start-up' is a different problem to management of the 'steady state'. Task assignments, for example, are less quickly accomplished, with quite a few innovations required to make new technology work. The proposed solution is to regard efficient start-up as a problem solving exercise in which careful selection of new information systems, accounting systems, and methods of task co-ordination are important. An incentive plan for both management and workers is regarded as an integral part of start-up requirements.

Finally, C. Burnet (British Aerospace, Weybridge) took a global view of the progress function insofar as he was concerned with relating learning curve parameters to readily identifiable properties of the product, such as aircraft speed, geometry, and operations related equipment such as powered flying controls and radar systems. An important consideration therein is the definition of product complexity so as to separate this influence on learning curve parameters from the effects of increased worker efficiency. The origins of learning, prediction of learning, and optimization of the workforce/organization/product interfaces are seen to be areas where insight needs to be gained. If a product 'mock-up' or pilot plant, is used to optimize the manufacturing process, a good specimen too late is no better, and may be worse, than a bad specimen too early!

The Conference has assisted participants to determine the future use of learning curves and progress functions so as to improve productivity. We have seen that there is a considerable body of knowledge, and some controversial opinions on cause-and-effect, and indeed, on how results should be interpreted. It is, of course, necessary for each participant to make his own value judgement on these issues. The authors will not be available on-line to assist in this process.

For the future we can say that the microelectronics revolution will give companies unprecedented opportunitites to institute new forecasting and monitoring procedures. Distributed computing power at shop floor level will provide the detailed breakdown of cycle times sufficiently early for local action to be taken to diagnose and correct malfunctions whether due to management, men, machines, or material. This information can then flow to the production scheduling section for manufacturing process control and to the cost accountant for budgetary control. Further transmission to a central analyst can then result in updating of company-based learning curves (for which, as we have seen, the algorithms already exist), thus allowing inter-departmental and inter-company comparisons to be made. Thus management will have a second tier of feedback control to add to the short time-scale shop-floor detection and correction system already defined.

D. R. Towill

Contributors to this issue*


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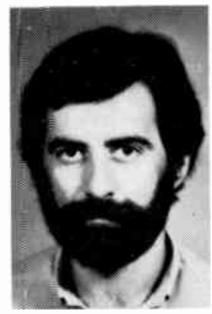

K. DAGAKIS


V. ZACHAROPOULOS

Philip Byrom is a graduate of the University of Liverpool where he read organic chemistry. He carried out research in organic photochemistry at the Royal Military College of Science, Shrivenham, before joining the Home Office Forensic Science Service in 1969. Dr Byrom is currently a Principal Scientific Officer and Head of a Chemistry Section at the Chorley Laboratory.

Peter Collar received the B.Sc. degree in physics from Bristol University in 1961 and the Ph.D. degree from Sheffield University in 1967, awarded for work on eddy currents in thin ferromagnetic materials. He joined the Institute of Oceanographic Sciences in 1968 after nearly three years spent at the Royal Radar Establishment, Malvern, and has worked on the development of sensors and instrumentation for the remote measurement of physical parameters. His present interests include the development and evaluation of techniques for measuring near-surface currents in the open sea.

Charles Clayson studied electronic and electrical engineering at the University of Birmingham between 1960–1966, attaining a B.Sc. in 1963 and a Ph.D. in 1967 for work on laser communication systems. He then joined the Institute of Oceanographic Sciences, Wormley and has worked on a variety of instrumentation and electronics projects ranging from a servo-controlled winch to telemetering wavebuoys. At present, he is mainly concerned with high resolution current and temperature measurements in the deep sea.

Pat Gwilliam started his career in electronics with the RAF from 1953–1957 as a ground radar fitter. He then joined the Royal Aircraft Establishment, first at Aberporth, then at Larkhill where he worked with a wide variety of equipment used on the missile test ranges. While with the R.A.E. he obtained his H.N.C. in electrical engineering with endorsements. In 1970 he joined the Institute of Oceanographic Sciences, Wormley, where he developed temperature and pressure sensors for deep sea tidal measurements. His present work is in developing a multi-instrument underwater package for deep ocean study of chemical fluxes.

Chris Hunter studied electrical engineering at Guildford Technical College and applied physics at Portsmouth Polytechnic. He has worked for 16 years on oceanographic instrumentation at the Institute of Oceanographic Sciences and has also worked for short periods at Woods Hole Oceanographic Institution in the USA and at the GEC Hirst Research Laboratories in Wembley.

Since 1972 Mr Hunter has been closely involved in the design, manufacture and subsequent operation of the UK Oceanographic Data Buoy, DB1. He is currently engaged in buoy instrumentation and in the development of an underwater telemetry system.

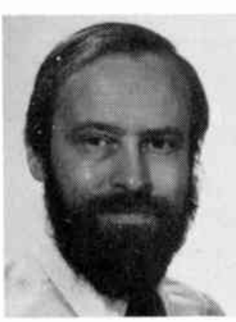
Charalambos Mantakas received the B.Sc. degree in physics at the University of Athens in 1957. He then undertook postgraduate studies in the Ecole Nationale Supérieure d'Electrotechnique, d'Electronique et d'Hydraulique of Toulouse, receiving his Ph.D. in electronics and telecommunications from Toulouse University (Doctorat 3^{e} cycle) in 1961. He also received a Ph.D. from the University of Athens.

From 1962–1971 Dr Mantakas was Head of the Scientific and Nuclear Instrumentation group of the Nuclear Research Centre 'Demokritos', Athens. During this time he worked also at the Centre d'Etudes Nucléaires, Grenoble, and in CERN, Geneva. From 1972–1974 he was Associate Professor of Telecommunications in the postgraduate course of electronics at the University of Thessaloniki. He then returned to the NRC 'Demokritos' where he established and has been in charge of the Digital Communications Group working now on voice and data encryption.

Kostas Dagakis received his Diploma of Electrical Engineering from the National Technical University of Athens in 1973. He worked in the Electronics Department of the National Defence Research Centre until 1977. Then he joined the Electronics Department of the Nuclear Research Centre 'Demokritos' where he has been working in the area of digital communications.

Vasilios Zacharopoulos graduated from the Faculty of Physics of the University of Patras in 1973. He is currently a research fellow in the Electronics Department of the Nuclear Research Centre 'Demokritos' and is working towards his Ph.D. in the area of digital communications.

**See also pages 223, 226, 234 and 258*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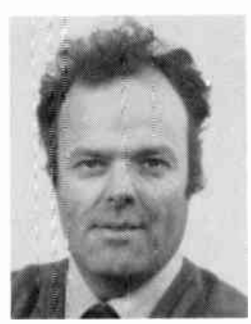
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General Interest Paper

Electronics in Forensic Science

PHILIP BYROM, B.Sc., M.Sc., Ph.D.*

The structure of the Forensic Science Service in England and Wales is outlined, followed by discussions of some of the many applications of electronics instrumentation in current use.

1 Introduction

The word 'forensic' is derived from 'forum', the Roman court of law. Forensic Science may therefore be defined as the application of scientific method to legal matters. In practical terms this generally means the scientific investigation of a wide range of criminal activities.

2 Organization of the Forensic Science Service

Evolution of the present Forensic Science Service in England and Wales began with the opening of a small laboratory serving the Metropolitan area in 1935 at Hendon. The Metropolitan Police now have comprehensive laboratory facilities at New Scotland Yard; the provinces are served by six Operational Laboratories administered by the Home Office. These latter laboratories are located at Aldermaston, Birmingham, Chepstow, Chorley, Nottingham and Wetherby.

Each of the Operational Laboratories undertakes to provide a full range of scientific services except for the specialized subjects of firearms investigation ('ballistics') and document examination (involving handwriting and typescript comparison). These particular activities are centralized at the Nottingham and Birmingham Laboratories respectively.

It is worth pointing out, in passing, that routine fingerprint work is carried out by specialist police officers working within their respective forces. The Forensic Science Laboratories have little involvement in this field.

A further organization, the Home Office Central Research Establishment (HOCRE), located at Aldermaston provides the Operational Laboratories with information services and certain advanced instrumental analysis facilities besides seeking to develop and evaluate new methods and instrumentation for use in routine casework.

Owing to popularization by the mass media the public probably has a somewhat distorted and glamorized view of the work of the Forensic Scientist. The fictional 'expert' is depicted as a jack-of-all-trades; solving cases, whatever their content or complexity, in splendid isolation and with amazing speed. The distinction between medical and pathology specialism and other branches of forensic science (chemistry, biology, etc.) is usually rather blurred in fiction. In reality forensic pathology is conducted by Consultant Pathologists retained by the Home Office. Although these Consultants are university or hospital based they are formally attached to an Operational Laboratory and close communications are maintained with the laboratory-based Forensic Scientists.

As a general rule cases involving more than one scientific discipline are seldom handled by an individual scientist. Rather, a teamwork approach is adopted in which the various aspects of a case are handled by scientists with relevant qualifications and experience. The scientific content of casework tends to fall into three main subject areas resulting in a natural departmental structure within an Operational Laboratory. This may be outlined briefly as follows:

BIOLOGY:	Offences against the person (rape, murder, assault, etc.) involving the examination of blood and other body fluids, hair etc.
CHEMISTRY:	Offences against property (burglary, criminal damage, road accidents, etc.) involving analysis of glass, paint, soil, explosives, etc.
DRUGS & TOXICOLOGY:	Drugs offences and cases of poisoning (intentional or otherwise)

The Operational Laboratories are committed to the analysis for alcohol of many thousands of blood or urine samples per annum under the provisions of the Road Traffic Act 1972. Thanks largely to automatic instrumentation this activity occupies a very small proportion of the available man-hours and is usually carried out on a rota basis by a number of authorized analysts drawn from Chemistry and Drugs & Toxicology Departments. The instrumentation and procedure for alcohol analysis will be described later.

Casework in the Operational Laboratory is undertaken by scientists having appropriate qualifications and experience. In-house training is provided in a wide range of subject areas and many junior staff take advantage of day-release facilities in order to gain higher qualifications.

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Selected groups of Police Officers receive training to ensure that the best possible use is made of laboratory facilities. Police Liaison Officers play an important role in the running of the laboratory. These officers are co-opted from Police Forces within the laboratory area.

Most forensic science investigations exploit the principle of contact trace analysis. Whenever two objects, animate or inanimate, come into contact there is inevitably a transfer of material between the two. Thus in an affray one can expect transfer of clothing fibres, hairs and possibly blood between the participants. Likewise when two vehicles collide in a road accident paint will be transferred.

The practical value of these types of evidence will depend on the sensitivity and specificity of the methods used for analysis as well as upon the intrinsic nature of the materials concerned. It is the pursuit of enhanced sensitivity and specificity coupled with the need to cope with ever-increasing case loads that have led to a dramatic increase in the use of electronic instrumentation in forensic science in recent years.

Many modern scientific instruments are designed so that they can be operated by virtually anyone, given a minimal amount of instruction. Others, however, are of such complexity that a specialist operator is required for realistic results to be obtained. For these reasons the more advanced instruments tend to be centralized in the Operational Laboratory, thus providing an analytical service.

No discussion of modern scientific instrumentation would be complete without mentioning the role of computers and microprocessors. These devices have added immensely to the ease with which complex instrumentation systems may be operated and with which analytical data may be processed and interpreted. Several of the instruments to be described later are microprocessor-controlled or have dedicated data processing facilities.

Each Operational Laboratory is equipped with a computer terminal comprising visual display unit, printer and British Telecom modem. This allows link-up to a large mainframe computer located at HOCRE, Aldermaston, for purposes of information retrieval (e.g. literature references) and interrogation of data banks to assist in assessing the evidential significance of casework findings.

3 Electrophoretic Techniques

Although most of the applications of electronics instrumentation are to be found in Chemistry and Drugs & Toxicology Departments of the operational forensic science laboratory there have been very significant advances in recent years in methods used by the biologist to group blood and other body fluids. Most people will be aware of the ABO system of blood grouping since this is widely used in hospitals for blood transfusion purposes. This system is also used by the forensic biologist but there are, in addition, several other grouping systems which depend on genetic variation of certain enzyme types. These greatly enhance evidential value when used in combination. The phosphoglucomutase (PGM) enzyme, in particular, can be resolved by electrophoresis techniques. In this method the enzyme molecules are made to migrate in a conductive gel medium by application of an electric field. The genetic variants migrate at different rates under standard conditions and can be visualized as discrete spots or bands by treating the gel with a specific chemical reagent causing colour development.

The traditional technique uses a starch gel medium maintained at a specified pH (acidity) by a chemical buffer and enables PGM to be resolved in three variants.

Iso-electric focusing, an electrophoretic technique involving the initial establishment of a pH gradient across a polyacrylamide gel, was recently introduced with considerable improvement of resolution. In this way the PGM system can be resolved into ten variants thus enormously increasing the evidential value of PGM. The instrumentation required for electrophoresis is fairly simple, consisting of a stabilized d.c. power supply and a means of providing thermostatic cooling to counteract ohmic heating in the gel medium.

This approach could well be applied to other enzyme systems in the future, the only serious practical restraints being the stability of the particular system under adverse storage conditions and the sensitivity of detection methods. Enhancement of the sensitivity of existing methods of detection might well be possible by use of instrumentation.

4 Gas Chromatography

In the other departments gas chromatography is one of the most versatile instrumental techniques in current use, finding applications in drugs analysis, blood alcohol analysis and many areas of chemistry. This technique relies on dynamic equilibrium of substances between the vapour phase and a solid or liquid phase; the latter generally being deposited as a thin film on a granular solid support. The vaporized sample is passed through a tube packed with the solid (or supported liquid) phase by means of an inert 'carrier' gas, typically nitrogen, argon or helium. Substances which have little affinity for

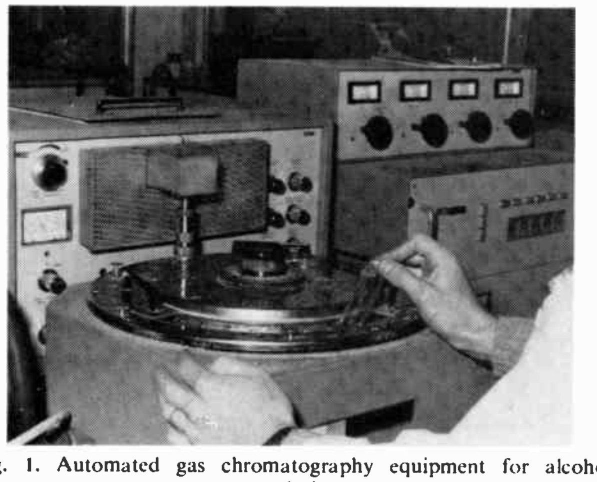

Fig. 1. Automated gas chromatography equipment for alcohol analysis.

the packing material pass quickly through the system whereas those with greater affinity take longer. Thus samples which are mixtures become separated. The separation is visualized on a chart recorder by some form of electronic detector located at the exit of the tube.

Many types of detector are available but two are worthy of description in the context of this article, namely the flame ionization detector and the electron capture detector.

In the flame ionization detector the carrier gas and emerging sample components pass into a hydrogen flame burning in a controlled air atmosphere. Polarized electrodes in the flame allow an ionization current to be monitored with the aid of a high-gain amplifier. In the absence of sample, i.e. carrier gas only, this current is very small and we observe a low-level baseline. The ionization current increases upon arrival of sample-components in the detector, the effect being particularly dramatic with organic compounds (compounds of carbon) such as may be found in paints, plastics, rubbers, fuels, drugs, etc.

The electron capture detector contains a small radioactive source emitting beta particles (i.e. electrons). Thus a steady current is observed with only carrier gas in the detector. Arrival of molecules which absorb or capture electrons causes the current to decrease. Current is monitored via an amplifier and chart recorder. Substances which lend themselves to electron capture detection include chlorinated or nitrogen containing species, e.g. many pesticides, and certain ingredients of explosives. Portable gas-chromatographic equipment incorporating electron-capture detection is available for field use in detecting traces of explosives in vehicles or premises.

Alcohol analysis, for the purposes of the Road Traffic Act 1972, is also achieved by gas chromatography. The sample of blood or urine is diluted precisely with a solution containing a reference compound or internal standard. Analysis by automated headspace gas chromatography results in separation of volatiles present in the sample. In practice these volatiles consist almost entirely of ethyl alcohol (the subject of the analysis) and the internal standard. Thus essentially two peaks are displayed on the chart recording. Digital processing allows the analytical signals to be related to calibrations conducted with standard alcohol solutions, thereby producing quantitative results.

Analysis of headspace, or vapour above the diluted sample, achieves a preliminary clean-up and avoids contamination of the instrument with undesirable materials such as blood solids.

Samples are always analysed at least in duplicate using completely independent sets of apparatus, thus avoiding instrumental errors. Furthermore standard alcohol samples are introduced at regular intervals to check instrumental performance.

Sample containers supplied for police use contain additives to arrest bacterial action and, in the case of blood, to inhibit clotting. The fluoride preservative levels of samples may be checked using a specific-ion electrode connected to an electronic voltmeter.

Urine and blood samples are treated in exactly the same way since, for the purposes of analysis, they are merely liquids containing alcohol. The legal limit is however, 80 mg of alcohol per 100 ml of blood and 107 mg of alcohol per 100 ml of urine.

Recent legislation will allow the introduction of automatic breath alcohol analysis equipment in order to streamline the current procedures. Tests carried out on prototype instruments have compared favourably with concurrent blood analyses.

The prototype instruments evaluated thus far operate on various principles ranging through automatic colorimetric analysis of chemical reagent solutions, gas chromatographic analysis, infra-red absorption and fuel cell devices. The latest generation of devices are microprocessor controlled. Large-scale introduction of this equipment will provide considerable impetus to the electronics manufacturing and servicing industries. Clearly considerable scope for innovation exists in this area.

5 Particle Examination

Automated equipment has long been used in hospital laboratories for relieving the tedium associated with blood cell counting. This same equipment has been adopted by the forensic scientist for soil particle analysis. A slurry comprising the silt fraction of a soil sample, i.e. particles smaller than 63 microns, suspended in an electrolyte solution is passed through a fine glass capillary tube. Electrodes at each end of the capillary allow the conductivity of the system to be monitored. Soil particles passing through the capillary influence the conductivity to an extent dependent on their size. Suitable data processing results in a histogram plot of particle count against size which may be used as a sensitive means of comparing soil samples. In this way soil samples weighing well under one gram may be processed, a considerable improvement on the traditional methods of soil analysis.

6 Measurement of Optical Properties

Glass is encountered in many aspects of crime and numerous types of glass are in existence. Refractive

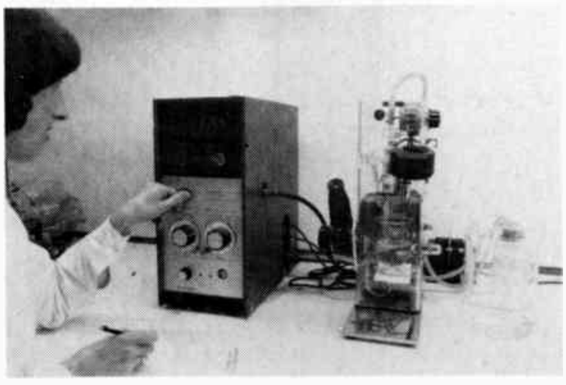

Fig. 2. Blood cell counting apparatus adapted for use in soil analysis

index is a convenient parameter to measure since it is strongly dependent on chemical composition and chemical composition varies considerably according to glass type and origin. Traditional methods of measuring refractive index require an optically flat polished specimen. Virtually all glass encountered in forensic science casework, however, takes the form of small irregularly-shaped fragments. Refractive index of such material is measured by immersion in a liquid and observation through a microscope using monochromatic (sodium) light.

In the event of a difference in refractive index between the liquid and the glass fragment a bright halo ('Becke line') is observed around the fragment. When the refractive indices coincide the Becke line disappears. This is brought about by adjusting the temperature of the system, refractive indices of liquids being much more temperature-sensitive than those of solids. Measurements are conducted in equipment designed for micro-melting-point determinations whereby the microscope stage is heated in a controlled manner and the temperature monitored by a platinum resistance thermometer. The operator may record the temperature at the chosen point on a digital read-out. A range of silicone oils is used for immersion. These are calibrated for use with the aid of a range of standard glasses having known refractive indices. Efforts are being made to automate the procedure, the most promising approaches involving closed-circuit television monitoring with computerized photometry facilities for match point determination. Alternatively a photo-diode array inserted in the microscope field would allow monitoring of Becke line behaviour.

7 Atomic Absorption Spectrometry

The ability to identify and accurately measure small amounts of a particular chemical element is of great value, particularly if methods are available which require minimal sample preparation and present few problems as regards interference from other elements or matrix effects. These criteria are largely satisfied by the technique of atomic absorption spectrometry which involves thermal breakdown of the sample into its constituent atoms. This is achieved by introduction of the sample as a solution into an extremely hot flame, typically acetylene burning in nitrous oxide. Atoms of the selected element are detected and quantified by passing light from a hollow-cathode discharge lamp containing the element of interest through the flame and measuring the amount of light absorbed by the atoms using a photomultiplier detector linked to a monochromator set to an appropriate wavelength. Sensitivity of this method may be enhanced by several orders of magnitude if the flame is replaced by an electrothermal atomizer; a popular form of this device consists of a graphite tube heated by an electric current in an inert atmosphere.

The usefulness of atomic absorption techniques would be enhanced considerably by development of viable multi-element facilities; current instrumentation is applicable only to single, selected elements.

Fig. 3. Atomic absorption spectrometer. Sample introduction and digital readout

8 Fluorescence

Many substances exhibit the property of fluorescence whereby excitation with light of a particular wavelength results in emission of light at longer wavelengths. Instrumentation is available consisting of a continuous light source, usually a xenon discharge lamp, with a monochromator to enable selection of the excitation wavelengths. Fluorescence emitted by the sample is analysed by a second monochromator equipped with a photomultiplier detector.

Two modes of operation are in common use:

(1) Using a fixed wavelength of excitation the detector monochromator is scanned to give an 'emission spectrum'.
(2) With the detector monochromator set at some optimum wavelength the excitation monochromator is scanned. This gives rise to an 'excitation spectrum' which is, incidentally, analogous to the absorption spectrum of the fluorescing species,

It has been found to be convenient in practise to scan both monochromators at the same rate with a fixed wavelength separation between the two. This is known as 'synchronous scanning' and various wavelength separations may be used to obtain maximum information. The results of synchronous fluorescence scans are difficult to interpret theoretically but they enable very sensitive comparisons to be made between samples. For example there is a high level of confidence in the comparison of traces of lubricating oil on the clothing of 'hit-and-run' accident victims with oil samples from suspect vehicles.

It may be possible with advances in instrumentation to record fluorescence information from microscopic material. Research has already shown, for instance, that drugs may be characterized in individual cells of tissue specimens.

9 X-ray Diffraction

Several techniques involving the use of X-rays are popular in forensic science because of the minimal sample preparation requirement and their non-destructive nature. This latter aspect is of particular importance since the sample is available for analysis by other methods if necessary or it can be retained for production in court.

X-ray diffraction makes use of the property of crystals to reflect X-rays in different directions and with different relative intensities depending on the exact nature of the atoms in the crystal and their spatial arrangement.

The interaction of X-rays with a crystal lattice obeys the Bragg equation:

$$n\lambda = 2d \sin \theta$$

where n = integer
λ = wavelength of X-rays
d = spacing of atom layers in the crystal
θ = angle of incidence and reflection of X-rays with respect to the atom layers.

The X-ray diffractometer works by subjecting the sample to a collimated beam of filtered X-rays and monitoring the reflected pattern either by a photographic film placed around the circumference of a circular camera or by a Geiger counter which is automatically scanned in a circular path around the sample. In the latter case results appear as a chart recording of intensity against angle. By reference to tabulated data or to a computerized data bank the individual crystalline components of a sample may be identified. Alternatively data may be used fingerprint-fashion for comparison purposes. The types of material routinely examined range from improvised explosives mixtures to paints, cements and plasters, geological material, etc.

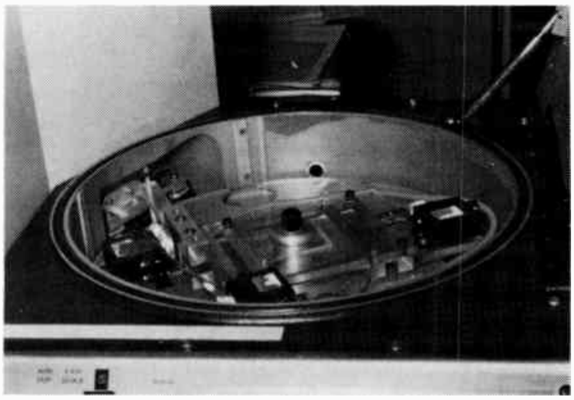

Fig. 4. A device for remote sample positioning in the x-ray fluorescence spectrometer

10 X-ray Fluorescence

Another form of X-ray analysis, X-ray fluorescence, involves irradiation of the sample with X-rays and analysing the emitted secondary X-radiation. The energies of the secondary X-rays depend on the chemical elements present in the sample. Lithium-doped silicon semiconductor detectors are available for this purpose and multi-channel analysers enable data to be displayed in the form of X-ray count as a function of energy. Digital processing permits rapid analysis by automatic identification of the elements present in a sample. X-ray fluorescence requires virtually no sample preparation; flakes of paint, glass fragments, counterfeit coins, etc., are amenable to analysis without pre-treatment other than, perhaps, lining up the sample in the centre of the incident X-ray beam.

Fig. 5. Scanning electron microscope equipped with microprobe analyser

Very small samples can present problems of alignment since positioning is critical for meaningful results to be obtained. Since measurements are conducted in an evacuated chamber the optimum positioning of a small sample can be very time-consuming because re-positioning requires release of the vacuum and several minutes are required for re-evacuation before the effects of any adjustment can be observed. To overcome this problem a remote 3-dimensional positioning device was constructed in the author's laboratory. This utilizes three servo mechanisms, as used in radio-controlled models, linked by cable to a three-channel controller. The latter uses 555 timers to produce suitably variable square-wave trains for servo operation. Thus the sample position may be adjusted precisely from outside the chamber without the need to disturb the vacuum.

11 Scanning Electron Microscope

The scanning electron microscope (SEM) finds many applications in Forensic Science, not only because of the superior imaging capabilities as compared to optical microscopy but the fact that the sample is bombarded with electrons inevitably results in the emission of X-rays. As in the case of X-ray fluorescence mentioned above, the emitted X-ray energies will depend on the particular chemical elements present in the sample. Linkage of the SEM to a semiconductor detector and data processor thus enables chemical analysis of microscopic samples. This combination is known as a Scanning Electron Microscope Microprobe Analyser, (SEM-MPA).

The usual specimen manipulation facilities of the SEM coupled with the high magnifications achievable allows chemical analysis of a small selected sample area. Thus, for example, the individual layers of a multilayer paint flake may be analysed. This technique is currently being used for the identification of gunshot residues which might be expected to be found on hands and clothing of a person who has used a firearm.

Advances in software are already enabling complex data from XRF and SEM-MPA instrumentation to be processed rapidly and presented to its best advantage. It should soon be possible, for example, to identify glass types automatically on the basis of multi-element analysis, reference being made to stored 'known' data.

12 Mass Spectrometry

Organic molecules as occurring in drugs, fuels and a whole host of products consist largely of a skeleton of carbon atoms with other light elements such as hydrogen, nitrogen and oxygen attached. The structural permutations possible with only a small number of such atoms are numerous. Consequently special methods are required to elucidate the structures of organic molecules and such a method is mass spectrometry. If an organic molecule is bombarded with high-energy electrons it will become fragmented and lose some of its own electrons in the process giving rise to positively charged fragment ions. In the mass spectrometer these ions are accelerated and focused before entering a magnetic field. In this area the ions assume a circular orbit in which centripetal force is balanced by magnetic forces, thus:

$$\frac{mv^2}{r} = Bev$$

or rearranging: $$\frac{m}{e} = \frac{Br}{v}$$

where m = ionic mass
e = ionic charge
B = magnetic field strength
r = radius of orbit
v = velocity of ion

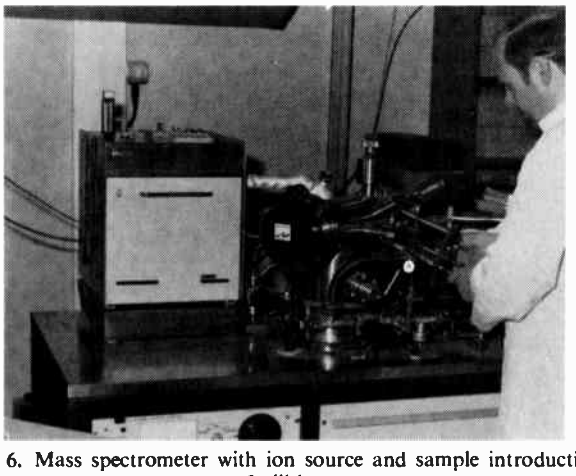

Fig. 6. Mass spectrometer with ion source and sample introduction facilities

The mass charge ratio (m/e) will characterize a given ion and may be determined from the instrumental parameters. For convenience the magnetic field may be scanned by supplying the electromagnet coils with a sawtooth waveform. Thus a range of ions can be repeatedly swept past the slit of a detector to give a spectrum which is linear in mass charge ratio. The detector for a mass spectrometer consists typically of a structure resembling a photomultiplier and working on similar principles. Interpretation of the fragment spectra produced may be facilitated by digital data processing. Many variations of data presentation are available for eliminating unwanted interferences or enhancing sensitivity by concentrating on a particular species with exclusion of all others. Several methods of sample presentation are provided on most instruments thus allowing a full range of materials to be examined. Vapours or volatile liquids are injected through a septum seal and allowed to bleed into the ion source through a capillary. Solids may be inserted on the end of a probe which enters the source via a vacuum lock, and are then volatilized by electrical heating. The most powerful device for sample introduction is, however, the gas chromatograph whereby the effluent is admitted directly to the ion source, sometimes with removal of excess carrier gas. In this way the individual components of a mixture, as they are separated by the gas chromatograph, may be structurally characterized as they enter the mass spectrometer.

Mass spectrometry allows rapid and precise identification of drugs and metabolites at levels often many orders of magnitude lower than previously measurable by traditional chemical techniques. It is also possible, for instance, to identify petrol at trace levels in fire debris on the basis of lead alkyl anti-knock additives which may be selectively detected by mass spectrometry.

Alternative modes of ionization will increase the scope of mass spectrometry techniques but advances in computerization and software will provide the key to optimization.

13 Conclusion

This discussion of electronics instrumentation in forensic science is by no means exhaustive but it is hoped that it will provide some insight of an area unfamiliar to most Electronics and Radio Engineers.

The Forensic Science Service owes a debt of gratitude to Electronics Research and Development and to many facets of the Industry for numerous contributions to the current 'state of the art'. There will undoubtedly be more exciting developments in future years as a result of the tremendous rate at which progress is being made in the field of Electronics.

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Indexing Terms: Diving apparatus, Helium–oxygen systems, Signal processing, Speech analysis and synthesis

The helium speech effect and electronic techniques for enhancing intelligibility in a helium–oxygen environment

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and

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SUMMARY

This paper considers the nature of the speech mechanism, and the effects on the speech spectrum of a high pressure helium–air environment. A comparison is made between the characteristics of speech distortions in a helium–air mixture and certain well-known characteristics of speech in normal air which give rise to similar effects.

The criteria for good intelligibility are related to the performance of various helium speech unscrambling techniques which have been used. These unscrambling techniques are classified here into two main categories: those essentially using signal processing in the frequency domain and those using signal processing in the time domain. Consideration is also given to waveform coding techniques which involve a combination of both of these classes of signal processing.

The ability of these categories to incorporate the various features required for good intelligibility in unscrambling helium speech is discussed, in order to highlight the potential importance of frequency domain approaches to future unscrambling developments.

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1 Introduction

Deep sea diving is a specialist industry which has grown dramatically in importance within Europe as a direct result of oil exploration and production in the North Sea. Divers provide an essential service without which underwater construction, inspection and maintenance tasks could not be carried out. At the same time diving is a difficult and hazardous occupation and in consequence a great deal of time and money has been spent to improve operational efficiency and the personal safety of the divers. The present widespread use of saturation diving techniques means that divers can now work in the sea at great depths for long periods of time and working dives to 700 metres are now possible. The frequency, complexity and cost of such dives have focused attention on the need for improved diver-to-diver and diver-to-surface communication.

Saturation diving refers to the technique whereby the diver can be considered as 'saturated' with gas for a specific depth and can remain at such depths without danger. Decompression time is necessary because dissolved gases in the body must be given sufficient time to reach a new equilibrium with the reducing ambient pressure during ascent from a dive. Failure to decompress the diver at a slow enough rate causes the condition known as 'the bends', in which bubbles of trapped gas are released in the tissues. In extreme cases, bubbles can appear in the bloodstream itself with possibly fatal consequences. Nitrogen, which constitutes 78% by volume of air, is an especially troublesome gas in this respect and requires long decompression times, even at shallow depths.

In saturation diving beyond approximately 50 metres, breathing normal air under pressure causes enhanced absorption of nitrogen into the bloodstream, and promotes a dangerous state of narcosis in the diver, reducing his capacity to work accurately and increasing the possibility of a fatal error in judgement.

The use of a gas mixture with reduced nitrogen content obviates this dangerous situation. The simple solution of increasing the proportion of oxygen in the breathing mixture is however to be avoided since oxygen gas dissolved in the bloodstream can cause convulsions if the partial pressure of the oxygen in the mixture exceeds two atmospheres. Further, at depth, the density of normal breathing mixtures would render breathing difficult and therefore use is made of a gas mixture containing high proportions of a gas with low molecular weight. Helium, which has the additional advantage of being an inert gas, is normally used. The use of helium in the respiratory gas mixture (of the order of 95% helium, 3% nitrogen, 2% oxygen), thus reduces the probability of these dangerous physiological and mental problems, but it creates a serious communication problem for those who must attempt to understand the 'squeaky' voice emissions of the diver. It is these distorted voice sounds which are termed 'helium speech'.

The distortions of speech sounds uttered in helium, when compared to similar sounds in air, render the speech unintelligible and the need for electronic techniques to 'unscramble' the helium speech has been identified.

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This paper considers in detail the nature of the speech mechanism, and the effects on the speech spectrum of pressure and of a helium environment are discussed. A comparison is made between the characteristics of speech utterances in helium and certain well-known characteristics of speech in normal air which give rise to similar effects.

The criteria for good intelligibility both with respect to the listener and with respect to the speaker himself are discussed, and these are related to the fidelity of the various unscrambling techniques which have been used.

A number of helium speech unscrambler techniques have been proposed to date, and these are classified here into two main categories: those essentially using signal processing in the frequency domain, and those using signal processing in the time domain. Consideration is also given to waveform coding techniques which involve a combination of both of these classes of signal processing.

Finally, the factors necessary for good intelligibility from unscrambled helium speech are re-iterated, and related to the performance in terms of intelligibility of present-day unscrambler devices.

2 The Speech Mechanism

In normal speech in air, the structure of the speech waveform is determined by a complex interplay between the shape of the vocal tract from the vocal cords to the lips and nose, by the shape of the excitation waveform injected into the vocal tract, and by the speed of sound in the exhaled gas mixture (Fig. 1). There are three main

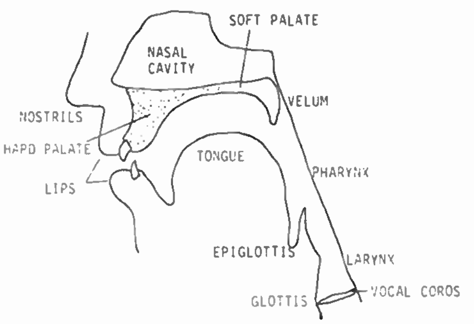


Fig. 1. Section through human vocal apparatus.

ways in which the airflow from the lungs is converted into an acoustic signal with components in the audio range. Firstly, the vocal cords situated in the larynx are adducted, or drawn together, while air is blown from the lungs through the glottis. Very rapidly, sub-glottal pressure builds up as the lungs continue to expel air, until the pressure necessary to blow the vocal cords apart again is reached. The cords separate under this pressure, allowing an impulse of air to be injected from the over-pressured sub-glottis through the gap between the vocal cords. This gap effectively takes the form of a Venturi tube due to the fashion in which the cords separate and as a result of the local drop in air pressure in the constricted passage between the cords, coupled with the elastic tensions acting in and on the vocal cords and the relieved sub-glottal pressure, the vocal cords are forced back towards each other. The instant of glottal closure is usually the point at which excitation of the vocal tract is most powerful.[1–3] This process is repeated at a rate of between 50 and 400 times a second in the voiced speech uttered during ordinary conversation. The spectrum of the vocal cord excitation exhibits a harmonic structure in which the strongest component is normally the fundamental repetition rate and the fall-off with increasing frequency is around 12 dB/octave. Vowel sounds such as 'ah' or 'ee' are produced by this mechanism.

The second main sound source consists of air turbulence produced at a constriction somewhere in the vocal tract. Since it involves a continuous stream of air, the spectrum of this sound source exhibits neither periodicity nor any harmonic structure, and resembles white noise with an essentially flat power spectrum. Sounds formed in this way are generally characterized by a 'hissy' quality, and are termed fricatives. Examples of fricatives include 'ss' or 'ff'.

The third type of sound source results from the build-up of pressure which occurs when the vocal tract is closed at some point, such as by the lips or tongue. A sudden release of pressure causes a transient excitation of the vocal tract which results in a sudden onset of sound. The speech resulting from this type of excitation can be classified into two types: voiceless stop consonants, where the vocal cords are not vibrating during the closure and the onset is preceded by silence (such as 'p' or 't'); and voiced stop consonants, where the vocal cords are vibrating during the closure and the onset is preceded by a low intensity voiced sound, such 'b' or 'd'.

Sound energy from these sources enters the vocal tract, whose configuration, as governed by the positions of the tongue, jaw, velum and lips (Fig. 1), determines its acoustic properties and modifies the spectrum of the sound source. The resonances of the vocal tract cause concentrations of energy at certain frequencies which are known as formants. The coupling at the point of radiation (lips/nostrils) produces a high-frequency emphasis of 6 dB/octave.

3 Effects of Ambient Pressure on the Speech Spectrum

Although the velocity of sound is essentially independent of air pressure, when a diver speaks in high pressure air

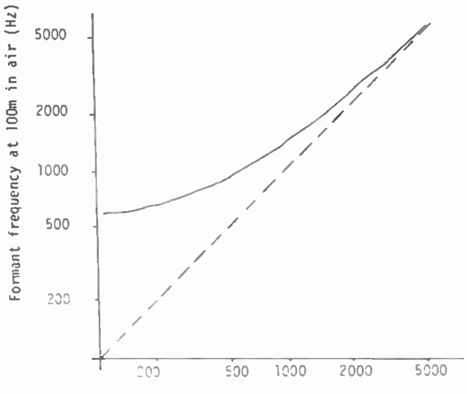


Fig. 2. Formant shift in high pressure air (from Ref. 6).

his speech exhibits a pronounced nasal quality and a steady decrease in speech intelligibility[4,5] is produced with increasing depth and air pressure. Published comparisons[6] of voice formant frequencies in air at atmospheric pressure and in air at high pressure (3 atmospheres) indicate a non-linear shift in formant frequencies (Fig. 2). This effect is considered to be a result of a variation in acoustic impedance mismatch between the vocal tract walls and the high density air.[7,8]

4 Effects of Helium–Oxygen Mixtures on Pitch Period

Published results[9,10] suggest that the fundamental period of repetition of the vocal cords (the pitch (or larynx) period) does not vary as a function of the composition or pressure of the breathing environment.

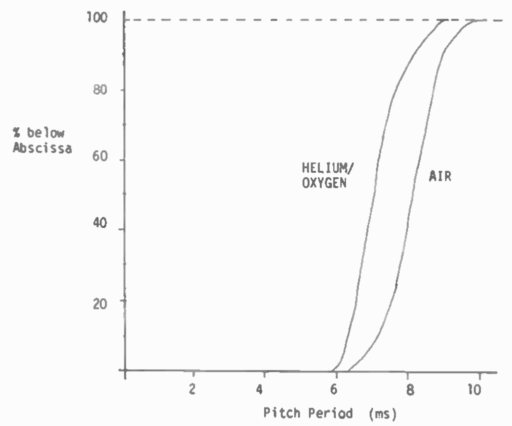


Fig. 3. Cumulative distribution of pitch period (from Ref. 7).

Although early results, shown in Fig. 3, demonstrate a slight decrease in pitch period when breathing helium gas under laboratory conditions, this effect has been attributed to a physical contraction of the larynx muscles since the helium gas was colder than room temperature.[7] Acoustic theory would predict that no increase in pitch period should occur as a function of increased helium concentration or increased ambient pressure. Published data[11] relating the pitch period for the same utterances

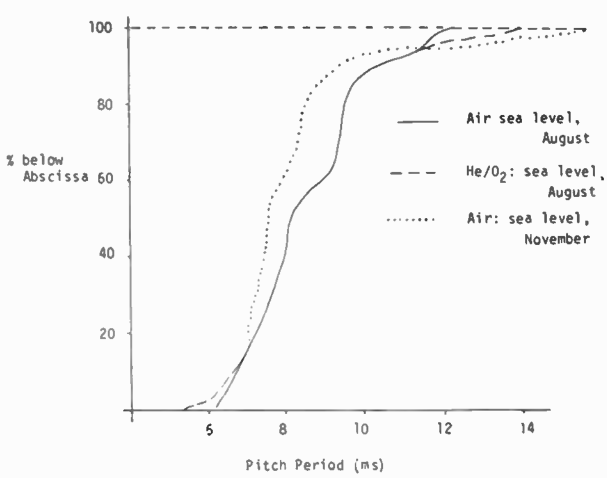


Fig. 4. Cumulative distribution of fundamental periods for air and helium/oxygen mixture at sea level (from Ref. 11).

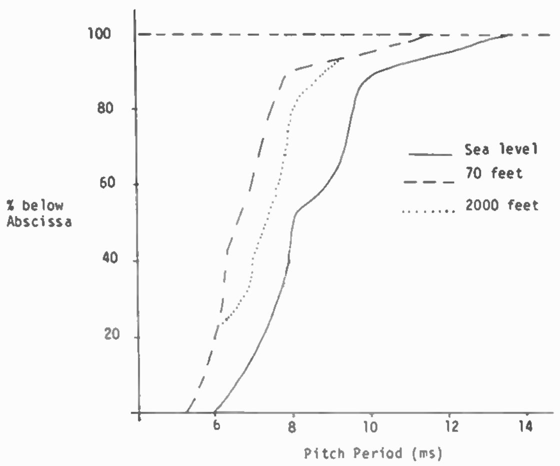


Fig. 5. Cumulative distribution of pitch period for helium/oxygen mixture vs. depth (from Ref. 11).

made in air and in a (pressured) helium environment are shown in Fig. 4 and demonstrate close correlation between the pitch distribution in air and helium at sea level on the same day. However, the pitch distribution measured some three months later for the same utterance in air at sea level shows a noticeable change. The results suggest that the effect of helium gas present in the vocal tract has no distinct effect upon pitch period, but however, the pitch distribution for a given speaker and a given utterance is subject to change with time. Results[11] shown in Fig. 5 demonstrate that pitch period in helium is reduced in comparison to the pitch period in air for the case where the helium is respired at depth. Notice here, however, that the reduction in pitch period does not vary directly with depth since the reduction is greater for 70 feet depth than for 200 feet depth, and that such changes fall within the range of expected pitch variations for normal speech in air. Several possible causes for these observed pitch changes at depth have been suggested.

First, at depth, and especially in a diving habitat, it has been observed that divers tend to speak with increased vocal intensity (loudness) in an attempt to overcome background noise levels experienced in such a chamber. Such increases in vocal intensity are normally accompanied by a reduction in pitch period. Secondly, environmental effects on the diver's speech mechanism, such as changes in the acoustic loading of the vocal tract might be expected to produce some change in pitch period. Finally, the diver may invoke modifications to his speech to alter his pitch period, in an attempt to enhance the intelligibility of speech as he himself judges.[12] Although changes in pitch period have been measured at depth, it has been shown[13] that changes of the magnitudes described in Fig. 5 have minimal effect on speech intelligibility. Pressure and gas composition appear to produce little effect on the spectrum of the vocal cord source.

5 Effects of Helium–Oxygen Mixtures on Voiced Sounds

Helium speech is best understood in relation to voiced vowel sounds. Vowel sounds are produced by excitation of the vocal tract by the vocal cords alone. The vocal

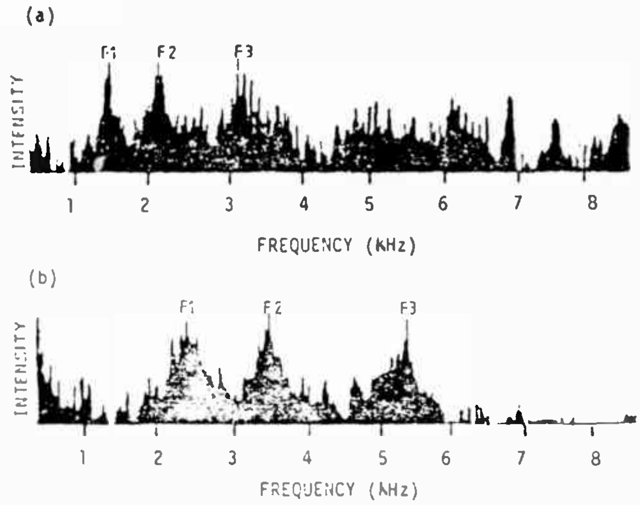


Fig. 6. Vowel spectra for speaker in (a) air and (b) helium atmosphere (from Ref. 16).

tract resonances produce peaks in the spectrum of the speech signal at frequencies determined by the positions of the movable elements (tongue, lips) of the vocal tract and a simple frequency translation in these formant frequencies might therefore be expected as a result of change in the speed of sound produced when breathing a gas of low molecular weight. Such a simple frequency translation effect applies in general; however, more detailed investigations have shown that the formant shift, in particular for the first formant, is non-linear and these non-linearities have been attributed to physiological effects produced by the change in gas density and by the increase in ambient pressure.[6,14,15] Figure 6 shows two speech spectra of the same vowel uttered by a speaker in air, (a), and in a helium environment, (b), (79% helium : 16% oxygen) at normal atmospheric pressure.[16] The frequency transposition ratios for formant centre frequencies F_1, F_2 and F_3 are approximately 1·65 : 1, 1·57 : 1, 1·56 : 1, illustrating the non-uniformity of frequency translation especially for the first formant. An increase in formant bandwidth is also

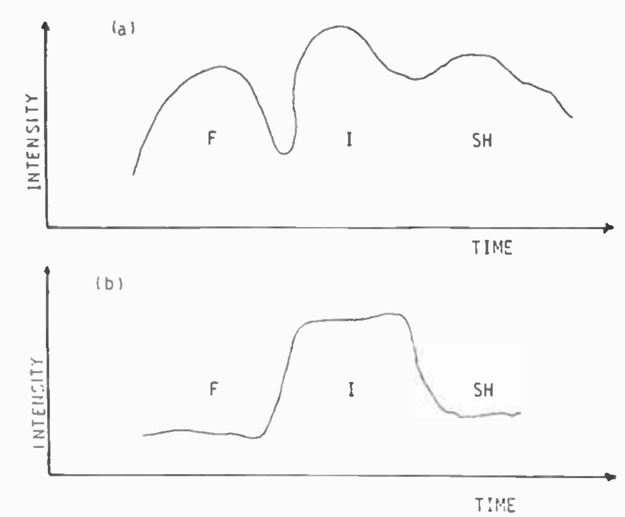


Fig. 7. The word 'fish' spoken (a) in normal air and (b) in helium at a depth of 300 ft (from Ref. 17).

apparent in Fig. 6(b) and, in particular, there is severe attenuation of high frequency components above 6 kHz. Thus, to compensate for the frequency expansions experienced in a helium–oxygen mixture, which correspond to a fall-off in excitation with frequency of 6 dB/octave, high-frequency pre-emphasis is indispensable in unscrambler design.

6 Effects of Helium–Oxygen Mixtures on Unvoiced Sounds

The attenuation of high-frequency components in helium leads to a dominance of low-frequency voiced sounds over high-frequency unvoiced sounds. This has a serious effect on the intelligibility of the speech signal since many consonant sounds in words fall into the latter category.

An example of consonant attenuation in helium speech can be seen in Fig. 7, which compares vocal intensity as a function of time for the word 'fish' taken at the surface, Fig. 7(a), and then at 300 feet, Fig. 7(b), in a helium atmosphere.[17]

7 Nasality Effects in Helium Speech

The loss of high-frequency energy and the non-linear formant frequency shifts present in helium speech can be directly attributed to an increase in the nasality of the helium speech.[14,15]

Nasality is normally due to a resonance of the nasal cavity. However, nasality can also be produced by any 'sidebranch' present in the vocal tract.[1] The main features of nasality in air at normal pressure are a general broadening of formant bandwidths, an overall loss in spectral energy, and a significant drop in intensity of the higher formants. The correlation between these features of nasality in air and the previously discussed parameters of the helium speech effect leads directly to the conclusion that nasal cavity resonance contributes to the main non-linear effects in helium speech.

In a helium environment, the likely path for transmission of energy into the nasal cavity is through the tissue of the soft palate (Fig. 1) whose acoustic impedance becomes more closely matched to that of a helium–oxygen mixture under pressure than to normal air.

8 Intelligibility of Helium Speech to the Listener

One of the most important accepted criteria for speech intelligibility is the ability to distinguish speech sounds by their relative formant frequency ratios. This is the property which allows distinction of the same sound spoken by a small boy, whose absolute formant frequencies are quite high, and a grown man, whose absolute formant frequencies will be much lower. In view of the non-linear formant shifts inherent in a pressured helium atmosphere, vowels and certain consonants involving voiced speech can therefore be expected to become more difficult to identify since their relative formant frequency ratios will not be maintained in the helium atmosphere.

Another important factor degrading intelligibility of helium speech is due to the enhancement of vowel sound intensity relative to the intensity of consonant sounds (see Fig. 7). This feature is important since consonants and the transitions between consonants and vowels provide cues to the next sound to be produced by a particular speaker.[17,18] Transitions of the first and second formant frequencies have been highlighted[19] as playing an important role in identifying the onset of, for example, nasal sounds. Hence, in the absence of such cues, the listener's perceptual system may be caught unawares and will effectively lag behind changes in the speech, and it may misinterpret and confuse one transition for another, thereby reducing intelligibility. The importance of vowel–consonant transitions has been widely reported and it has been shown that the vowel–consonant intensity ratio is an important vehicle upon which many speakers reduce articulatory effort, using instead the variation of the intensity ratio to impose a message structure.[18]

9 Self-intelligibility in a Helium Atmosphere

The problem of self-intelligibility for the diver himself is a more complex affair. It has been shown that the helium atmosphere itself affects the diver's auditory ability and provides a high-frequency emphasis of +10 dB/octave for those frequencies above 5 kHz and an attenuation for those below 5 kHz.[15] In addition to this inherent compensation in the auditory system some divers appear to adapt their voices (usually over a matter of days) in order to sound more intelligible to themselves, with varying opinions as to the success or otherwise of such manipulations with respect to the listener.[6,14]

	Vowel 1			Vowel 2			Vowel 3			Vowel 4		
	F1	F1:F2	F1:F3	F1	F1:F2	F1:F3	F1	F1:F2	F1:F3	F1	F1:F2	F1:F3
Early in experiment	1000	1.9	3.25	1100	1.91	3.27	950	2.3	3.3	1450	1.93	2.97
Late in experiment	900	2	3.28	950	2	3.47	800	2.6	3.56	1100	2.09	3.36
In air	550	3.09	4.45	600	3	4.17	400	3	5.75	700	2	3.36

Fig. 8. Self-adaptation of formant ratios in a helium environment towards normal (air) values.

Formant frequency ratios for four different vowel sounds uttered by the same subject are shown in Fig. 8 and suggest that relative formant ratios can be varied in a conscious manner by the diver in a helium environment.[14] The subject spent several days in a deep diving chamber and the respective vowel formant frequencies were measured in air before the experiment, in the chamber at the start of the dive and, finally, prior to leaving the chamber after several days in the helium environment.

The trends of Fig. 8 show that the first formant frequency F_1 has ultimately tended towards its value in air and that the relative formant frequency ratios ($F_1:F_2$, $F_1:F_3$) have also demonstrated a consistent trend towards their values in air.

Such observations relating to the self-compensation on the part of the diver suggest that the speech mechanism can be regarded as a closed-loop feedback system (Fig. 9). Feedback path P1 represents acoustic pathways to the ear through the surrounding environment, and paths P2 and P3 represent tactile feedback, or feedback by sensation through body tissues. The sample-and-hold unit represents short persistence memory, and stores information relating to the next sound which the speaker intends to make. This sampled information passes immediately to the feedback nodes of the error processor which analyses by how much the actual sound output differs from the expected output. It is thought that the decision element works in a predictive mode, projecting ahead to the zero-error condition and permitting input of the next speech unit on this basis.[20]

10 Helium Speech Unscrambling

The earliest attempts to unscramble helium speech involved the use of tape recorders where the helium speech was recorded at a fast speed and was subsequently played-back at a lower speed at which the resulting speech was more intelligible.[21] A modified version of this technique which enabled real-time processing involved use of a continuous tape loop on which the helium speech signal was recorded, and was subsequently read by pick-up heads which were themselves rotating past the tape loop. However, these tape methods inherently involved bulky, moving mechanisms and, more importantly, were limited in their ultimate fidelity since, in addition to shifting the formant frequencies of the speech signal, the fundamental frequency was also shifted.

The search to produce unscrambling methods which would enable correction of only the formant frequencies of the helium speech spectrum led to the development of a series of real-time helium speech unscramblers based on a variety of signal processing algorithms. These signal processing algorithms can be divided into frequency-domain and time-domain techniques. Frequency-domain techniques involve unscrambling of the speech signal by real-time frequency manipulation, usually by some form of band-pass filtering. These techniques offer the intrinsic advantage of preserving basic pitch information which is important to intelligibility. Time-domain techniques involve processing of the time-

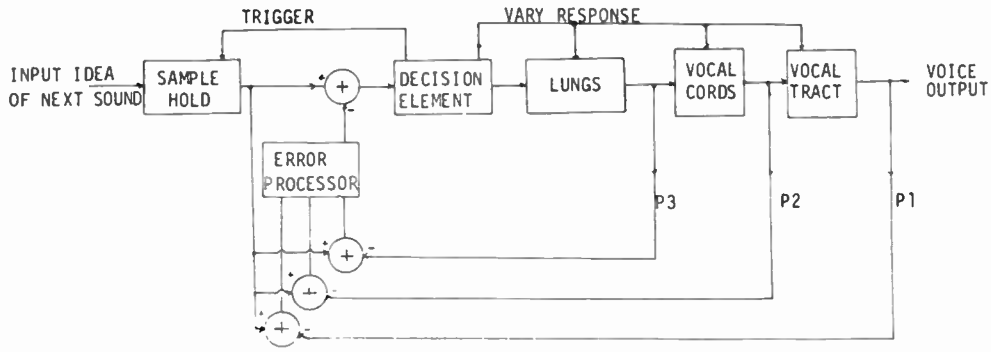


Fig. 9. The speech mechanism as a closed-loop feedback system.

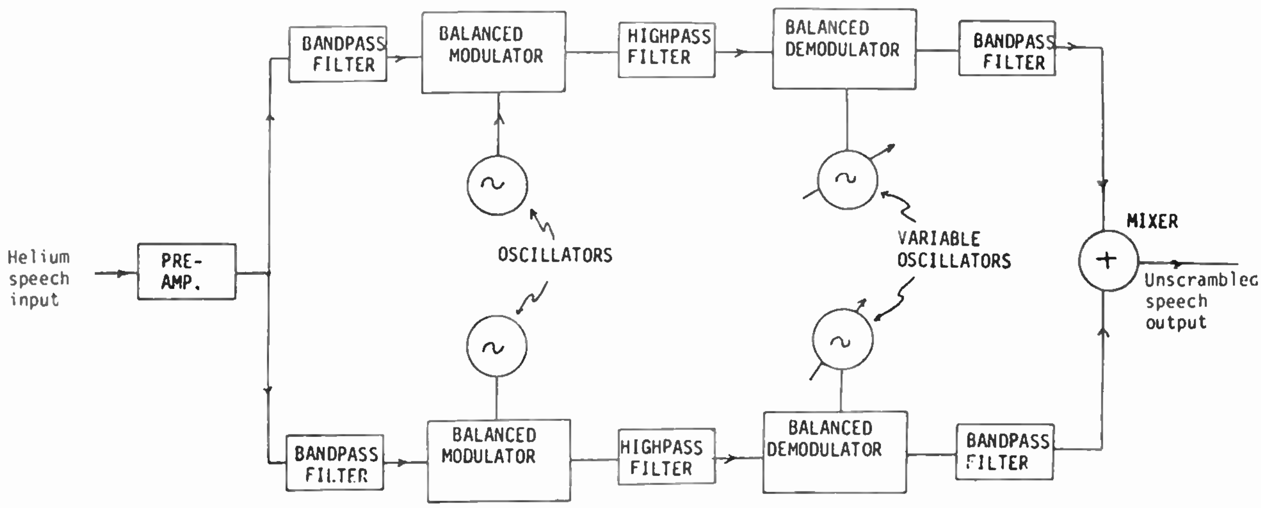
varying helium speech signal directly, the most widely used technique being to segment the helium speech signal either asynchronously or in synchronism with the pitch period, prior to subsequent time expansion of each speech segment by the required correction ratio. Time-domain techniques thus need special precautions to maintain pitch information.

Recent developments in unscrambler design involve the use of unscrambling techniques based on waveform encoding methods, which are a mixture of time and frequency domain manipulations. These encoding techniques, however, invariably involve computer processing, which is currently too slow to allow realization of a real-time unscrambler. These recent developments are reviewed here after a consideration of the three techniques used in unscrambler design, two of which can be regarded as frequency-domain techniques and the third being a time-domain processing technique.

10.1 Unscrambling by Frequency Subtraction

The first of the frequency domain techniques involves frequency subtraction by heterodyning, where the input helium speech is first converted upwards in frequency by a balanced modulator, and is subsequently down-converted by a balanced demodulator of different (and variable) frequency.[11] Use of such a variable-frequency mixing oscillator allows some form of fine tuning and hence caters for variations in helium mixture.[22]

An extension of this principle which incorporates dual-band frequency subtraction (Fig. 10), permits each band to be shifted downwards by different amounts, thereby improving intelligibility. Note that every frequency



(a)

(b)

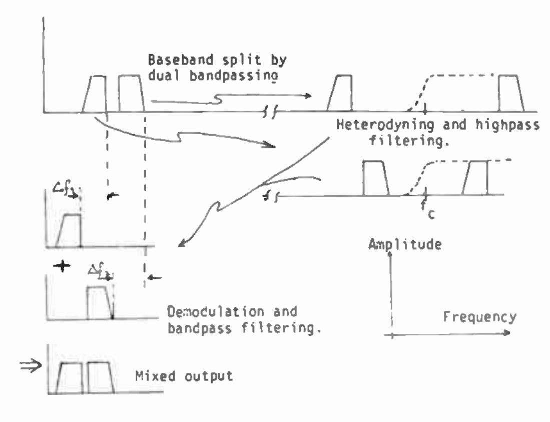


Fig. 10. (a) Helium speech unscrambler using dual-band frequency subtraction. (b) Helium speech translation by frequency subtraction.

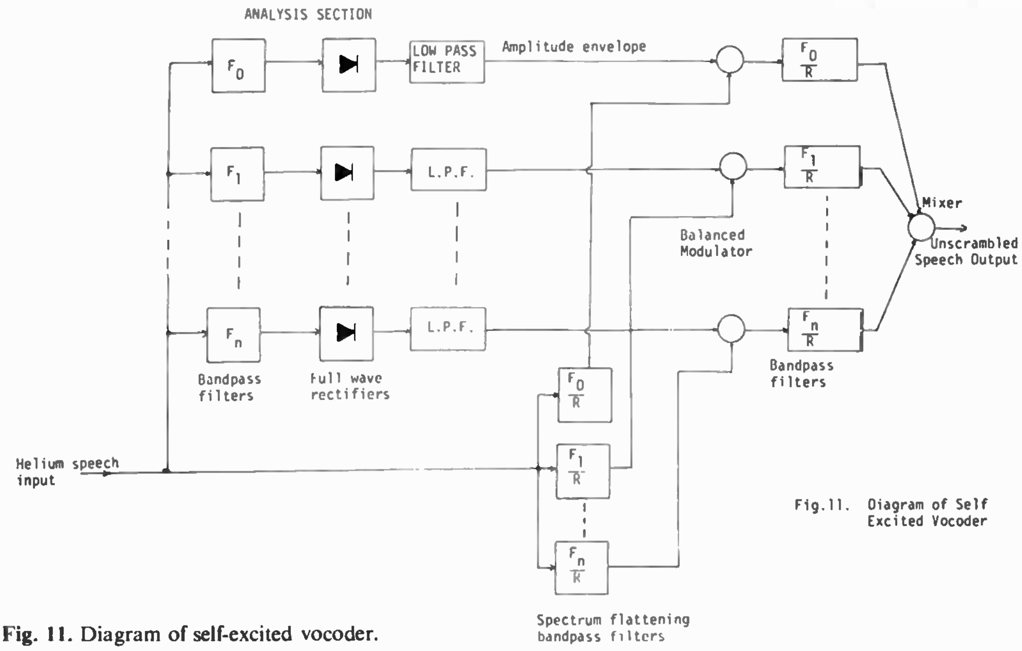


Fig. 11. Diagram of self-excited vocoder.

within the individual bands will be shifted by the same amount from its original position in the helium speech spectrum, thus the original formant bandwidths are conserved, and since helium speech is, by nature, very nasal in quality with an attendant increase in formant bandwidth, this nasal quality will be conserved in the unscrambled output, thereby limiting the fidelity of this technique.

10.2 Unscrambling by Frequency Multiplication

The second of the frequency domain techniques involves a variant of the analysis-synthesis (vocoding) scheme.[3] This technique offers unscrambling by scaling the frequency spectrum of the helium speech by ratio multiplication as opposed to the heterodyne technique, which used arithmetic subtraction. Use of the ratio multiplication technique means that formant bandwidths can be scaled by a geometric ratio and hence improved performance in terms of reduced nasality of the unscrambled helium speech can be achieved.

In general, in the analysis operation the input speech signal is decomposed into several contiguous passbands and the amplitude envelope (spectral energy) in each passband is extracted. The system employed for this is

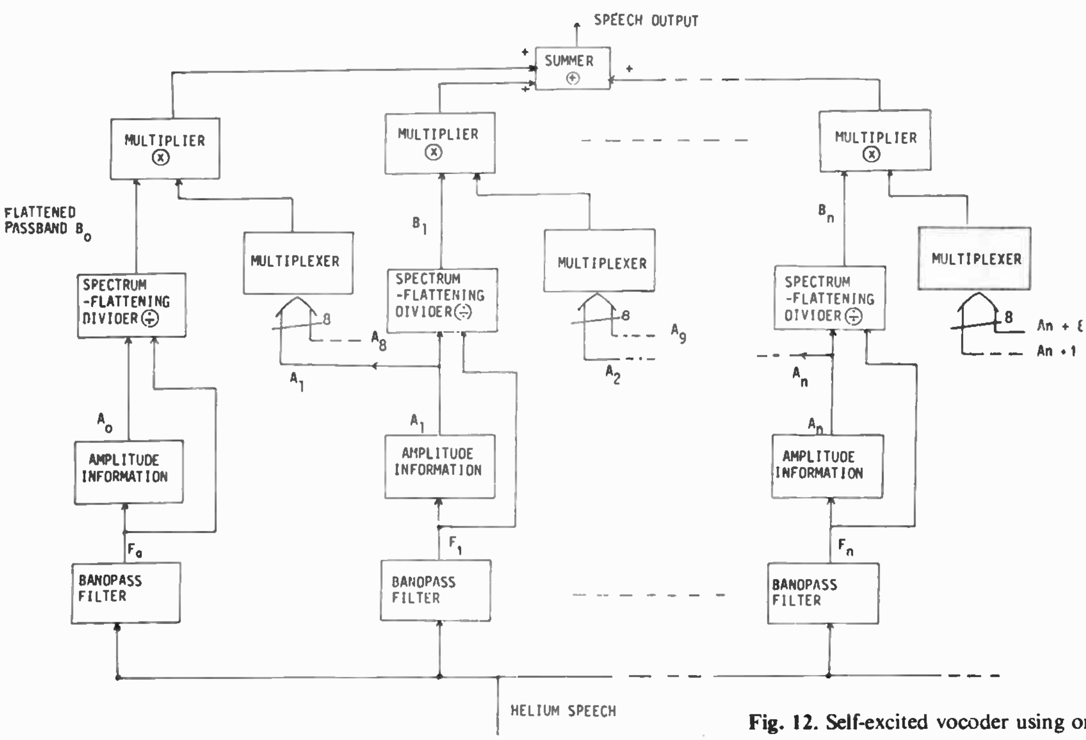


Fig. 12. Self-excited vocoder using one set of bandpass filters.

similar to that shown in the analysis section of Fig. 11. In classical vocoder design, additional circuitry exists to determine if the input speech is voiced or unvoiced and, when voiced, the fundamental pitch period is extracted. The amplitude information from each passband, the pitch information and the voiced/unvoiced decision are subsequently transmitted to a synthesizer, where a 'voiced' excitation source (driven at the pitch period) or an 'unvoiced' random signal source is connected to a second set of bandpass filters. The excitation level for each bandpass filter in the synthesizer is dictated by the energy level of the corresponding analysis filter. An extension to this analysis/synthesis scheme has been applied to the helium speech problem in the form of the Self-Excited Vocoder (SEV) system[23] (see Fig. 11). The main difference between this arrangement and the classical vocoder design outlined above lies in the excitation source used to drive the synthesis filters. Here, instead of using an artificial voiced or unvoiced source, the input helium speech is spectrally flattened (uniform spectral intensity) and is itself used as the excitation source to drive the synthesis filters, thereby preserving the natural quality of the speech. For the purpose of unscrambling helium speech, the centre frequencies of the synthesis bandpass filters and the corresponding analysis bandpass filters are related by the factor R which is defined as the ratio of the velocities of sound in the helium mixture and in air respectively. Thus for an analysis filter having centre frequency F_c, the amplitude information it derives relates to the synthesis filter whose centre frequency is equal to F_c/R. The synthesizer outputs are then summed to produce the formant-shifted speech output.

A recent, further development[24] of the SEV principle uses only one set of contiguous, bandpass filters to derive the signals required for both analysis and synthesis tasks (Fig. 12). The principle of operation depends on there being a mathematical relationship between both adjacent centre frequencies of the contiguous bandpass filters and also the velocity ratio R for certain helium–oxygen mixtures. Here, the permissible velocity ratios which the unscrambler can handle form a geometric series; as does the sequence of bandpass filter centre frequencies, both series being linked by the same geometric ratio. This means that a simple multiplexing arrangement can be employed to combine the appropriate synthesis signal (amplitude envelope) with the corresponding spectrally flattened passband (excitation signal) and hence essentially provide a shifted segment of the original helium speech spectrum.

Both the number of bandpass filters required in the system and the amount of different velocity ratios which can be catered for are directly dependant on the choice of value for the geometric ratio. A convenient value which has been chosen for this system constant[24] is 1·21.

Since the maximum velocity ratio possible is ≈ 3 (100% helium), then the system may cope with six different helium–oxygen mixtures in the range 1·21–3, ($1{\cdot}21^6 \approx 3$). Also, twenty-two contiguous bandpass filters are required with this choice of system constant in order to unscramble helium speech successfully in the range 250 Hz–16·5 kHz. ($250\ \text{Hz} \times 1{\cdot}21^{22} = 16{\cdot}5\ \text{kHz}$).

The main advantages of this system are that it provides a selectable range of helium–oxygen mixtures over which it will operate and, since it uses only one set of bandpass filters to provide analysis and synthesis signals, the number of electronic components is reduced thereby reducing the cost of the system. The very nature of operation of this system, however, precludes the ability to correct for non-linear formant shifting in the speech spectrum, although the possibility of correcting for high-frequency attenuation exists by selective manipulation of gain within passbands.

10.3 Unscrambling by Time Expansion

The most widely used real-time signal processing technique for helium speech unscrambling involves time expansion of pitch-synchronous segments of the speech signal. The precursor to this technique involved the use of a tape recorder, where playback of the recorded helium speech at half-speed was reported to increase intelligibility.[21] However, with this tape recorder method, fundamental pitch period is also shifted, and the technique is difficult to implement in real time. The time expansion process involves segmenting the helium speech signal in time using the start of each pitch period as the segment marker. The signal within a segment is sampled at a fast rate commensurate with Nyquist requirements; the signal is then stored and subsequently read out of the storage register at a slower speed. There are many available means of determining the start of a pitch period.[25–30]

Several unscramblers have been designed which employ signal processing techniques based on bandwidth compression by simple waveform time-base expansion in pitch synchronism with each pitch period.[31–33] Such signal processing demands use of electronic storage and systems have tended to employ digital memory components for this function. However, developments in analogue memory components now permit the realization of unscramblers which employ the basic pitch-synchronous time-expansion technique but are based on analogue charge transfer devices for waveform storage and c.m.o.s. digital circuitry for control logic functions. Use of low-power analogue technology in the unscrambler permits realization of a compact unscrambler system which can operate in an underwater environment.[34,35]

A schematic block diagram of a helium speech unscrambler based on the time-expansion technique is shown in Fig. 13. After high-frequency equalization in

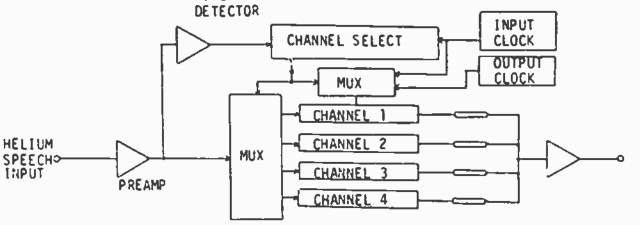


Fig. 13. Diagram of helium speech unscrambler using the pitch synchronous time expansion technique.

the pre-amplifier, the helium speech signal is input to the data stores.

From acoustic theory, the speed of sound in a gas is given by:

$$C = (\gamma P/\rho)^{\frac{1}{2}}$$

where γ is the adiabatic constant (ratio of specific heats), P is the gas pressure and ρ is the density of the gas. The adiabatic constants and the densities can be calculated using the following relationships:

$$\gamma = \sum_i Q_i \cdot \gamma_i$$

$$\rho = \sum_i Q_i \cdot \rho_i$$

where Q_i is the percentage of the ith gas in a mixture. For air the main constituents are nitrogen and oxygen in the proportions 78% : 21% by volume. A measure of the worst-case condition can be made by considering a pure helium environment in which case the velocity ratio is given by:

$$\frac{C_{\mathrm{He}}}{C_{\mathrm{Air}}} = \left(\frac{\gamma_{\mathrm{He}}}{\gamma_{\mathrm{Air}}} \times \frac{\rho_{\mathrm{Air}}}{\rho_{\mathrm{He}}}\right)^{\frac{1}{2}}$$

$$= \left(\frac{1 \cdot 66 \times 1 \cdot 27}{1 \cdot 40 \times 0 \cdot 179}\right)^{\frac{1}{2}} = 2 \cdot 9.$$

Since the maximum time expansion required on the inter-pitch stored waveform is 3 : 1, four independent storage channels are required so that, in the worst case, three channels can be reading signal out whilst the fourth channel is available for reading signal in, thus preventing loss of any pitch intervals.

The pitch detector used in available unscramblers is typically an amplitude peak detection circuit with hysteresis. The increased speed of sound in the helium/oxygen mixture—in comparison to air—produces a faster decay time-constant for the helium speech waveform envelope, forcing the pitch peaks to become more pronounced than in air, and simplifying the pitch detection circuitry so that a peak detector circuit can be successfully employed. With unvoiced speech, characterized by a noise-like waveform, the pitch detector operates repetitively.

The output from the peak detector circuit defines the start of a pitch period and a segment of the input helium speech, of bandwidth up to 16 kHz, is stored in one of the four channels at a fast sample rate. At the end of the frame, the clock frequency for this store is reduced by a factor which is dependent on the gas mixture being used by the diver. The stored signal is then read out at a lower rate with a bandwidth compression to the normal 3–4 kHz speech bandwidth. On detection of the start of the subsequent pitch period, the clock and signal multiplexers change and another store reads in the helium speech. Typical stored segments are in the range 2·5–20 ms permitting operation with diver fundamental frequencies up to around 400 Hz. This is acceptable for most male voices, whose pitch frequency will on average vary from 70 Hz to 150 Hz in normal speech, and for most female voices which will vary from 100 Hz to 400 Hz.

More than one channel may be producing an output at any one time, a feature which is acceptable since, in normal speech, the vocal tract response due to a glottal pulse has not died away before the next response appears.

A section of input helium speech waveform (85 metres depth) is shown in Fig. 14(a), with Fig. 14(b) indicating the pitch detector output defining a 3 ms storage interval from the pitch peak. The loss of signal in discarded sections of each pitch period causes minimal degradation in the output speech. The pitch period in Fig. 14(a) is seen to be of the order of 6·5 ms. Another section of input helium speech waveform (Fig. 15(a), 85 metres depth) is compared with the corresponding unscrambled output (Fig. 15(b)). Note that the pitch interval (approx. 6 ms) remains unchanged although the stored signal is expanded in time.

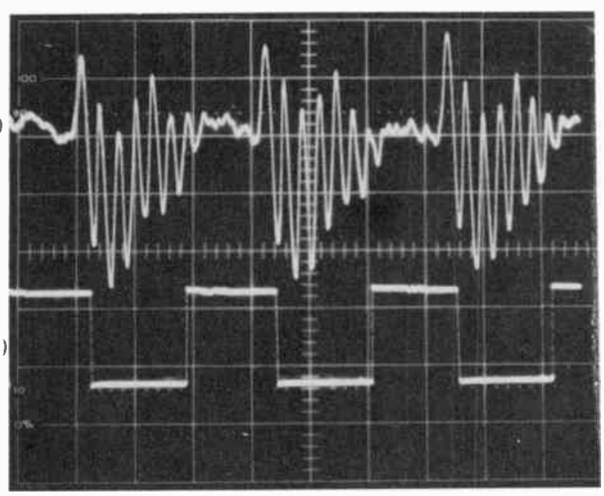
(a)
(b)


Fig. 14. Pitch detector operation showing 3 ms frame. (a) Helium speech waveform 20 mV/div. (b) Pitch detector output 10 V/div horizontal scale 1 ms/div.

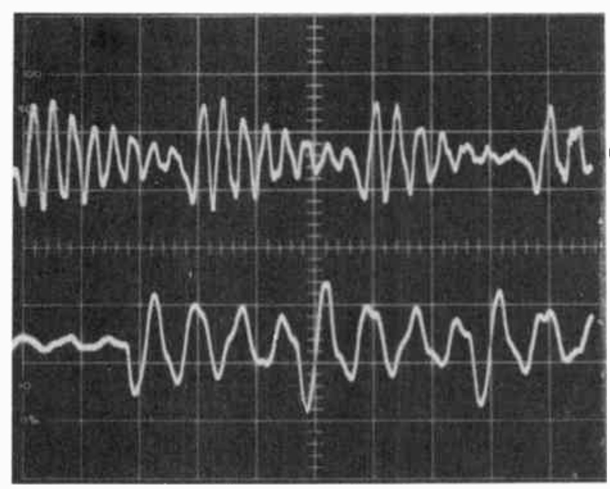

(a)
(b)

Fig. 15. Unscramble output. (a) Helium speech input waveform 50 mV/div. (b) Final unscrambler output 1 V/cm horizontal scale 2 ms/div.

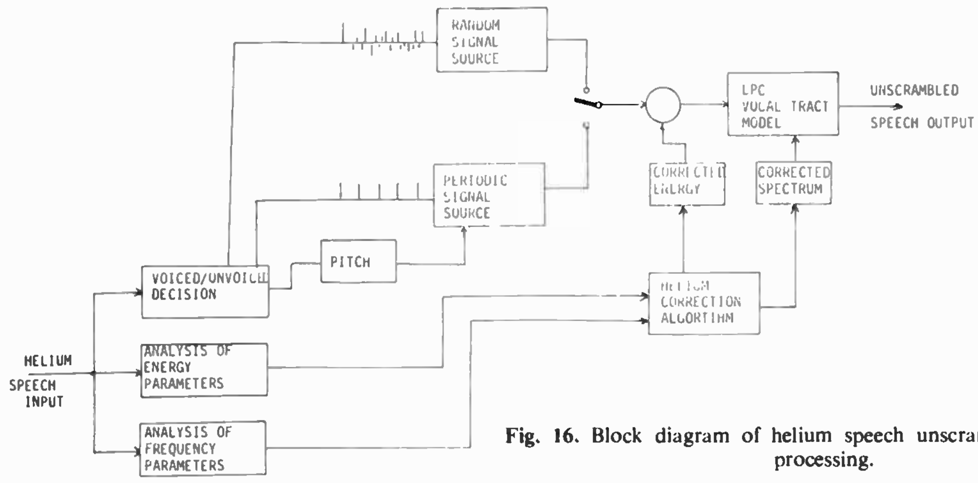


Fig. 16. Block diagram of helium speech unscrambler using LPC processing.

10.4 Unscrambling Techniques Based on Waveform Encoding

In addition to the previously described techniques which have all been implemented practically, linear predictive coding[36] and homomorphic deconvolution[37] have also been applied to the problem of helium speech unscrambling, although neither has yet been implemented practically in a real-time unscrambler. Although lengthy and time-consuming algorithms are at present necessary for their implementation, these techniques offer the potential to perform non-linear frequency shifting for future high performance unscrambling systems.[38]

10.4.1 *Unscrambling by linear predictive coding*

In linear predictive coding (LPC), the speech signal is sampled and analysed to yield data such as pitch period, voiced or unvoiced decision, voice intensity and vocal tract response parameters. The latter parameters can be used to control the response of a time-variant, multi-stage, recursive, digital filter which is excited by a voiced/unvoiced excitation source (Fig. 16). The filter represents a model of the vocal tract, and is capable of predicting the current speech sample based on a weighted linear combination of previous speech samples. The system works on the assumption that the vocal tract characteristics remain constant over a 20–25 ms period so that new filter values from the analysis section are computed and corrected according to some helium–air correction algorithm every 20 ms.

Although computer-simulated unscrambling using LPC analysis has been evaluated,[39] no real-time device has yet been used for helium unscrambling based on this technique, and the successful use of any such LPC system will depend heavily upon the ability to model successfully the helium speech effect in order to apply dynamic (non-linear) correction to the filter coefficients.

10.4.2 *Unscrambling by homomorphic deconvolution*

The speech signal is the result of a convolution operation between the glottal air excitation injected through the vocal cords into the vocal tract and the impulse response of the vocal tract. Signal processing by homomorphic deconvolution[37] involves a transformation which yields a function known as the cepstrum. The cepstrum contains information relating to fundamental pitch period and frequency/amplitude information for the speech signal. If the cepstrum of the glottal air excitation itself is then known or can be approximated, it can be subtracted directly to yield a modified cepstrum which contains only information relating to the deconvolved vocal tract response. The vocal tract response is assumed constant over the 20 ms analysis period of the cepstrum computation. Figure 17 shows the homomorphic (cepstrum) process applicable to this situation.

The required non-linear frequency processing to achieve unscrambling of the helium speech can be applied in computation of the inverse cepstrum.[40] The resulting output of the homomorphic analysis system is

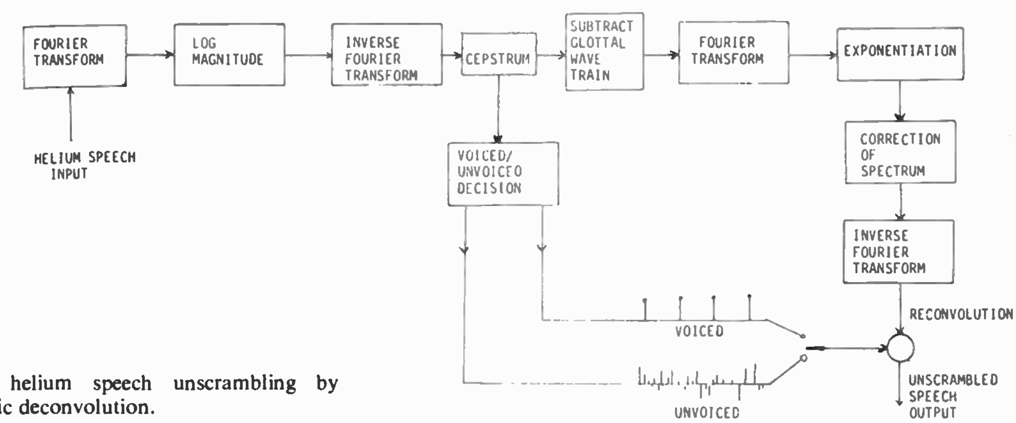


Fig. 17. Block diagram of helium speech unscrambling by homomorphic deconvolution.

then reconvolved with a periodic pitch impulse source or unvoiced random signal source. This proposed technique is ultimately limited by the requirement to compute Fourier transformations in real time, even when using fast Fourier transform (FFT) techniques.[41,42] In addition, the system assumes prior knowledge of the glottal air excitation waveform in helium to permit deconvolution at the cepstrum stage. Thus the derived vocal tract impulse response may be corrupted by remnants of the incorrectly deconvolved glottal air excitation function, thereby limiting the performance of the technique.

11 Factors Degrading the Intelligibility of Unscrambled Helium Speech

Many of the above unscrambling techniques involve some form of voiced/unvoiced sound detection based on a pitch extraction algorithm whereby if a pitch cycle is detected, the sound is assumed voiced, if not, it is assumed unvoiced. On the basis of this judgement the time domain 'expanded segment' principle stores each pitch-synchronous segment on detection of a pitch cycle. In the absence of a pitch decision, the unscrambler segments the speech in a repetitive fashion, or alternatively, may output white noise. In the waveform coding techniques of linear predictive coding and homomorphic deconvolution, either a pitch-synchronous impulse train or a random signal source is used to resynthesize the speech. This strategy tends to form an important source of degradation of intelligibility, since the air excitation waveform injected through the vocal cords, although up to this point assumed to be a perfect Dirac impulse, is in fact a complex waveform.[43]

In view of the fact that pitch detection is carried out on the emitted helium speech signal (which is the convolution of the vocal cord excitation and vocal tract impulse response) and especially since the diver may be working a high ambient noise environment, it seems reasonable to assume that there may be a significant probability of pitch extraction errors in helium speech unscrambler operation.[29]

In analysis/synthesis unscrambler systems, an unvoiced decision causes a random noise source to be used as the synthesis excitation source and, for the pitch-synchronous time-expansion unscrambler, the sampled and expanded waveform loses pitch synchronism for such an unvoiced decision. Assuming that such unvoiced decision errors occur amidst true voiced speech, these errors can be viewed as 'noise' intervals, as far as the human ear is concerned. Figure 18 shows results of the effects on intelligibility of alternating intervals of equal length of speech and white noise.[44] The intelligibility is measured in terms of the percentage of words heard correctly, and is plotted against frequency of interruption. Interruption frequencies ranged from 0·1 to 10 000 per second, and the signal-to-noise ratio was varied from −18 to +9 dB. Shown also is the response, marked 'Quiet', when the intervals between speech were silent, with no added noise.

It can be seen from Fig. 18, that, in the range 10 to 560 noise interruptions per second, intelligibility is increasingly impaired. The pitch frequencies of most male and female speakers fall within this range; hence if a succession of unvoiced pitch decisions are made in error of actual voiced speech sounds, then that sound will be effectively masked in noise as the pitch detector causes unvoiced sound to be output. In addition to masking the actual voiced speech sound, it is conceivable that vowel-consonant transitions, which have already been identified as playing an important role in intelligibility,[17,18] may be degraded.

Another possible source of degradation of intelligibility and loss of voice quality can be appreciated from the facts, stated earlier, that in a helium atmosphere the frequency spectrum is shifted in a non-linear fashion and that, due to amplitude spectrum distortion, voiced sounds are enhanced at the expense of unvoiced sounds (high frequency energy is heavily attenuated). All of the unscramblers developed to date which are based solely on time or frequency domain techniques involve a shift of the frequency spectrum in a piecewise linear fashion and make no attempt to correct amplitude spectrum distortion. However, since it has been shown that vowels are recognized by their formant frequency ratios, and that speech intelligibility is a function of vowel–consonant intensity ratio fluctuations, then a correlation exists between these facts and loss of intelligibility when using present unscramblers in that formant frequency ratios may not be conserved by simple linear frequency transformations, and since no amplitude compensation is applied, vowel–consonant ratios are adversely affected. Hence, although the technology used in present-day unscramblers continues to advance in terms of reduction of unscrambler size and minimization of power consumption, device performance continues to be non-optimal in terms of intelligibility.[45,46]

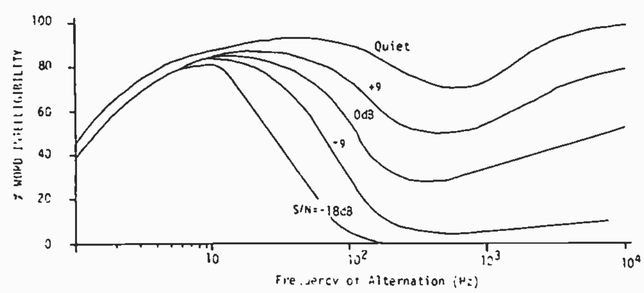


Fig. 18. Word intelligibility as a function of the frequency of alternation between speech and noise, with signal-to-noise-ratio in dB as the parameter (from Ref. 44).

12 Conclusions

The mechanisms affecting speech intelligibility in normal air have been identified. These include the dependence of vowel recognition on the relative formant frequency ratios, and the recognition of nasality by the attenuation of formant amplitudes and the broadening of formant bandwidths. Transitions of formant frequencies and the fluctuation of vowel–consonant intensity ratios have likewise been identified as playing an important role in speech intelligibility.

Similarly, the effects of a helium atmosphere upon the speech spectrum have been discussed. These include a non-linear shift of the frequency spectrum, particularly at low frequencies, which destroys relative formant frequency ratios, thereby confusing the recognition of vowel sounds; broadening of formant bandwidths due to alterations in the acoustic impedance of the vocal tract, hence causing the characteristic nasal quality of helium speech; and severe attenuation of high-frequency sound, thereby affecting vowel–consonant intensity relationships which are important in providing transitional cues from one sound to another.

Of the various unscrambling techniques reviewed in this paper, those techniques which involve frequency-domain processing either as the main source of unscrambling or as part of some composite strategy, as in waveform coding techniques, offer the potential of providing the non-linear shift necessary for helium speech unscrambling. Here, segmentation of the helium speech spectrum and subsequent piecewise relocation of each passband within the normal speech frequency range restores formant frequencies to their original areas of the spectrum. However, correction of individual formant bandwidths, and hence successful denasalization of the helium speech, is not necessarily implicit. Frequency domain techniques in which unscrambling is carried out by scaling each segment of the helium speech by a geometric ratio, such as in the Self-excited Vocoder, provide the best opportunity of denasalizing speech since the formant bandwidths are also scaled. On the other hand, techniques involving frequency subtraction, such as by heterodyning, have diminished capability of correcting nasality since the helium speech formant bandwidths are conserved. Unscrambling techniques based on frequency domain processing strategies, including waveform coding techniques, offer the possibility of manipulating selectively the amplitude response of each passband to achieve correction of the high frequency attenuation present in helium speech, thus restoring the vowel–consonant intensity relationships which have been identified as having an important bearing upon speech intelligibility. None of the unscramblers developed to date, however, has incorporated any useful means by which the amplitude spectrum may be corrected for this high frequency attenuation.

Time-domain processing techniques such as pitch-synchronous segmentation and expansion are inherently unable to provide non-linear frequency shifting and lack the capability to fully-correct the amplitude spectrum of helium speech. Notwithstanding the limitations of the basic technique, however, this type of unscrambler is currently in widespread use, due mainly to its simplicity, low cost and reliability.

Future developments in unscrambler design appear to be evolving on two major fronts. First, advances in implementation technology have already led to a miniaturization of the time-domain technique into a compact form for diver-borne use, with a view to further development to single-chip form, opening up the possibility for diver-to-diver and diver-to-surface through-water communications. Secondly, improvements in unscrambler system design, based on frequency domain techniques, will lead to a more faithful reproduction of the diver's voice. Techniques which use frequency-domain processing are the only options with the capability of including all the features necessary for good intelligibility of the unscrambled helium speech.

Thus, over the next few years, both new technological implementations and new systems developments can be expected in helium speech unscrambling. These new systems will result in a new generation of unscramblers, with high quality performance in terms of intelligibility of the unscrambled helium speech, increased diving efficiency and above all, increased diver safety.

13 Acknowledgments

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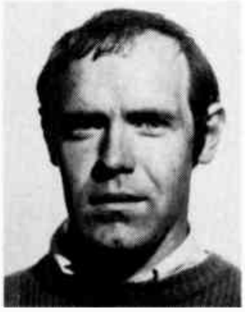
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Indexing Terms: Circuit design, logic, Integrated circuits

Optimizing gate interconnections in four-phase dynamic logic m.o.s. l.s.i. technology

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SUMMARY

Dynamic logic circuit technology has been used widely to implement large-scale integrated (l.s.i.) circuits using metal–oxide–silicon field effect transistors (m.o.s.f.e.t.). One of the factors which determines the overall dimensions of a custom-designed random logic l.s.i. circuit is the number of interconnection tracks as they occupy a large part of the chip between functional modules. This paper describes a multiplexing technique, which allows a reduction in the number of interconnection tracks between modules in l.s.i. circuits implemented using the four-phase dynamic logic technology of the major–minor configuration. It is shown that the operating speed or performance of the circuit is not affected by this technique.

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1 Introduction

With the advent of large-scale integrated circuits, it is feasible to implement complex functions on a single silicon chip. A typical l.s.i. circuit may incorporate several hundred gates or several thousand transistors. Advances in semiconductor fabrication technology and in mask making techniques have led to a reduction in the geometrical dimensions of each individual transistor, thereby increasing the packing density and the circuit complexity.

When realizing l.s.i. random logic circuits the interconnection tracks between gates are an important factor in determining the chip size. A typical circuit has data highways containing many lines and each line runs over a significant length. Interconnection tracks occupy significantly more area of the chip when compared with the area occupied by the active transistors. Consequently, the overall packing density of gates per unit area is less with random circuits than with regular circuits such as memories. This paper describes a multiplexing technique which reduces the number of lines in the data highways for customized m.o.s.l.s.i. circuits, using the major–minor configuration of the four-phase dynamic logic circuit implementation.[1] The operating speed or the performance of the circuit is not affected in any way although there is a reduction in the chip dimension and hence an increase in the overall packing density in terms of gates per unit area of the chip. The advantages of four-phase dynamic logic technology are that the power dissipation is reduced as there is no d.c. path between the supply and ground, the circuit may be implemented using the minimum size device thereby increasing the packing density, and several functions such as shift-register can be implemented using fewer devices. Because clocks are used to synchronize the operation of the circuit, the design is free of race hazards.

2 Optimization of Data Highway

To facilitate the implementation of an l.s.i. circuit, the designer partitions the logic into functional modules. Those with many interconnection wires are located adjacent to each other to minimize the chip area. Because of the very nature of the custom-designed integrated circuits, the layout is complex and random; even after optimum partitioning, a significant chip area is occupied by data highways containing many interconnected tracks. Modules which have a smaller number of interconnection tracks may be separated topologically by large distances and each interconnection wire increases the chip dimensions. Figure 1 shows part of a

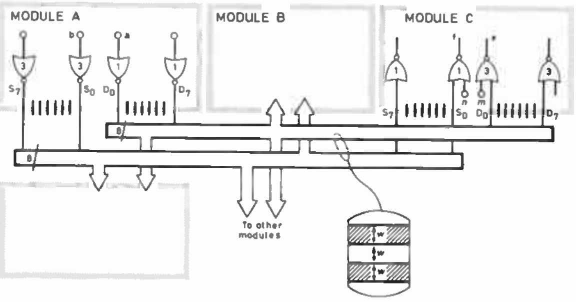

Fig. 1. Chip partitioning. Modules and data organization.

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chip with 16 lines in the data highway, corresponding to eight data signals (D_0–D_7) and eight control signals (S_0–S_7), being shared between several functional modules. The circuit is designed so that each of the data signals forms an output of a gate type 1 and each of the control signals an output of a gate type 3 to make the optimum use of the technique to be described in Section 2.1. Assuming that the minimum width of a track and the minimum separation between adjacent tracks is w microns and the total length of each line is l microns, the area occupied by tracks S_0 and D_0 would be $3lw$ square microns. Figure 2 shows the same circuit in which two signals S_0 and D_0 are multiplexed and transmitted over a single interconnection track. The area occupied by the single interconnection wire is lw square microns, resulting in a saving of $2lw$ square microns. Signals S_0 and D_0 are demultiplexed in each of the functional modules by introducing decoupling devices as shown in Fig. 2. Similarly, other data and control signals are paired to half the number of lines and signals demultiplexed in each module. It should be noted that the area occupied by the decoupling devices is negligible.

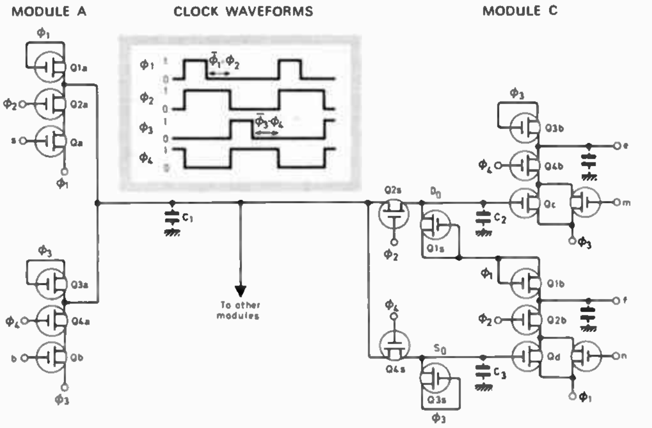

Fig. 2. Multiplexing circuit.

2.1 Circuit Operation

Figure 2 of the paper is a schematic diagram of the circuit. When it is laid out, it becomes apparent that the decoupled output, as it forms a gate of an input, is available in metallization for probing and fault-finding if necessary. The waveform, of the decoupled output has one-to-one correspondence with the waveform of an output where the optimization technique is not used. After familiarization, the complex waveform on the multiplexed line is easy to read.

The detailed operation of the circuit shown in Fig. 2 with respect to the clock waveforms is as follows.

Let inputs a and b be at logic levels 1 and 0 respectively. Devices Qa and Qb are in the on and off state respectively. During ϕ_1 period, devices Q1a and Q1s are turned on and capacitances C1 and C2 charge up to a logic level 1 (precharge). In the sampling period $\bar{\phi}_1 . \phi_2$ devices Q2a, Q2s and Qa are on and both C1 and C2 discharge to a logic level 0. As ϕ_3 and ϕ_4 are at a logic level 0, devices Q3a, Q4a and Q4s are off, there is no discharge path to ground and the state of C3 is not affected. When ϕ_2 goes to a logic level 0, device Q2s is turned off and C2 is isolated from C1.

During ϕ_3 period, devices Q3a and Q3s are switched on and C3 and C1 precharge to a logic level 1. During the sampling period $\bar{\phi}_3 . \phi_4$, devices Q4a and Q4s are on; but there is no discharge path for C1 and C2 as device Qb is off. Hence capacitances C1 and C2 remain charged at a logic level 1. As ϕ_1 and ϕ_2 are at a logic level 0, device Q2s, Q2a and Q1a are off and do not affect the charge on C3. When ϕ_4 goes to a logic level 0, the device Q4s is turned off and C3 is isolated.

When both a and b are at a logic level 1, devices Qa and Qb are on. During ϕ_1 period C2 is precharged to a logic level 1; but will discharge to logic level 0 via devices Q2a, Q2s and Qa during the sampling period $\bar{\phi}_1 . \phi_2$. When ϕ_2 goes to a logic level 0, C2 is isolated. Similarly during ϕ_3 period C3 will precharge to logic level 1; but will discharge to logic level 0 via devices Q4a, Q4s and Qb during $\bar{\phi}_3 . \phi_4$ period. When ϕ_4 goes to a logic level 0, C3 is isolated.

When both a and b are at a logic level 0, devices Qa and Qb are off. C2 and C3 will precharge to logic level 1 during ϕ_1 and ϕ_3, respectively. However, there is no discharge path to ground during the corresponding sampling periods as devices Qa and Qb are off and C2 and C3 remain charged at a logic level 1.

When a is at a logic level 0 and b at 1, device Qa and Qb are off and on, respectively. During ϕ_1 period C2 will charge up to a logic level 1 and retain the charge as no discharge path exists during $\bar{\phi}_1 . \phi_2$ period. During ϕ_3 period C3 will charge up to a logic level 1; but will discharge to a logic level 0 via Q4a, Q4s and Qb during $\bar{\phi}_3 . \phi_4$ period.

Thus the state of capacitances C2 and C3 is the inverse of logic at inputs a and b, respectively, and the signals are demultiplexed.

Two other advantages of the technique are as follows:

(a) In four-phase dynamic logic circuits faults which could result in an incorrect operation of the circuit and be directly attributed to the layout design are as follows:

- (i) an error in the interconnection pattern or violation of the layout design rules,
- (ii) an error in design of the device size, or
- (iii) the effect of voltage coupling due to certain harmful interconnection crossovers causing a gate to crash.

Of the above faults, error types (i) and (ii) could be located on chip by monitoring the signal propagation with the aid of probes and oscilloscope. This method is not applicable when the incorrect operation is caused by the error type (iii) as the circuit capacitance and coupled voltage are altered when probes are used. Error type (iii) occurs as a result of the cumulative effect of voltage transitions on outputs of certain gates coupled into the input of a gate by interconnection crossover capacitances. It is difficult to locate because of the dynamic and unpredictable nature.

By introducing the decoupling devices, only C2 and C3, which are local to the module, form the relevant inputs to the gate. The layout design is eased as the problem of harmful coupling voltages inherently associated with a long run of interconnections and other crossovers with it is eliminated.

(b) The circuit may be improved slightly by removing devices Q1a and Q3a. This results in a small saving in dissipated power. The dissipated power is proportional to CV^2 for each charge/discharge cycle. The peak voltage of C1 with Q1a is approximately $(V-V_t)$ where V is the peak ϕ_1 voltage and V_t is the threshold voltage of the device. Without Q1a the peak voltage of C1 is approximately $(V-2V_t)$ as the peak ϕ_2 voltage when derived on chip is $(V-V_t)$. The approximate saving in power consumption per cycle is given by $C_1(2VV_t+3V_t^2)$.

3 Conclusion

The reduction in the device size and improved circuit techniques have enabled the doubling of the complexity of integrated circuits every year. This trend will slow down as an ever-decreasing size will lead to the fundamental problems in the device physics and fabrication.[2] It will be difficult to continue to discover new circuit techniques. This paper describes a useful optimizing technique at the interconnection level for reducing the chip dimensions and would lead to a significant saving depending on the randomness of the layout/logic design.

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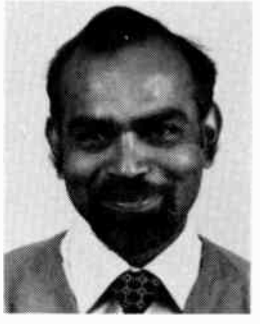

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Indexing Terms: Materials, thick film, Printed circuits

Thick-film materials for hybrids

MARTIN COLEMAN, B.Sc., M.Sc*

SUMMARY

This paper briefly describes four areas of work involved in trying to establish the reliability and limitations of thick-film materials. One of the most widely used resistor systems, Du Pont 1400, has been extensively examined and long-life stability predictions are made. The possibility of replacing gold conductors with lower cost silver alloys depends critically upon controlling or eliminating silver migration, and the conditions under which migration occurs are described. The replacement of alumina substrates with alternative materials such as enamel steel, polymer board or glass is also reported, and the behaviour of the thick film materials has been found to be generally inferior to those on alumina. The Du Pont CMS nitrogen-fireable resistor, conductor and dielectric system has been evaluated and the properties found to be comparable with air-fireable systems.

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1 Introduction

Since the early days of thick-film technology, the variety and applications of available materials have expanded beyond recognition. The two main criteria of the hybrid manufacturer continue to be a drive to reduce costs and maintain, or improve, quality and reliability. The incentive increases to replace gold by other lower cost conductor materials, and the hermetic package with some other form of passivation or encapsulation. The materials options open offer a bewildering choice, some with known potential hazards such as silver, and any change has to be based upon confidence that circuit reliability will be maintained. Thorough investigations have to be undertaken not only into the likely life behaviour of new materials, but also into identifying the failure mechanisms involved. On the basis of their determined performance, a direct comparison can then be made with the established system and the test options chosen. Some of the areas examined during the past few years at STL are described briefly in this paper.

2 Tight T.C.R. Resistor Systems

The first resistor systems were based upon grains of palladium oxide and silver-palladium alloy in lead borosilicate glasses.[1] The temperature coefficient of resistance was typically about 300 parts/10^6 deg^{-1}C, the resistors were firing sensitive and the long-life stability was poor, especially in the presence of moisture.[2]

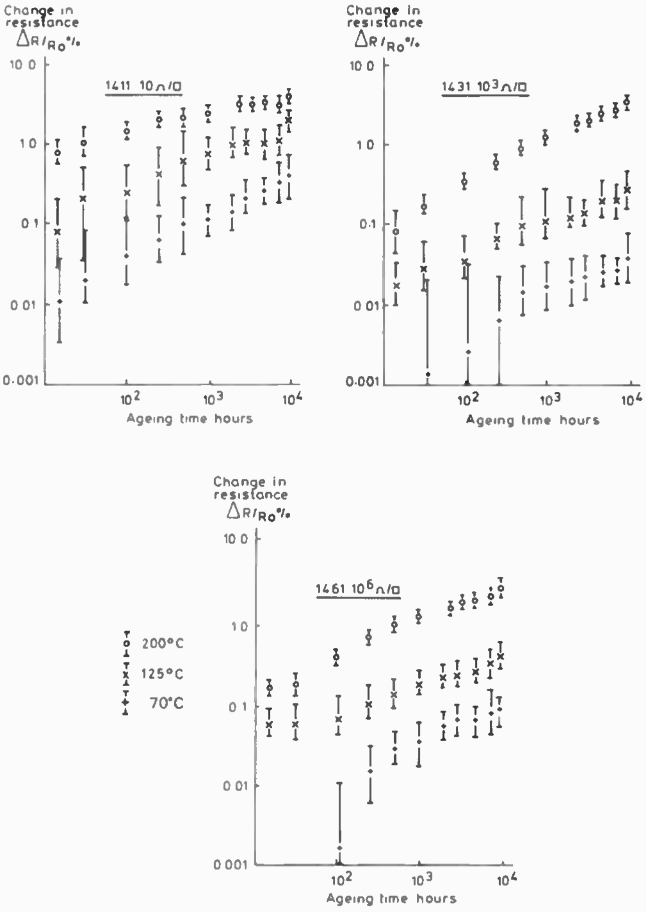

Fig. 1. Change in resistance with time of DP 1400 after ageing.

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A second generation of resistor systems was developed during the early 1970s based upon ruthenium and iridium oxides. About the same time, Du Pont produced the first of their 'Birox' series resistor systems, based upon bismuth ruthenate in lead borosilicate glasses, which were more stable, had tighter t.c.r.s and were less sensitive to firing conditions.[3,4]

Typical of the better available resistor series is the Du Pont 1400 system. The range of resistivities offered is from 10 Ω/□ (1411) to 1 MΩ/□ (1461) with a claimed temperature coefficient of resistance (t.c.r.) of less than ± 100 parts/10^6 deg^{-1}C from -55°C to $+125$°C (typically <50 parts/10^6 deg^{-1}C between 0°C and 70°C). All inks are blendable and can be used with either silver-palladium or gold terminations. Quantitative analysis has confirmed that the conducting phase is bismuth ruthenate but the glass composition varies across the series and DP 1461 contains a boron-free lead silicate glass.

As part of a study of the properties of the DP 1400 resistor series at STL, three elevated temperatures, 70°C, 125°C and 200°C were chosen to induce accelerated ageing under no-load conditions in air. The logarithm of the percentage change in resistivity, $\Delta R/R_0$, was plotted against log time for each of the resistivity inks for both gold (DP 9260) and silver palladium (DP 9061) terminations. The curves for DP 1411, DP 1431 and DP 1461 are shown in Fig. 1, and the behaviour of laser-trimmed and untrimmed resistors was identical. The temperature dependence of the resistance of the DP 1400 series had semiconducting behaviour below about 300 K and metallic behaviour above this resistance minimum (Fig. 2). The amount of resistance change per degree (<100 parts/10^6 K^{-1}) of the resistors was much less than for lead ruthenate (about 3000 parts/10^6 K^{-1}) and bismuth ruthenate (about 500 parts/10^6 K^{-1}).

The changes in resistance of the DP 1400 resistors after thermal ageing were well defined and followed a

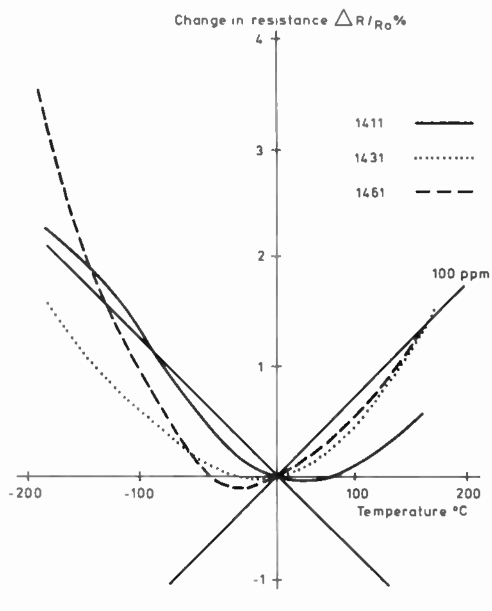


Fig. 2. Temperature dependence of resistance of DP 1400.

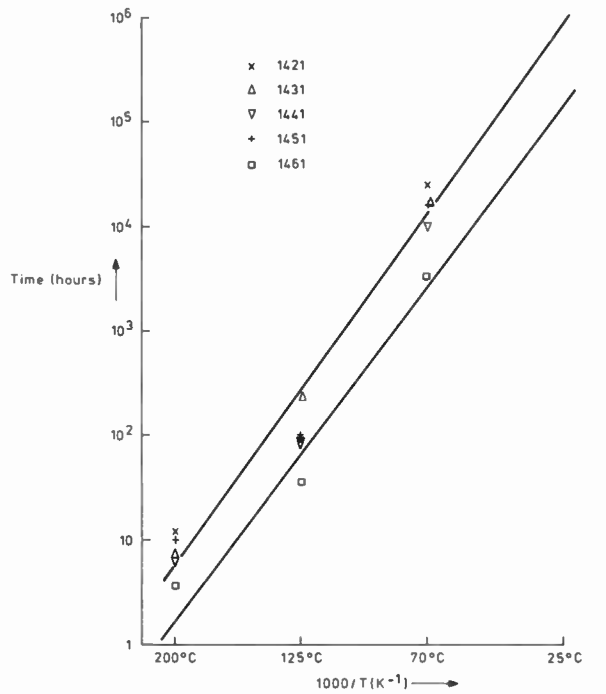
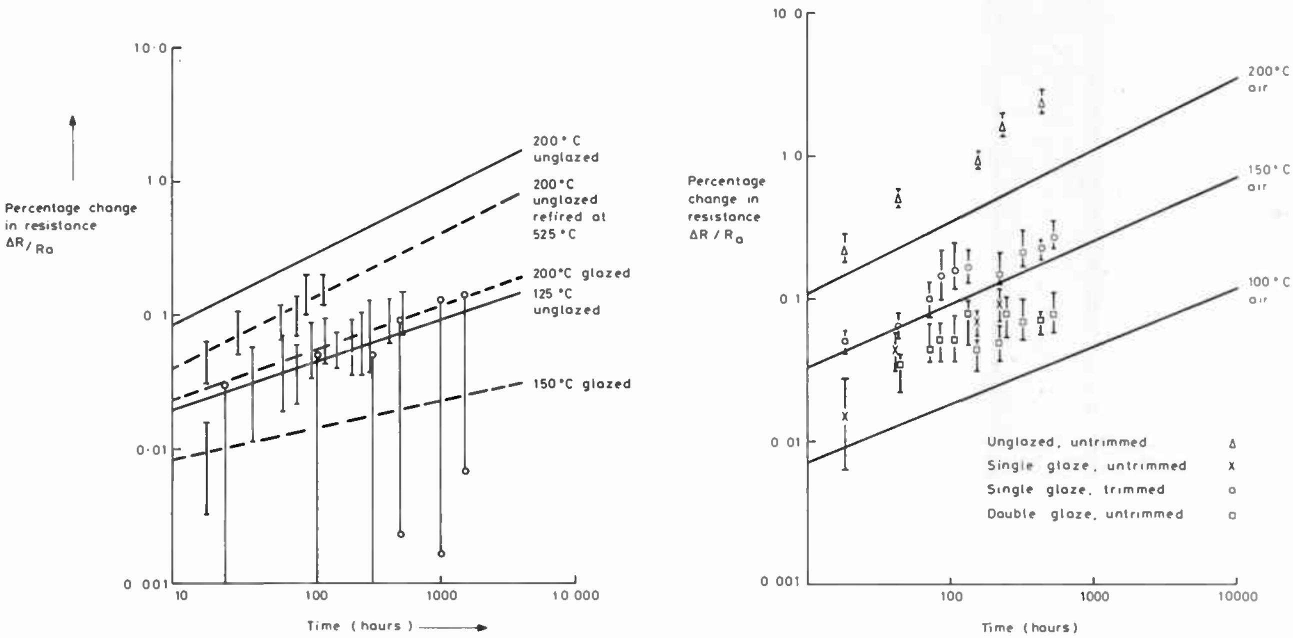


Fig. 3. Time for 0·1% change in resistance for inks of different resistances.

consistent pattern especially in the range 100 Ω/□ to 10^5 Ω/□. The time dependent behaviour varied from temperature to temperature (approximating to a square root of time dependence at 200°C and a cube root of time dependence at the lower temperatures) suggesting no single ageing mechanism. It was not possible, therefore, to extrapolate precise life behaviour from the higher temperatures to lower operational temperatures. However, the time dependence was sufficiently similar for the 100 Ω/□ to 1 MΩ/□ resistivity inks to enable an approximate Arrhenius plot to be made for the three temperatures, 70°C, 125°C and 200°C (see Fig. 3). The estimated times for 0·1% change at 25°C were about 9×10^5 hours for the 100 Ω/□ to 10^5 Ω/□ resistors and $1{\cdot}5 \times 10^5$ hours for the 10^6 Ω/□ resistors.

When the DP 1400 resistors were overglazed with DP 9137, the resistance change after thermal ageing was considerably less than for the unglazed resistors and the slope at 200°C was reduced (Fig. 4(a)). Refiring through the overglaze profile which has a peak temperature of 525°C, also reduced the magnitude of change but by much less, and the slope at 200°C remained the same. Thermal ageing unglazed resistors at 200°C in nitrogen produced the surprising result that the magnitude of the resistance change was higher than for equivalent resistors aged in air (but still obeying a square root of time dependence) (Fig. 4(b)[5]). The behaviour of glazed resistors was identical with those aged in air but laser trimming, which exposes a small part of the resistor along the cut, produced results between the glazed and unglazed.

The aged behaviour of the bulk resistor is probably related to changes occurring in the interface region between the ruthenate particles. If the glass contains



(a) DP 1431 overglazed with 9137 aged in air.

(b) DP 1431 aged at 200°C in nitrogen.

Fig. 4.

ruthenium ions which act as impurity centres that aid tunnelling, these centres may diffuse together or towards the ruthenate-glass interface reducing the centres available for conduction. The resultant increase in resistance would be related to a diffusion time dependence law which is normally based upon the square root of time, $t^{\frac{1}{2}}$. The temperature dependence would be related to an exponential function of the type, $e^{-E/kT}$, where E is an activation energy and k is Boltzmann's constant. The unglazed resistors aged at the higher temperatures obeyed the diffusion law and the higher magnitude of change for the resistors in nitrogen is consistent with the diffusion loss of oxygen from the resistor top surface.

It is difficult to explain time dependence less than $t^{\frac{1}{2}}$ but more complex diffusion laws have been observed involving $t^{\frac{1}{4}}$ relationships. The rate of change at the lower temperatures may be fitted to a $t^{\frac{1}{3}}$ law and a suggested thermal ageing mechanism is stress relaxation.[6] The behaviour of overglazed resistors would be determined by stress relaxation but trimming and exposing part of the resistor would lead to ageing due to a combination of diffusion and stress relaxation at 200°C. If stress relaxation is the basic bulk mechanism for the DP 1400 resistor system, further firings might anneal out some of the stresses, and therefore reduce the magnitude of change without changing the rate of change. This would explain the reduction in magnitude of change in resistance found between unglazed resistors as fired, and those refired through the overglaze profile.

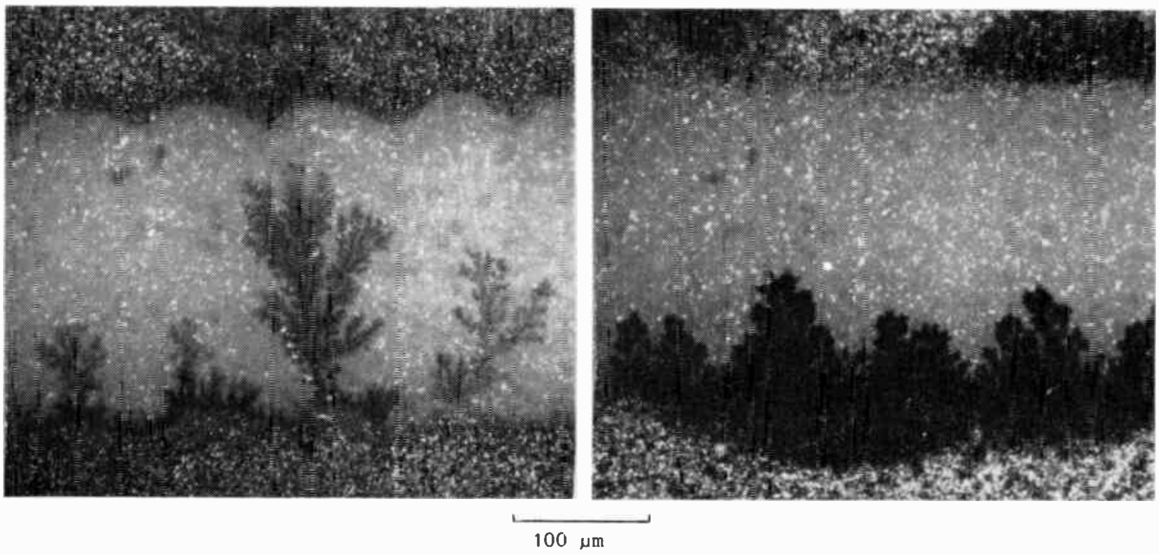
It may be concluded that unglazed resistors aged at high temperatures are subject to resistance changes which are related to a diffusion process, probably at the resistor top surface, and these changes can be inhibited by overglazing. At lower temperatures the dominant mechanism is probably stress relaxation. The combination of overglazing and control of the firing profile can produce highly stable, reproducible resistors.

3 Silver Alloy Conductors

Gold has been used as the traditional conductor material for high reliability applications and, in combination with platinum, provides a solderable termination material. The possibility of silver migration occurring in the presence of moisture has led to considerable doubts as to the suitability of silver alloy conductors for such applications. Nevertheless, the drive to reduce costs encourages circuit manufacturers to re-examine the possibility of the use of silver-loaded conductors.

The migration of silver under voltage loading in the presence of moisture is well known but the conditions under which the migration is sufficiently inhibited, so that it no longer presents a problem, are less well known. A variety of claims have been made by ink manufacturers that their proprietary formulations are resistant to migration. The claims are often based upon a so-called 'water drop' test in which a drop of de-ionized water is placed across two conductor tracks and d.c. voltage is applied. Either the time taken to produce shorting at a fixed voltage is measured or the voltage is increased in steps until shorting occurs. Claims based upon such tests are not always useful in determining likely behaviour under less severe conditions and testing under better controlled humidity and temperature conditions is required.

As dendrites form in a gap between conductor tracks, the resistance across the gap drops considerably and, if no protective resistor is put in series, the power dissipated could destroy the dendrites. Three protective resistors (1 MΩ, 10 MΩ and 100 MΩ) were used in experiments in dendrite growth at STL, and short-circuits only occurred between tracks with the 1 MΩ and



(a) 10 MΩ resistor. (b) 1 MΩ resistor.

Fig. 5. Dendrites formed in 85% r.h. at 85°C with different values series resistors.

10 MΩ resistors in series.[7] The dendrites were of a much finer filamentary nature for the 10 MΩ resistors than for the 1 MΩ resistors (see Fig. 5(a) and (b)). In order to form dendrites, current must flow between the tracks to allow migration to occur, but as the resistance across the gap decreases more of the voltage is dropped across the limiting resistor. If this resistor is too high there will never be enough current to drive the silver across the gap, and if it is too low, fine dendrites will be destroyed. Hence the conductors in series with the 10 MΩ resistor had the fastest failure times.

The classical description of silver migration is that, at the anode, a uniform zone of colloidal silver oxide and silver migrates towards the cathode and dendritic silver grows from the cathode to the anode.[8] From examination of the failed samples where dendritic growth had not bridged the gap between conductor tracks, dendrites were always associated with the cathode. In the case of a silver-platinum fired on a glass substrate, the zone from the anode was also observed, and within this zone there were thicker filaments of silver.

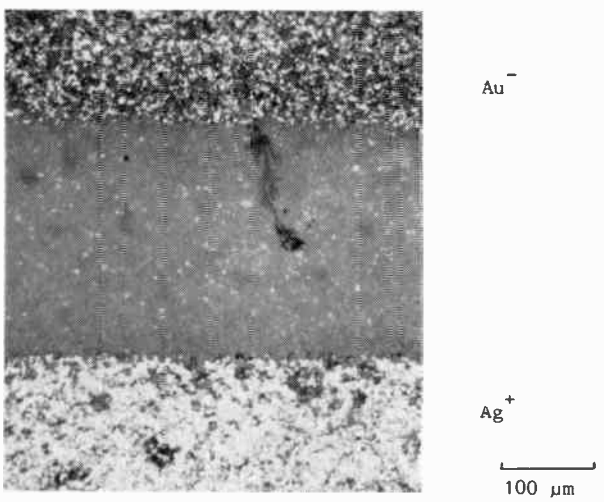


Fig. 6. Silver dendrite growing from gold cathode.

A test pattern was devised so that each half of an interdigitated pattern with lines separated by 250 μm gaps could be printed and fired separately with different inks. Samples of silver were paired with gold and silver-palladium. One sample from each pair was biased with the silver electrode negative and the other with the silver positive. After 370 hours in 85%r.h. at 85°C, with 20 V d.c. load, both polarity AgPd/Ag combinations and the Ag+/Au− had shorted but the Ag−/Au+ was unaffected. Both the AgPd/Ag samples had dendrites growing from the cathode but there were more in the sample with the AgPd as the cathode, and when the gold was the cathode, dendrites were observed originating from the gold (Fig. 6), i.e. silver transfers from the anode to the cathode and silver dendrites grow back from the gold cathode.

As well as surface migration, concern has been expressed at the possibility of silver movement through dielectrics as in cross-overs. Tests have been carried out with silver-palladium and silver-platinum palladium alloys sandwiched around double thicknesses of dielectrics in 85%r.h. and 85°C and no failures were found after 5000 hours with 25 V d.c. applied. By 6500 hours, however, evidence of dendrites was found on the dielectric top surface and, in two cases, electrical shorts had occurred between the top and bottom conductors. Silver is known to dissolve readily in glasses with concentrations of up to a few percent and some silver dissolution may occur during firing. Multilayer dielectrics are heterogeneous mixtures of crystalline particles either as fillers or as a devitrifiable phase. Further bulk diffusion may occur during the test, but a more likely mechanism is the movement at the interfaces between the crystalline and glass phases.

Silver movement has also been observed at high temperatures, typically 150°C under dry no-load conditions and the mechanism is related to the formation

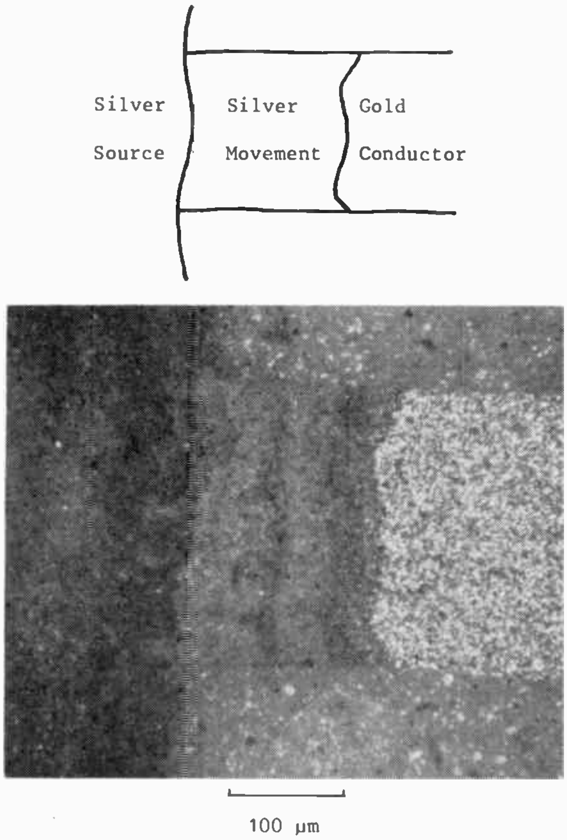


Fig. 7. Silver movement over a gold thick film conductor at 150 C.

of silver chlorides or sulphides with chlorine and hydrogen sulphide in the atmosphere (Fig. 7). The movement is concentrated on gold conductors which at these temperatures provide clean, oxide-free conducting surfaces.

In all cases of silver movement, either by dendritic growth between electrodes or ionic movement through dielectrics, a covering of a dielectric overglaze significantly inhibits this movement. A resin encapsulant also usually inhibits migration but there is considerable variation in the effectiveness, depending upon the resin composition. Silicone resins perform well as long as the substrate is clean enough to allow the resin to thoroughly bond to the substrate, but epoxide resins only delay dendrite formation. Silicone resins do not prevent moisture ingress but effectively stop condensation at the substrate surface, whereas epoxide resins, which are less permeable, do allow eventual moisture condensation.

Under sufficiently severe conditions silver will always migrate and the amount of migration will depend upon the silver concentration and the ambient conditions. The only sure way of preventing this migration, therefore, is to exclude moisture or industrial gases from the conductor surface, either by encapsulation or by sealing in a dry atmosphere.

4 Alternative Substrates to Alumina

The development of enamel steel substrates with controlled purity of the enamel offers the potential for extending hybrid circuitry to large areas.[9] A number of sources of enamel steel substrates are now available; two of which were evaluated, Alpha Metals, which has enamel on both sides and Erie Ceramic Arts, which is single sided. The composition of the enamels of the two substrate types was found to be based upon a borosilicate glass containing oxides of barium, calcium and zinc. The compositions were similar except that the Alpha enamel had a higher silica and zirconia content and lower alumina content. Sections through the enamelled steel substrates were examined on the scanning electron microscope. The zirconia was found to be a separate discrete phase within the enamel and in the Alpha substrates there was also a separate silica phase.

The enamel softening points were determined using thermomechanical analysis, and were found to be 583°C and 595°C for Alpha and Erie respectively. This means that both enamels will have low viscosities at 650°C and explains the considerable interaction between conductor glasses and the enamels when fired at this temperature. It was also noticeable that mesh marks were clearly visible at a peak temperature of 650°C. To a lesser extent, the same effect was observed when an Erie substrate was fired through the same profile with the enamel side in contact with the belt.

A number of silver alloy conductors have been produced for firing on to enamel steel substrates. The resistivity values of silver and silver-platinum conductors are comparable with the values obtained on alumina, but

Table 1

Resistivity and adhesion strengths on alpha substrates

Ink type	Ink number	Resistivity $\Omega/\square \times 10^{-3}$	Thickness μm	Resistivity Ω cm $\times 10^{-6}$	Adhesion strength kgf†
Ag	DP 7713	$3{\cdot}48 \pm 0{\cdot}04$	12·7	$4{\cdot}42 \pm 0{\cdot}05$	$1{\cdot}8 \pm 1{\cdot}7$
	ESL 590	$13{\cdot}9 \pm 1{\cdot}2$	7·0	$9{\cdot}7 \pm 0{\cdot}8$	$2{\cdot}8 \pm 1{\cdot}4$
AgPd	DP 7711	$57{\cdot}7 \pm 6{\cdot}2$	28·3	163 ± 18	$4{\cdot}5 \pm 3{\cdot}4$
	EMCA 6143–2	161 ± 12	7·0	112 ± 8	$3{\cdot}9 \pm 1{\cdot}5$
	ESL 9695	186 ± 21	9·0	167 ± 19	$3{\cdot}1 \pm 1{\cdot}9$
AgPt	DP 7712	$3{\cdot}22 \pm 0{\cdot}34$	22·0	$7{\cdot}08 \pm 0{\cdot}75$	$4{\cdot}4 \pm 3{\cdot}1$

† 2·5 mm square pads.

silver-palladium conductors had higher resistivities because of palladium oxidation during firing (i.e. palladium oxide does not dissociate in the presence of silver until about 690°C).[10] (Table 1).

Adhesion strengths in excess of 2 kgf for a 2·5 mm × 2·5 mm pad were typical and failures tended to be in the enamel. The resistor series available were all found to give considerably higher than expected resistance values for the 1 MΩ/□ inks. Stability and t.c.r. values were not as good as has been found on alumina.

An alternative substrate material is polymer board as used in conventional printed circuit boards, on to which can be printed and cured silver-loaded resins to act as conductors (rather than etched copper), carbon-loaded resins as resistors and alumina or silica-filled resins as dielectrics. Conductor adhesion strengths were found to be limited to the inherent strength of the polymer board and the track resistivities were about 0·7 Ω cm comparable with silver-loaded epoxide resins used for chip attachment.

Resistor inks with resistivities between 100 Ω/□ and 60 kΩ/□ were examined and found to be stable, within ±0·5%, after ageing at 125°C for an hour but showed a considerable increase in resistance (up to 40% for the high resistivity) when exposed to solder bath temperatures (about 265°C) for sixty seconds (see Table 2).

Table 2

Percentage change in resistance of carbon/resin resistors after thermal ageing

Substrate	After 1 hour 125°C	After 1 min at 265°C	7 days at room temperature	5 cycles −77°C to +150°C
110 Ω/□	−0·34±0·06	16·8±1·2	−1·8±0·1	−2·4±0·9
850 Ω/□	0·28±0·21	13·4±2·8	−2·1±0·2	−1·2±0·8
6·5 kΩ/□	0·25±0·07	4·7±1·4	−3·5±0·3	−5·7±0·3
60 kΩ/□	0·04±0·21	14·2±5·6	−6·3±0·5	−12·8±2·0

After a further 7 days at room temperature there was some recovery and after five rapid thermal cycles between −77°C and +150°C there was further recovery. This ageing mechanism is related to the internal destruction of the conducting path between carbon flakes at temperatures in excess of the glass transition temperature T_g of the resin.

A third substrate material is soda lime silicate glass used for liquid crystal and other displays. Palladium alloy conductors were precluded because of oxidation in the firing temperature range used, typically up to 550°C.[10] The level of sintering of conductors becomes critical at these low firing temperatures. The problem is especially noticeable when the peak firing temperature is below 550°C, particularly for gold conductors. At higher firing temperatures adequate adhesion strengths have been found although the spread in values is greater than for alumina. Care must also be taken when soldering to poorly sintered conductors because the leach-resistance will be reduced. The conductor surface will also be less suitable for wire bonding.

Resistor performance on glass substrates was generally inferior to that found on alumina. The lower peak firing temperature means that the glasses used must have lower softening temperatures and are likely to be more susceptible to attack, especially by moisture. Nevertheless the stability after thermal ageing was remarkably good. The resistivity of the inks is more difficult to control especially with the high resistivity members and this may be related to excessive interaction with the substrate.

5 Base Metal Systems

A variety of base metal conductors have been available for a number of years, both for firing in air and in nitrogen atmospheres. Of these conductors, however, nitrogen-fireable coppers are the most interesting because of their high adhesion strengths and good solderability properties. The d.c. conductivity is much lower than for bulk copper which may be related to the inherent conductor porosity and perhaps partial oxidation of the copper between grains in the matrix of the conductor.

Du Pont have recently released a nitrogen-fireable resistor system[11] based upon an experimental system which has been available over a limited range of resistivities from 20 Ω/□ to 10 kΩ/□ to selected users since 1979. This resistor system is compatible with

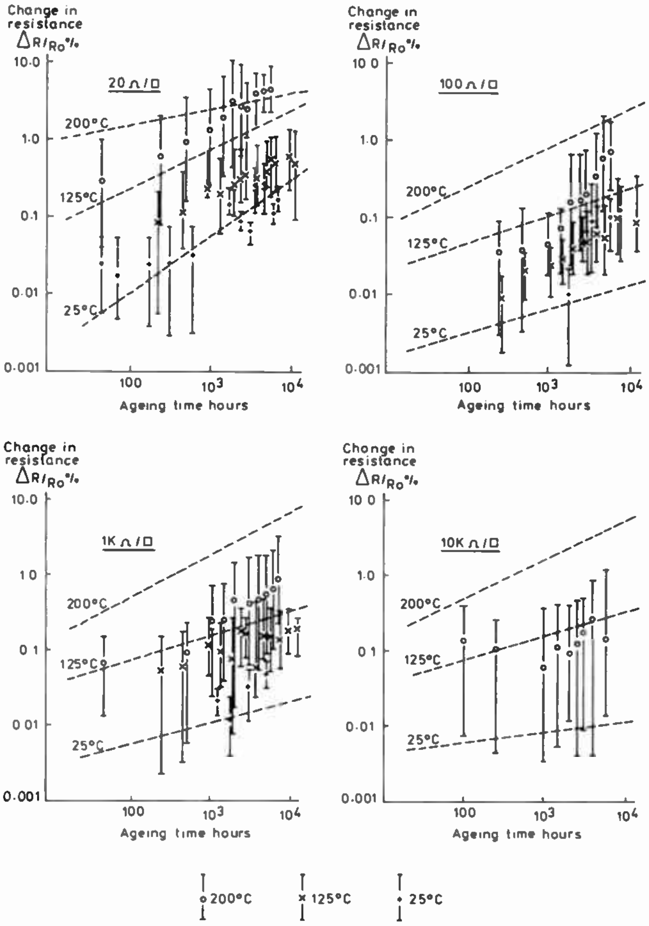


Fig. 8. Experimental nitrogen-fireable Dupont resistor system.

copper conductors and the composition is known to be similar to that covered by an STC patent[12] which was based upon lanthanum hexaboride as the conducting oxide in an alumino-borosilicate glass.

Resistors fired at a peak temperature of 900°C in nitrogen, from each resistivity ink of the experimental system, were aged in air at 200°C, 125°C and room temperature and the percentage change in resistance from the initial value $\Delta R/R_0$ was plotted on log/log graphs (see Fig. 8). The changes in resistance followed positive trends with the exception of the 10 kΩ/□ ink which was negative when aged at 125°C. The equivalent average changes for the unglazed DP 1400 resistor series are plotted for comparison and the stability especially at the lower temperatures compares favourably with this series. The rate of change of resistance for the 20 Ω/□ and 10 kΩ/□ resistors aged at 200°C and 125°C fitted reasonably well to a square root of time law consistent with a diffusion process. This is a faster rate of change at 125°C than for the DP 1400 'Birox' series and, if continued, would lead to greater increases at longer times (i.e. a crossover will occur probably after between 10^4 and 10^5 hours). The rate of change for the 1 kΩ/□ at 200°C is greater than $t^{\frac{1}{2}}$ but at 125°C is less than $t^{\frac{1}{2}}$ as for the DP 1400 series. The 10 kΩ/□ results were disappointing in comparison, with sometimes negative changes of over 1% being found after a few thousand hours at 125°C.

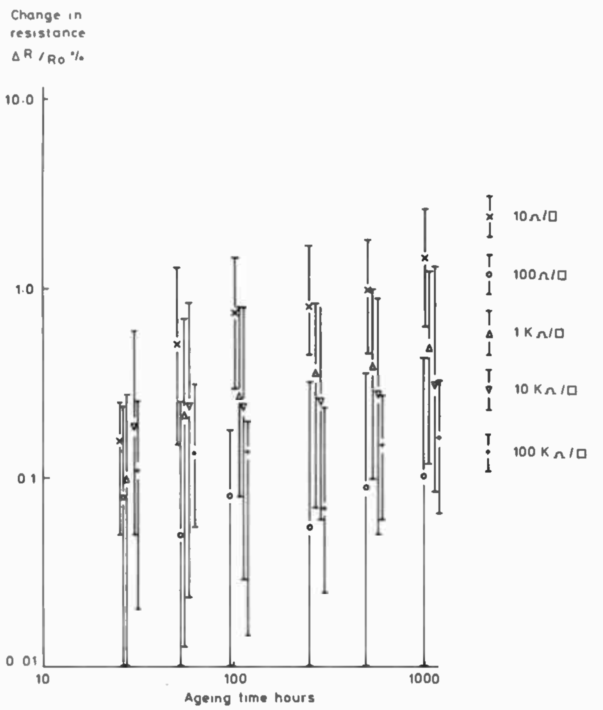


Fig. 9. DuPont CMS nitrogen-fired resistor system.

Further resistors were fired at peak temperatures between 850°C and 950°C and the change in resistance after ageing at 150°C for 2500 hours was found to be less than ±2% in all cases. More recently, laser trimmed samples of the 10 Ω/□ to 100 kΩ/□ (actually nearer 70 kΩ/□) Du Pont commercial resistor series have been aged up to 1000 hours at 150°C (Fig. 9). The changes in resistance were generally positive and the 100 Ω/□ and 100 kΩ/□ resistors were much more stable (less than 0·4% change) than the other three members of the series (up to 2·7% change). Untrimmed resistors aged up to 3000 hours behaved similarly and the rate of change of resistance for both trimmed and untrimmed resistors was less than $t^{\frac{1}{2}}$. The longer resistors generally changed by a greater amount than the shorter ones, and the shorter resistors had initial resistance values lower than expected, consistent with copper diffusion from the conductor into the resistor during firing.

As well as thermal ageing, experiments in 85% relative humidity at 85°C and under d.c. load after laser trimming have given results that are similar to those found for the unglazed DP 1400 'Birox' resistor system. At present, however, no overglaze is available for the nitrogen fireable system, and generally the overglazed DP 1400 system performs much better than when it is unglazed.

The ageing mechanism at 200°C was consistent with a diffusion process and may be associated with oxygen movement as was the case for air fireable systems. Whether oxidation of any component within the resistor system occurs was not determined but on heating the unfired inks to 750°C in air for two hours, weight gains were found. Reference is made in the STC patent to the ability to fire lanthanum hexaboride in air but the resistance values obtained in air were considerably greater than those in nitrogen. The ageing mechanism at the lower temperatures may be due to stress relief as for the DP 1400 'Birox' system. A second mechanism related to interaction between the conductor and the resistor was identified for the commercial resistor series. The most likely explanations are either diffusion of copper into the resistor during the firing which reduced the internal stresses, or some change is occurring during ageing which leads to negative changes in resistance values such as further copper diffusion into the resistor.

The shape of the t.c.r. curves and the room temperature resistance values were found to be sensitive to the peak firing temperature. Metallic type conduction consistent with lanthanum hexaboride tends to dominate the behaviour of the low resistivity inks, but at sufficiently low temperatures a negative t.c.r. consistent with semiconductor behaviour was observed.

Nitrogen-fireable dielectrics are also now available but a problem has been found with blistering when using them as crossover materials. This blistering is associated with the copper conductors sandwiched about the dielectrics and is thought to be due to partial oxidation of the copper which on refiring is reduced, releasing oxygen. By reducing the drying temperature of the copper from 150°C to 120°C, less oxidation occurred and the blistering was reduced. More recently a Du Pont dielectric 4275 has been produced which shows no evidence of blistering, but there was some suggestion that this was due to increased porosity which may make it more susceptible to the development of shorts between copper cross-overs. However, samples of DP 9922 copper conductors sandwiched across DP 4275 have survived 600 hours in 85% r.h. at 85°C with 24 V d.c. load without any evidence of shorting.

6 Conclusions

To maintain a system performance comparable with DP 1400 and gold conductor terminations, but at a lower cost, the choice must lie between a nitrogen-fireable base metal system or a silver alloy terminated air-fireable resistor system. The use of alternative substrate material such as enamel steel or polymer board may be suitable for particular applications such as for large areas, but the performance of the available materials is inferior. The choice, therefore, between a copper or a silver-based system depends upon whether the capital costs of setting up a nitrogen-fireable facility can be justified, or whether moisture can be prevented from reaching a silver surface with any certainty. Whichever route is chosen, results comparable in quality with gold systems are possible but considerably more effort must be put into maintaining a high level of process control.

7 Acknowledgments

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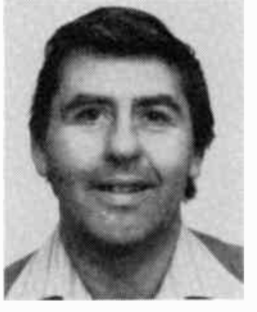
The Author



Martin Coleman graduated in physics at the University of Sheffield in 1964 and gained an M.Sc. in solid state physics the following year. For the past ten years at Standard Telecommunication Laboratories, he has been head of the Thick Film Group which has been involved in examining the stability and limitations of materials used in hybrid microcircuits. The Group has been particularly concerned to identify the underlying mechanisms behind the behaviour of these materials from which it is hoped meaningful predictions can be made of life expectancy in high reliability applications.

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Indexing Terms: Multiplexing equipment, Security Digital Speech

Voice encryption in multichannel paths

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SUMMARY

A proposal is presented for a system for the encryption of the signal resulting from the time division multiplexing of N channels carrying digital voice signals. The channel synchronization principle is such that no extra sync-signal is transmitted but synchronization is extracted from the information signal. Furthermore, it provides a first echelon encryption of the signal because there is a continuous and random shift of the position of each channel in time. The main encryption of the signal is done on a bit-by-bit basis, by modulo-2 adding to the signal a sequence that is derived from a non-linear combination of several past bits of the encrypted signal. This is achieved by using an e.p.r.o.m. in conjunction with a r.a.m. In an alternative system, this sequence is derived from a pseudo-random sequence generator that is also controlled by some past bits of the encrypted signal. In both cases synchronization and encryption are based on the transmitted signal itself.

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1 Introduction

Digital voice encryption is achieved, in general, by adding modulo-2, on a bit-by-bit basis, the digital signal representing voice to a synchronous digital random sequence. The analogue voice signal is converted to a digital one by using either pulse code modulation or delta modulation techniques. In this application of voice encryption, delta modulation is preferred because of its simplicity and low bit rate in comparison to p.c.m. Actually, the continuously variable slope delta (c.v.s.d.) modulation technique provides satisfactory voice quality at bit rates as low as 9 Kbit/s.

The bit stream used in the encryption of the digital voice signal is usually generated by a pseudo-random sequence generator (p.r.g.) of maximum length with non-linear feedback. For the sake of security, a p.r.g. should have a very long period, several hours or even days, so that it is necessary to send a characteristic signal at the beginning of each communication to synchronize the p.r.g. at the receiving end. This procedure creates the problem that any incorrect reception of the synchronizing message causes a failure in communication. Also, after the transmission of this synchronization message is completed, it is impossible for newcomers to join the radio network. A re-synchronization must occur at reasonable time intervals to overcome these difficulties.

In multichannel communications, based on time division multiplexing (t.d.m.), the synchronization of channels between the transmitter and the receiver is of great importance. A widely used way of achieving synchronization is to send a particular bit pattern that gives a time reference to the receiver. This sync. signal occupies all or a part of the capacity of one information channel.†

The system described here has been designed mainly to satisfy the encryption requirements in a multichannel transmission. The channel synchronization principle is quite different from the usual one in that no channel space is used for the synchronization which is extracted from the transmitted signal. This is done in such a way that it provides a first-echelon encryption. For the main encryption of the t.d.m. signal, the process followed is different from the one mentioned above, where a pseudo-random sequence generator is used. On the contrary, the sequence used in the bit-by-bit encryption of the t.d.m. signal is generated by combining, in non-linear logic, several past bits of the transmitted signal.

2 Description of the System

2.1 Self-synchronizing T.D.M.

The block diagram of the N-channel time division multiplexer is shown in Fig. 1. The analogue signal of each channel is converted into a digital one via a continuously variable slope delta (c.v.s.d.) modulator. The encrypted t.d.m. signal, coming from the encryption unit, is applied to the serial input of a shift register and is shifted to the right at the rate of the clock f_0 coming also from the encryption unit. The shift register has a length

† Dagakis, K., Mantakas, C. and Papadopoulos, G., 'New technique for channel synchronisation in time division multiplexing', *Electronics Letters*, 16, no. 25, pp. 944–5, 4th December 1980.

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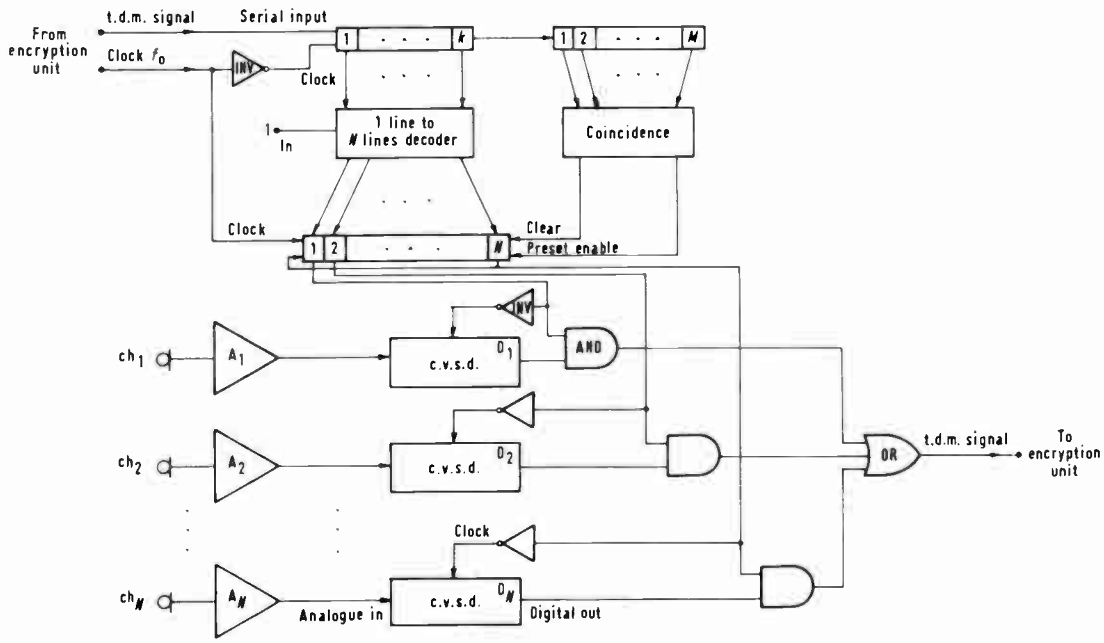

Fig. 1. N-channel multiplexer.

of $K+M$ bits, where K is related to the number of channels by $N = 2^K$.

The first K parallel outputs of the shift register serve as addresses to the decoder. The decoder input is always kept in the logical '1' state. Only one of the decoder N outputs will be in the state '1' during each clock, depending on the K-bit input address. Each one of the N lines is connected to the corresponding preset input of an N bit shift register. It has the same clock (f_0) and there is feedback from the last bit to the first one. Each one of the N outputs is inverted and serves as the clock to the corresponding c.v.s.d. modulator. Its frequency is equal to f_0/N, in practice it is about 16 kHz. The digital output of the c.v.s.d. modulators is a bit stream having a bit duration equal to N times the clock f_0 period. The AND gates that follow pass only the part that coincides in time with the logical '1' pulse appearing in the corresponding output of the N-bit shift register. The outputs of the AND gates pass through an OR gate thus forming the t.d.m. signal going to the input of the encryption unit.

The block diagram of the demultiplexer is shown in Fig. 2. The channel synchronization circuit, consisting of two shift registers, the decoder and the coincidence circuit, is exactly the same as that described in the multiplexer. If the received signal is the same as the transmitted one, these circuits have the same input signal and clock, and consequently they both yield the same output signal. Synchronization occurs every time the

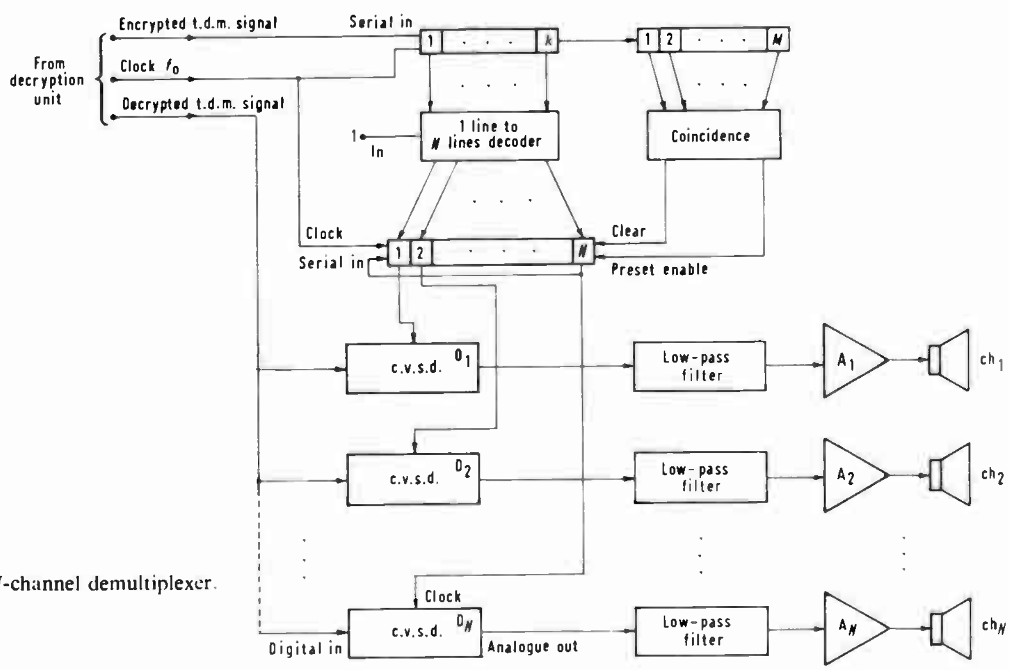

Fig. 2. N-channel demultiplexer.

coincidence circuits at both ends are triggered. The N parallel outputs of the second shift register serve as the clock to the corresponding c.v.s.d. demodulators, which have the decrypted t.d.m. signal, coming from the decryption unit, as common input. Thus, demultiplexing is completed and digital-analogue conversion follows.

As has already been mentioned, synchronization occurs every time that M consecutive bits of the encrypted t.d.m. signal coincide with those of the coincidence circuit. This occurs in a random way and depends only on the statistics of the information signal. It has an 'average' frequency equal to $f_0/2^M$. In the time interval between two successive instants of synchronization the channel sequence in time is always the same, but the choice of the first channel is random.

2.2 Main Encryption Process

The block diagram of the encryption/decryption unit is shown in Fig. 3. The digital t.d.m. signal is encrypted, bit-by-bit, with a sequence ⊗ generated in the following way: The encrypted signal, at the output of the exclusive-OR gate, is applied to the serial input of a n-bit shift register SR1 and is shifted to the right at the rate of the clock frequency f_0. The shift register parallel outputs serve as the addresses for an e.p.r.o.m. that has been written using a random digital signal. The same shift register provides the addresses for a r.a.m. that has the same size as the e.p.r.o.m. One e.p.r.o.m. output, one r.a.m. output and one of the second shift register parallel outputs are added modulo-2 to give the sequence ⊗ used for the encryption of the t.d.m. signal. Notice that the

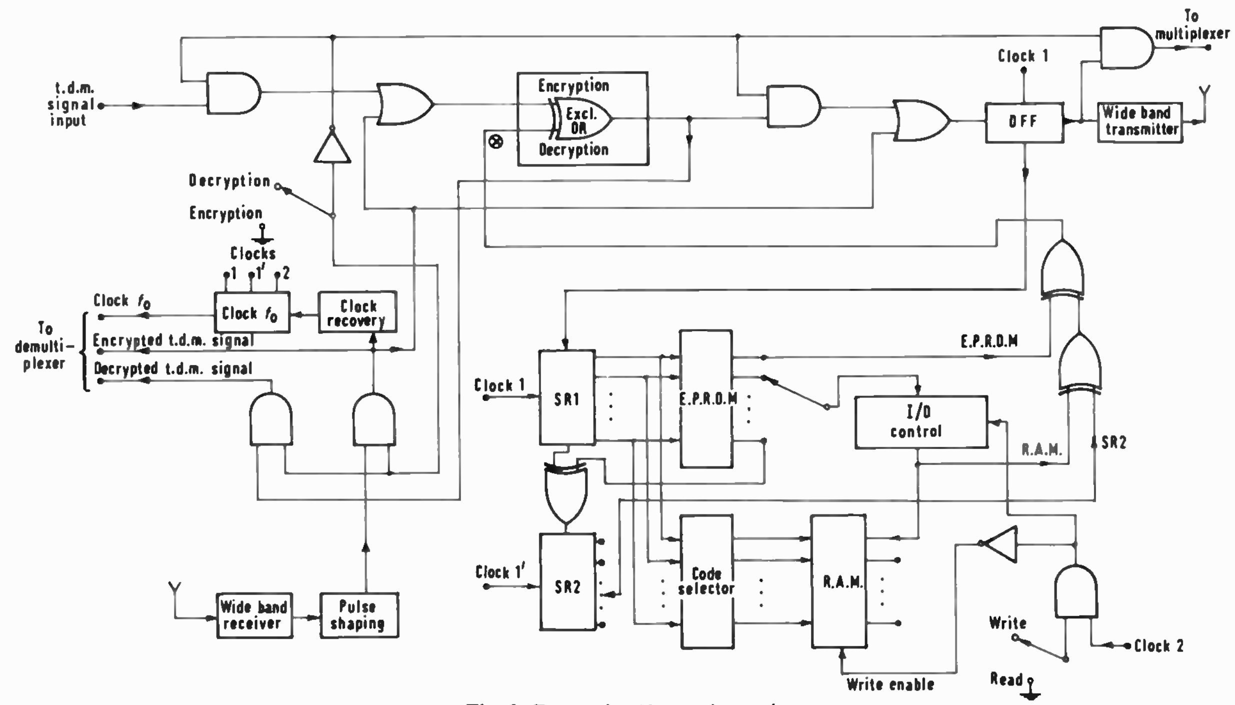

Fig. 3. Encryption/decryption unit.

The effect of this synchronization scheme on the information signal carried by each channel is now examined. The clock signal to the c.v.s.d. modulators/-demodulators would have a stable frequency f_0/N if a conventional synchronization technique was used. As soon as a coincidence occurs, a new synchronization sequence is established, which means that the first clock pulse of each c.v.s.d. modulator/demodulator presents a shift from the time it is expected. This time shift can take one of the following discrete values: 0, $\pm 1/f_0$, $\pm 2/f_0$, $\ldots \pm (N-1)/f_0$. The effect of this on the signal is negligible, if the average sync.-frequency is low. There is an optimum value for this frequency, depending on the transmission channel conditions. If this frequency is too low, any loss in synchronization will cause a failure in communication till the next sync. time. In our implementation synchronization occurred randomly with an average period of about 2 seconds and there was no detectable alteration in the audio signal of channels.

content of the second shift register SR2 is not the transmitted signal, but results from the modulo-2 addition of the transmitted bit stream and one of the e.p.r.o.m. outputs.

In normal operation, the READ-WRITE switch is in the READ position. In the beginning of a communication, this switch is turned in the WRITE position, thus connecting one e.p.r.o.m. output to one of the r.a.m. inputs. This occurs during the time that the clock pulse 2 is in the logical '1' state so that the data from the e.p.r.o.m. are enabled to be stored at the selected r.a.m. addresses. Each different arrangement of the shift register parallel outputs that are connected to the address inputs of the r.a.m. results in a different content for the r.a.m. This can be used as a key for the encryption.

The same unit can operate in the decryption mode by turning the corresponding switch to the decryption position. The sequence ⊗ is generated by a non-linear combination of n past bits of the transmitted signal.

Thus, if this signal is reconstructed correctly at the receiver, it generates the same sequence ⊗ used in the decryption.

2.3 Alternative Main Encryption System

The block diagram of such a system that uses a pseudo-random sequence generator for the encryption of a digital signal A, (t.d.m. signal) is shown in Fig. 4. A quite similar system is used for the decryption in the receiver. The modulo-2 addition of the signal A to the coding sequence ⊗ gives the encrypted signal S that is transmitted. This signal is also applied to the serial input of a shift register of N_0+N_1 bits length. The N_0 parallel outputs are rearranged in the block named 'key' and are connected to the preset inputs of a p.r.g. of N_0 bits length. Each different arrangement gives a different key for the encryption, with a total number of possible keys equal to $N_0!$

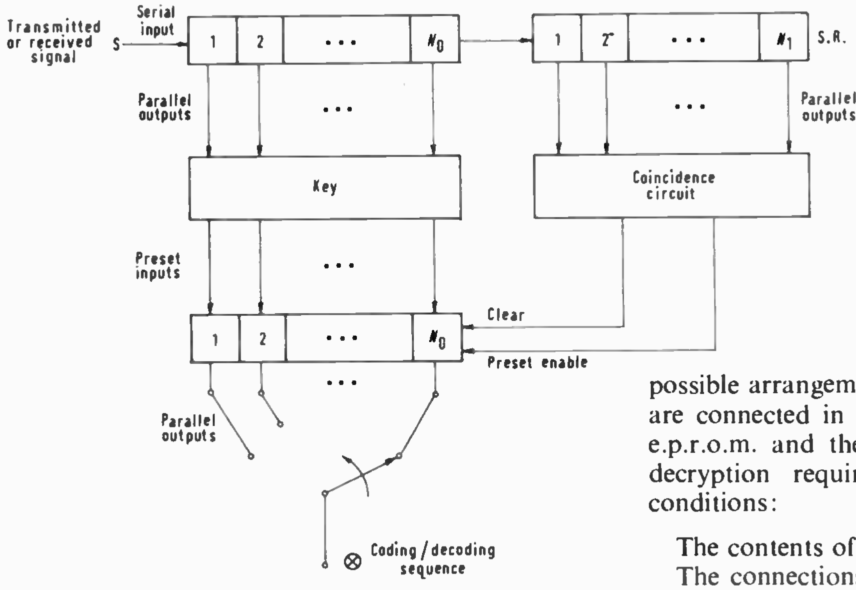

Fig. 4. Encryption/decryption unit using a pseudo-random sequence generator.

The 'coincidence' circuit is activated only for one specific combination of the N_1 parallel outputs of the shift register. In its simplest form, it will be a N_1-input AND gate. When this circuit is activated, a CLEAR pulse followed by a PRESET ENABLE pulse stops the normal cycle of the p.r.g. and enables the N_0 bits of the shift register to form the new starting point of the p.r.g.'s cycle.

The only difference in the receiver is that the input signal to the shift register is the received encrypted signal $\hat{s}$ which, of course, is identical to s, if there were no errors in the reception. Thus, the system generates a sequence quite similar to the coding sequence ⊗ and the modulo-2 addition of this new sequence to the received signal, bit-by-bit, gives the decrypted signal A.

From the above description it is evident that synchronization is achieved by the first time that the 'coincidence' circuit is activated and it is repeated automatically, in a random way, with an average frequency of $f_0/2^{N_1}$, where f_0 is the signal bit-rate. According to the value of f_0, N_1 must be selected so that synchronization is achieved in a reasonable time.

3 Conclusions

The proposed system is designed for voice encryption in multichannel transmission. The method of self-synchronizing time division multiplexing/demultiplexing is simple and presents some advantages over some others in use: channel synchronization is derived from the transmitted signal itself, so that there is no need to send an extra sync. signal that would occupy all or a part of the capacity of one voice channel. Furthermore, this method provides a first-echelon encryption as explained previously.

For the main encryption of the signal, the proposed self-encrypting system is very simple because the encrypting sequence is derived from the transmitted signal using a non-linear logic involving an e.p.r.o.m. and a r.a.m. The e.p.r.o.m. can be written in numerous different ways that may be considered as an internal key for the encryption. Another variable key consists of the possible arrangements of shift register SR_1 outputs that are connected in parallel to the address inputs of the e.p.r.o.m. and the r.a.m. It is evident that a correct decryption requires the following identical system conditions:

The contents of the e.p.r.o.m.

The connections of the shift registers outputs to the address inputs of the e.p.r.o.m. and the r.a.m.

The selection of one output of the e.p.r.o.m. and one input of the r.a.m. that is connected to it during the writing process.

The selection of one output of each of the e.p.r.o.m., the r.a.m. and the shift register for the modulo-2 addition that gives the sequence ⊗.

In the alternative system the encrypting sequence is derived from a p.r.g. that is controlled by the transmitted signal. In this case a correct decryption in the receiver requires the correct connection of the N_0 outputs of the shift register to the corresponding preset inputs of the p.r.g. and the correct selection of one of the p.r.g. outputs that will give the sequence ⊗. This system can be used either in the place of the process mentioned in Section 2.2, or in conjunction with it.

All these systems were tested in practice with excellent results.

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Drifting buoy observations using satellite telemetry

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Based on a paper presented at the IERE Conference on Electronics for Ocean Technology in Birmingham in September 1981

SUMMARY

Polar orbiting satellites, equipped with random access data collection and position fixing systems, have made practicable long-term remote oceanographic and meteorological observations by means of instrumented drifting buoys fitted with u.h.f. telemetry transmitters. The paper discusses the performance of nine buoys, deployed by the Institute of Oceanographic Sciences as part of the FGGE Southern Hemisphere Drifting Buoy Network, and presents a small fraction of the considerable data set acquired to date. The requirements for further development of certain sensors and components of buoy hardware are brought out as a result of the experience with these buoys.

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1 Introduction

In the past decade, polar orbiting meteorological satellites have been launched which have included random access systems for data collection from inexpensive low power u.h.f. transmitters and means of locating the position of these. The latest of these systems, ARGOS, is carried by the NOAA-A/F series of satellites and differs from its predecessors using the *EOLE* and *Nimbus*-6 satellites in that, together with the ground services, it is an operational, rather than an experimental, system: a minimum lifetime of ten years is envisaged.

The first major use of the new system has arisen in conjunction with the Southern Hemisphere Drifting Buoy network of the First GARP† Global Experiment (FGGE), undertaken by member countries of the World Meteorological Organization.[1] The objectives of the experiment required a reasonably uniform global coverage of meteorological data. This is particularly difficult in the Southern Hemisphere with its large area of ocean largely devoid of shipping or other means of obtaining observations. The buoy network was to provide, for a year beginning 1st December 1978, observations of sea level atmospheric pressure and sea surface temperature between 20°S and 65°S.

A contribution of nine drifting buoys was made by the Institute of Oceanographic Sciences to the Southern Hemisphere network and a related experiment in the Indian Ocean. Each buoy carried surface pressure and sea temperature sensors, and seven of the buoys were equipped with drogues in order to aid the study of large-scale, near-surface ocean currents, and to complement concurrent oceanographic observations made in the area by the research ship RRS *Discovery*. Two of the buoys were designed with good wave following characteristics and contained accelerometers and simple processors so as to give wave information.

The paper examines the performance of these buoys and assesses requirements for further development in the light of present experience.

2 ARGOS Data Collection and Dissemination

The system operates with two satellites in orbit at any given time. Currently these are *Tiros-N*, launched in October 1978 and NOAA-6, launched in June 1979. Identical receivers are carried by each satellite, developed from the successful *Nimbus*-6 Random Access Measurement System (RAMS). Each Platform Transmit Terminal (PTT) transmits, at intervals preset between 50 and 60 seconds, a message of duration between 360 and 920 ms. This consists of 160 ms of unmodulated carrier followed by bit and frame synchronization, identification, and up to 32 sensor channels of 8 bits each. Split-phase p.s.k. modulation is applied to a carrier of nominal frequency 401·650 MHz. Differences in received carrier frequency allow the satellite to process up to four PTT transmissions simultaneously, and it can cope with up to 1000 PTTs in view at any time as a result of the random distribution of received signals. With this

† GARP = Global Atmospheric Research Programme.

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type of system, it is clearly important that the satellite receiver should be adequately protected against inadvertent continuous PTT transmissions and against excessive transmission power. The carrier stability also is important since the PTT position is subsequently computed from the Doppler shift of the carrier measured at the satellite. These aspects are covered in the PTT specifications and certification procedures for new transmitters.

All data received by the satellite are read out to ground stations in Gilmore, Alaska, and Wallops Island, Virginia and thence to the spacecraft operational centre at Suitland, Maryland. From there the PTT data are forwarded to the Centre Nationale d'Etudes Spatiales at Toulouse for processing by Service ARGOS. For latest position information and data the user is able to interrogate the computer at Toulouse at any time: complete data sets are despatched fortnightly on magnetic tape. Meteorological data can be disseminated worldwide using the Global Telecommunications System (GTS) of the WMO.

3 Drogued Buoys

The seven drogued buoys were basically of Canadian design and manufacture (Hermes Electronics Ltd), although they were modified to incorporate an IOS drogue loss sensor as described below. In view of the primary FGGE requirement for long term reliability in the barometric pressure sensor it was considered preferable in this experiment to use the product of a well established development programme.[2]

The principal features of the buoys are shown in Fig. 1. The hull was a truncated cone of welded aluminium, with a foam-filled flotation collar, which contained a 160 Ah manganese alkaline battery pack, a Paroscientific Digiquartz 230A pressure sensor, two thermistors, a drogue loss sensor, and the PTT. An upper reinforced plastic cone housed the antenna, and provided a suitable position for the atmospheric pressure port. The buoy is 185 cm high and 85 cm diameter.

The specification for FGGE buoys called for ± 1 mbar accuracy in pressure measurement and $\pm 1°$ in temperature. Extensive testing had been carried out in selecting an adequately stable pressure sensor; also in designing the pressure port for minimal dynamic pressures in high winds ($< 0{\cdot}9$ mbar at 25 m/s) and in wave-induced buoy motion. Further protection was given against wave induced errors by averaging the pressure signal over approximately 80 seconds before transmission. Interconnection of the barometer and pressure port so as to maintain a free air-passage at all times, and avoid entry of water, is as shown in the diagram.

A reinforced Dacron window blind drogue was used, of approximate dimensions $2{\cdot}3\ \text{m} \times 7{\cdot}6$ m, carried between angle-iron spreader bars. Some buoys were drogued at 20 m mean depth, others at 100 m or 300 m (Table 1). Multi-plaited polypropylene rope connected a 1 cm chain drogue bridle to a welded lug at the base of the buoy hull and provided some necessary compliance. For the deeper drogues, of course, this tether line added considerable additional drag.

3.1 Drogue Loss Sensor

In view of the unreliability of drogues attached to surface buoys, due to the constant chafing and dynamic loading of the components, it was highly desirable that some means of sensing drogue loss should be incorporated: previous experience has shown[3] that it may not be possible to deduce this reliably from an examination of drift rate. Although various techniques have been tried, a completely satisfactory solution has yet to be devised. Ideally the sensor should not involve any cables or acoustic links external to the buoy, otherwise the

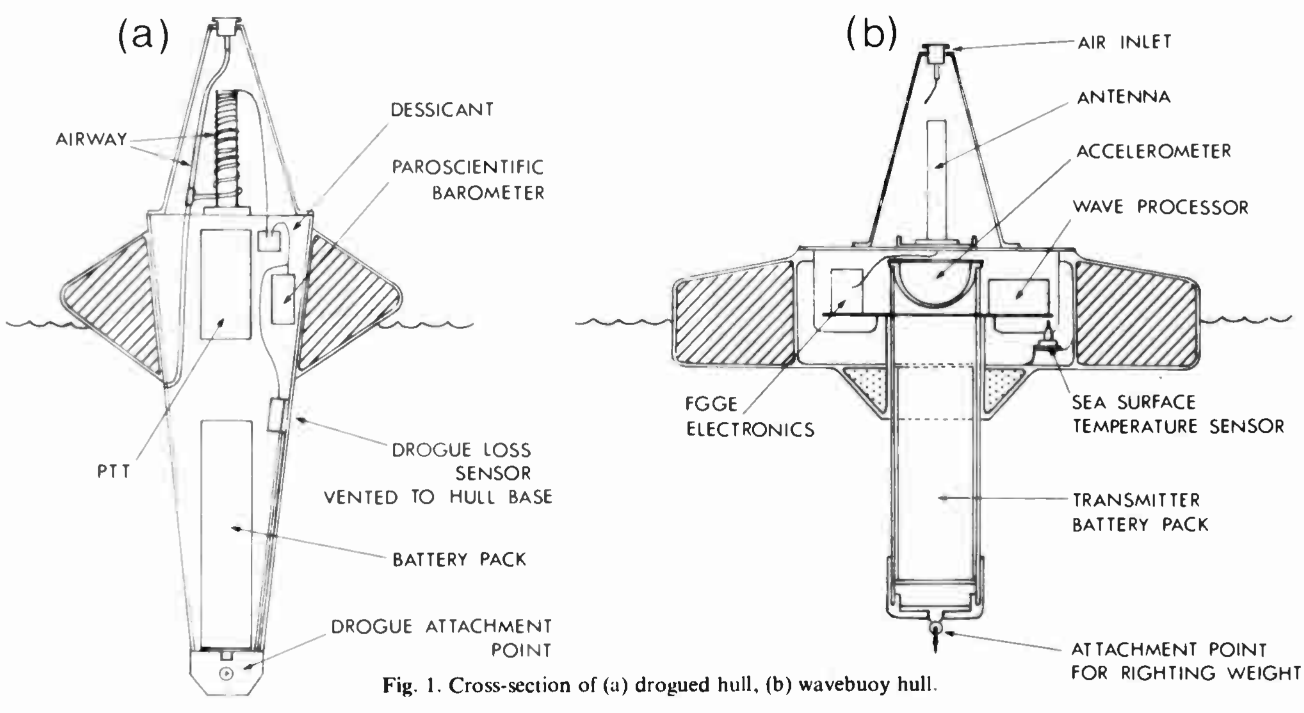

Fig. 1. Cross-section of (a) drogued hull, (b) wavebuoy hull.

reliability of the sensor may turn out to be comparable with that of the drogue. The technique tried in this instance was developed at very short notice and was of limited success but it is felt that, with further development, it could prove reliable.

The method used was to measure the mean depth of immersion of the buoy hull by means of a pressure sensor set into the bottom plate of the hull. Loss of the drogue, which was weighted to 50 kg, resulted in a change in depth of immersion of 7 cm which would cause a pressure change of -7 millibars at the sensor. Since this change was comparable to the possible change in atmospheric pressure in between transits of the satellites, it was preferable to use a relative pressure sensor with the reference side vented to the atmosphere. A suitably protected reference was obtained by tapping into the air line to the Paroscientific atmospheric pressure sensor.

The 350 millibar gauge pressure inlet was joined by a short length of pipe to an inlet in the buoy base plate. The strain gauge bridge was connected into a specially-developed low-power oscillator circuit.[4] The frequency analogue output of this circuit facilitated integration of the sensor output, thereby reducing the effect of fluctuations in pressure due to imperfect surface following. The integration period was set at 64·5 seconds, reducing any fluctuations at the heave resonance of the buoy by a factor of 135. This was further augmented by a factor of 6 due to hydrodynamic attenuation at the depth of the sensor inlet. The six least significant bits of the frequency count were converted to an analogue voltage, in the range 0 to +5 V, which was connected to the PTT multiplexor input normally used for battery voltage sensing. The resolution achieved was approximately 3 mm of water.

The operation of the sensors was checked statically in a wave tank with the appropriate drogue weight attached and detached. The effect of wave action was also checked and the heave and pitch resonances were found to occur at 1·5 and 2·0 seconds period, respectively. It was found that, whereas 17 cm (crest to trough) waves at the pitch resonance had a negligible effect on the drogue loss sensor output, waves of the same amplitude at the heave resonance changed the indicated depth of immersion by +6 mm and -20 mm, with the drogue weight attached and detached, respectively. These changes were due to the non-linear displacement/depth relationship of the hull. The wave amplitude, although seemingly low, was broadly comparable with the high frequency content of a developed sea. The energy contained between 1·25 and 1·75 seconds period in a fully developed spectrum amounts to $8{\cdot}7\times10^{-4}$ m^2 which is equivalent to a significant wave height (for the band limited spectrum) of 12 cm. Thus one would not expect wave action to cause any level changes sufficient to be confused with the 7 cm due to drogue loss.

4 Wave Measuring Buoys

The wave buoys[5,6] were designed to give good wave-following performance. For simplicity of manufacture a discus-shaped hull form was used. Since breaking waves would be expected to capsize such a hull occasionally, it was necessary to incorporate a self-righting mechanism. This consisted simply of a weight suspended from a lever arm below the centre of the hull, giving the overall righting moment characteristic shown in Fig. 2. These buoys are 135 cm high and 150 cm diameter.

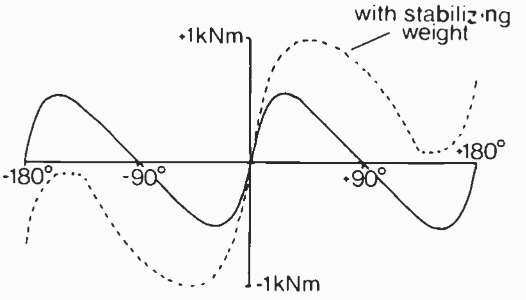

Fig. 2. Righting moment of discus buoy.

The wave sensor consisted of an accelerometer mounted on a damped short-period pendulum and a double integrator incorporating a 4th-order high-pass filter with a cut-off frequency of 0·06 Hz. This was necessary to ensure that the low-frequency error spectrum due to the system of accelerometer stabilization did not contaminate the results. The integrator output had a full scale of $\pm7{\cdot}5$ V for ±15 m displacement. This signal was digitized and subjected to a simple analysis for 20 minutes in every 2 hours by means of a processor using c.m.o.s. m.s.i. circuits. The processor had been used previously in buoys telemetering via *Nimbus*-6 and on the UK Data Buoy DB1. The output consisted of four 8-bit binary parameters which were transmitted after the standard format (atmospheric pressure, battery voltage, sea surface and internal temperatures) in each transmitted frame. The wave data were input to the Handar 620A transmitter serially: this necessitated only minor modification of the transmitter circuits since the latter already incorporated a serial data bus onto which the wave data could be connected via a transmission gate. Care was taken to protect the interface from latch up and other problems in the event of differences developing between the transmitter and processor supply levels. The wave sensor package consumed a total of 90 mW and sufficient alkaline manganese batteries were installed to power it for at least a year at a temperature of 5°C. These batteries were distributed round the buoy's central compartment whilst the main 160 Ah transmitter battery pack was accommodated in a pod below the discus for extra stability.

5 Results

Table 1 summarizes the performance of the buoys, which were deployed from RRS *Discovery* in the Southern Ocean in March 1979 and off the Somali Coast in the Indian Ocean in May and June 1979. Figure 3 shows the complete history of the buoy movements. Apart from Wavebuoy D, early buoy failures occurred without warning although the likelihood is that C was lost in ice in view of the low temperatures (ca. $-1{\cdot}5$°C) recorded shortly before failure. Beyond about a year of operation battery exhaustion is the most likely failure mode: A, for

Table 1

Summary of buoy performance

Buoy	Drogue depth	Deployed Date	Deployed Position	Last received	Last position	No. of days with locations	No. of locations	Probable failure mode	Data quality†
A	20 m	9.3.79	47 58'S, 19 59'E	15.11.80	42 23'S, 145 14'E	614	4961	Switched off on recovery in Tasmania.	Comparisons with Norwegian drifter Day 146 ($\Delta P = -1{\cdot}8$ mbar $\Delta T = 0{\cdot}2$°C). Unflagged.
B	100 m	9.3.79	48 00'S, 19 59'E	21.6.81	42 46'S, 144 56'W	825	6493	Battery exhaustion.	Unflagged. Comparison with Norwegian buoy 24.4.79 ($\Delta P = -0{\cdot}2$ mbar $\Delta T = 0{\cdot}0$°C).
C	300 m	11.3.79	51°49'S, 20°4'E	29.8.79	55 42'S, 59 10'E	171	1429	Abrupt failure. Lost in ice?	Noisy T data. Unflagged.
D	Wave-buoy	11.3.79	51°51'S, 20°8'E	19.4.79‡	53 15'S, 28°35'E	25	77	Ingress of water.	T unusable. Wave data good.
E	300 m	12.3.79	56°02'S, 19 52'E	10.10.80	61 24'S, 164 40'E	579	7032	Battery exhaustion.	P flagged after 24–29.3.79. T flagged 3° low 6–12.4.79.
F	20 m	14.3.79	60 02'S, 20 05'E	11.6.79	59 28'S, 27 31'E	89	763	Lost in ice?	P flagged low 13–19.4.79. T flagged 3° low 6–12.4.79.
G	20 m	13.5.79	00 30'S, 48 58'E	6.5.80	1°44'S, 97 57'E	360	2030	Abrupt failure. Reason unknown.	Unflagged.
H	20 m	15.5.79	00°28'N, 49 01'E	22.1.80	0°31'S, 86 38'E	253	1031	Reason unknown.	Unflagged.
J	Wave-buoy	17.6.79	07°46'N, 50 16'E	20.7.79	13 12'N, 60 00'E	23	137	Abrupt failure. Reason unknown.	No T. Wave data apparently good.

† Data flagged as suspect or bad where appropriate by FGGE Buoy Control Centre, Toulouse, or Marine Environmental Data Service, Canada.

‡ Recovered in New Zealand, August 1980.

example, was washed ashore in Tasmania in November 1980 still transmitting, but the battery voltage had fallen from 13·5 to 7·5 V. With the exception of D, the numbers of position locations per day accorded well with published data.[7] The location accuracy was checked satisfactorily prior to shipment of the buoys and against the ship navigation system shortly before deployment. In one of the initial tests, over several days, the mean position given by the ARGOS system differed by less than 100 metres from the UK Ordnance Survey datum, with a standard deviation (20 samples) in latitude of 300 m and in longitude of 170 m. Throughout the buoy lifetimes data were generally of excellent quality, with only very occasional errors at the start or end of a

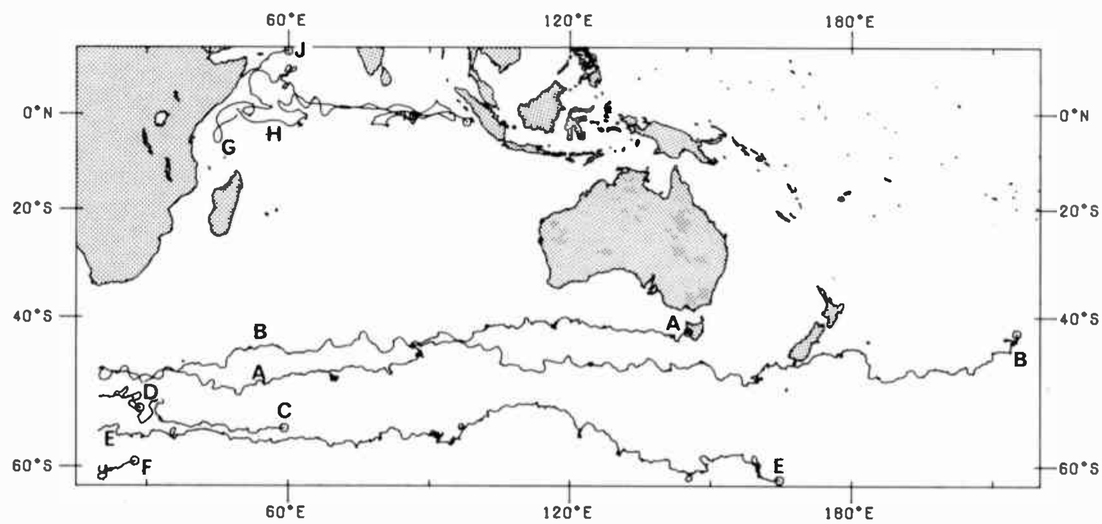

Fig. 3. Drift tracks of buoys.

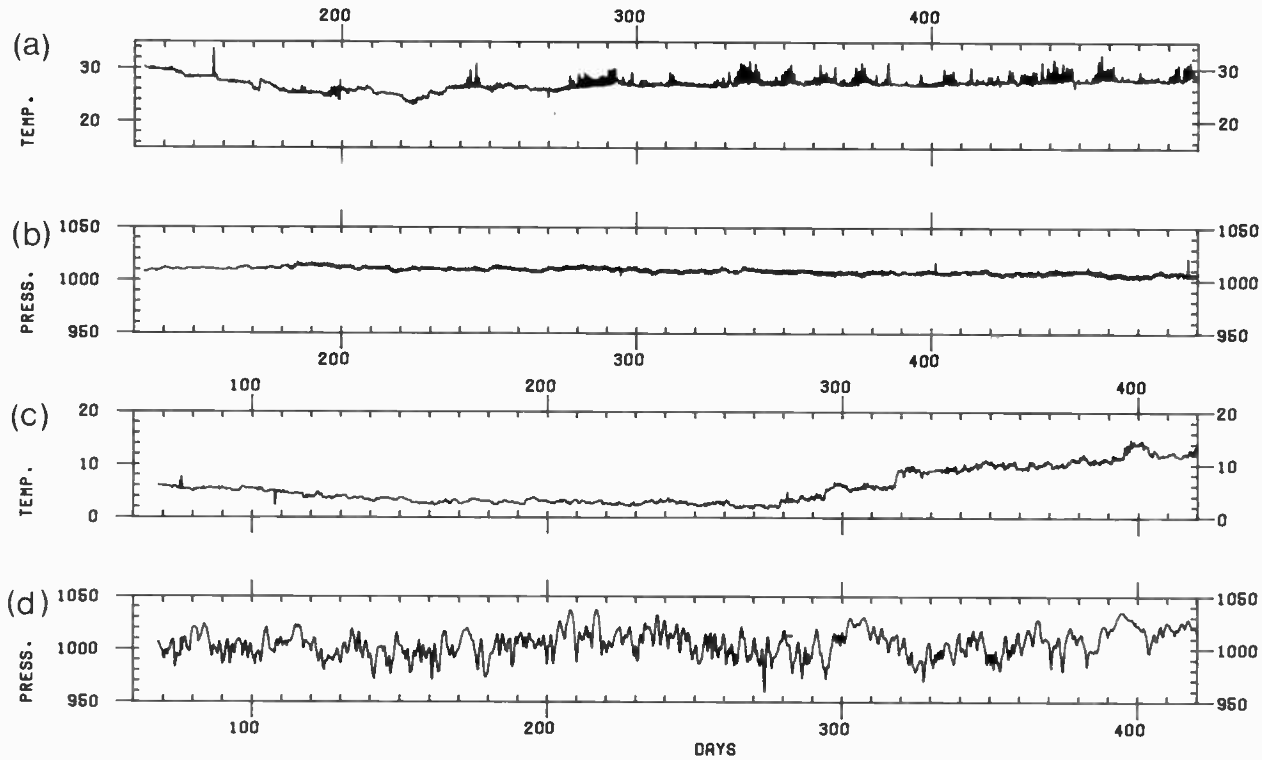

Fig. 4. Temperature (°C) and pressure (mbar) records for buoys G (a), (b) and A (c), (d).

transmission sequence, these due no doubt to low elevations of the satellite at such times.

It is not possible to discuss in detail the considerable data set obtained from the buoys, but pressure and temperature records representative of a year's operation are shown in Fig. 4. The upper traces (a and b) are data from Indian Ocean buoy, G: the lower traces (c and d) from Southern Ocean buoy, A. Very different meteorological conditions obtain in each area, with relatively minor changes taking place in barometric pressure in the Equatorial region, while along the track of buoy A much variability exists at periods of a few days. The thickening evident in trace (b) is accounted for by semi-diurnal fluctuations of $\sim$ 1–2 mbar, namely the thermal atmospheric tide, which reaches a maximum near the Equator. Diurnal fluctuations in temperature also contribute to the 'noisy' appearance of (a); these correlate well with similar increases in the internal buoy temperature (not shown). Much lower temperatures were encountered by buoy A. The slow changes here were partly seasonal and partly due to changes in latitude over the year. The increase in temperature of $\sim 4^\circ$ on day 320 is due to the buoy crossing the Subtropical Convergence.

The accuracy of data from an individual buoy is not immediately obvious in this type of experiment, although there are some possible means of verification. In the first place, a number of comparative observations were made prior to and immediately after buoy deployment using precision sensors with calibrations traceable to standard instruments. Secondly, the Buoy Control Centre, established at Toulouse for the duration of the FGGE in order to monitor and coordinate operations, was able to make some direct comparisons between buoys which drifted together. The BCC also received reports of apparently anomalous behaviour from those users of real time (GTS) data who were able to look at large scales of variability. For sea surface temperatures, for example, the Canadian Marine Experimental Data Service compared five-day-mean temperatures from individual buoys with the monthly climatological mean for 5° squares and were thus able to detect apparent anomalies. The comments in columns of Table 1 are derived mainly

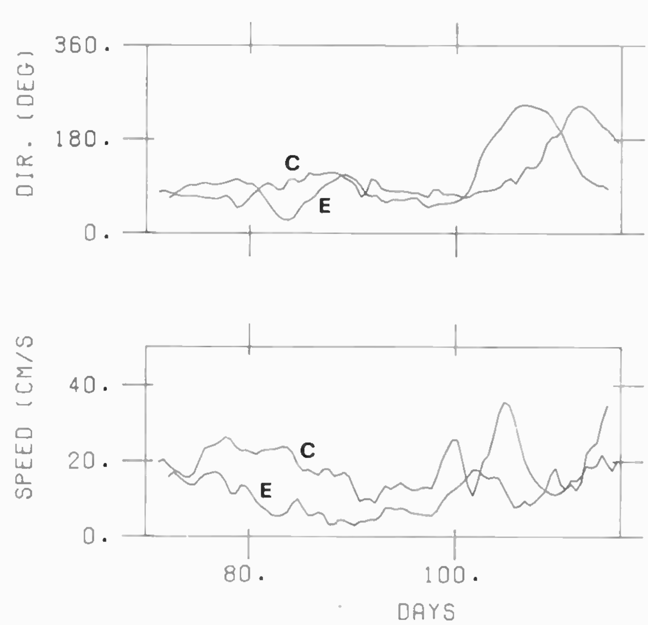

Fig. 5. Speed and direction of buoys C (51 S) and E (56 S).

from data published by the BCC: data for which no adverse reports were received during the FGGE year are unflagged. Often, a study of the self-consistency of the record can reveal erroneous data. However, for adjacent buoys E and F sea temperatures were reported as 3° low within a month of deployment (Table 1) having been verified as correct on deployment. Neither record bears evidence of sudden change or instability and further investigation is clearly necessary in this instance.

Interpretation of buoy drift rate in conjunction with the drogue loss sensor has so far proved difficult, although recent acquisition of the complete FGGE data set may assist in permitting correlation with winds computed from the observed pressure field. The Southern Ocean buoys were deployed along a line of longitude through a zone noted for the predominance of westerly winds. The latitude at which strongest winds occur is variable, at 20°E it occurs at about 48°S. To the south, wind strength decreases, and beyond about 62°S easterly winds prevail. The proximity of buoy F to the transition region accounts for its low drift rate, evident in Fig. 3. No direct indication was received of drogue loss and it seems likely that in this case the drogue survived for the duration of the buoy transmissions. Further north, E, equipped with a 300 m drogue, drifted until Day 149 at a mean speed of 15 cm/s and showed some spatial coherence in its drift (Fig. 5) with C, approximately 300 miles to the north. From the lack of correlation, remarked below, between the drifts of C and D we conclude that until Day 115 the drogued buoys were responding primarily to a large-scale feature in the ocean circulation. At Day 149 the mean speed of E increased abruptly to 40 cm/s and remained at that level for the next 16·5 months. It seems reasonable to suppose that the drogue or its tether failed at this point, although no evidence was supplied by the drogue loss sensor to support this. Similar interpretational difficulties occurred with the Indian Ocean deployments and for the deployments at 48°S.

Although positive indication of drogue loss occurred for one Indian Ocean buoy (G) the very variable and often large currents obtaining during the Monsoon (up to 250 cm/s) precluded correlation with drift rate. In the case of Southern Ocean buoys A and B, launched with drogues at 20 m and 100 m respectively, A moved much more rapidly eastwards at first than B, which was initially trapped in an eddy. Apart from some periods of rapid transport, notably between Days 80 and 140 when a maximum speed > 60 cm/s was recorded, A drifted eastward at a mean speed of ~ 30 cm/s; this seems to have been typical of other buoys, drogued and undrogued, at this latitude. At no time is there any evidence in the rate of drift to indicate drogue loss, although this buoy was recovered in Tasmania with the drogue tether lug completely worn through. Buoy B drifted at speeds < 20 cm/s until Day 100, when it began to exhibit large variations in speed over a few days' period, not unlike the changes generally observed in barometric pressure in this zone, suggesting that it might have been driven directly by the wind. This behaviour persisted for many months and it seems likely that the 100 m drogue was lost approximately one month after deployment.

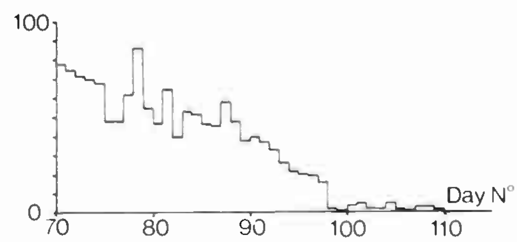

Fig. 6. Buoy D transmissions received per day.

In relation to the standard FGGE configuration buoys, the wave buoys had disappointingly short lives. Although good wave data were obtained from D, the total number of transmissions received per day dropped steadily after the buoy was launched, as shown in Fig. 6. Position fixes were not achieved after Day 95, this being the last day on which an adequate number (5) of consecutive transmissions were received in any transit.

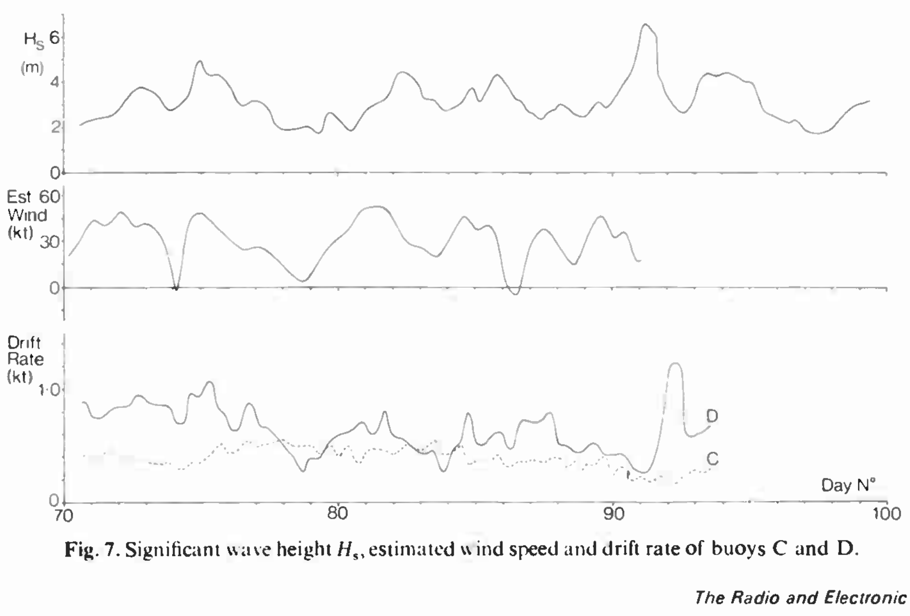

Fig. 7. Significant wave height H_s, estimated wind speed and drift rate of buoys C and D.

This was indicative of progressive reduction in transmitter power or in frequency stability although the transmitter supply voltage was monitored as being steady at 11·7 V up to the time of failure. The buoy was subsequently recovered in New Zealand and examination by the New Zealand Oceanographic Institute showed that sea water had leaked in past the sea surface temperature sensor O-ring, causing damage to the transmitter circuits. The wave data acquired from this buoy are shown in Fig. 7, together with the west wind strength estimated from the North-South atmospheric pressure gradient, derived from buoys A, C, F and E. The buoy drift rate, which clearly is determined mainly by wind speed is also shown. There is quite good correlation between the significant wave height and estimated wind speed initially. Since the wave buoy was drifting east faster than the drogued buoys, the estimated wind speed becomes progressively less valid for the locality of the wave buoy. In addition the pressure gradient measurements were only possible along an approximately N-S line so that any N-S component in the wind was not calculable; however, since the wave buoy drift was predominantly east, neglect of the N-S wind component should not affect the comparison seriously.

6 Conclusion

Although final evaluation and application of the FGGE data set can obviously not be expected for some time, the Southern Hemisphere Drifting Buoy Network—to which the buoys described here contributed—is generally considered to have been very successful. In just one example, use of the meteorological data in near-real time by Australia has had considerable impact in improving analyses of oceanic weather systems and hence has resulted in improvements in forecasts.[8]

Our experience, as recipients of non-real time data, has generally reflected the experience of others, namely that the ARGOS system has functioned extremely well. Provided that the cost to the user of the data acquisition and processing system can be held to reasonable levels, the use of drifting buoys for large-scale data collection has considerable potential. As yet it is in its infancy and further development is clearly required before sensor reliability and longevity can be considered to match the telemetry and location system. We see the more difficult problems arising in the use of drogued buoys as reliable indicators of Lagrangian drift; particularly with regard to drogue integrity in the long term and drogue loss indication, whose reliability should, ideally, exceed that of the drogue.

A more fundamental question also arises of the extent to which the buoy will remain 'locked' to the local water mass when it encounters current shear, wind forcing, and waves. Each of these aspects requires further attention. A vector averaging current sensor is under development; its outputs will be telemetered via the satellites. Measurement of current relative to the buoy, in addition to buoy displacement over the Earth's surface should result in improved estimates of mean near-surface current along the buoy track.

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The potential application of analogue matched and adaptive filters in spread-spectrum communications

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Based on a paper presented at the IERE Conference on Digital Processing of Signals in Communications held at Loughborough in April 1981.

SUMMARY

Spread-spectrum communication techniques use a common wideband channel to obtain subscriber accessing in secure, jam resistant links. Phase shift keying (p.s.k.) and frequency hopping (f.h.) under the control of a pseudo-noise (p.n.) code are favoured signal waveforms. The paper reviews the performance of surface acoustic wave and charge-coupled device fixed coded and electronically programmable analogue matched filters for the detection of short (13 to 1023 chip) 1 kHz to 20 MHz chip (clock) rate p.n.-p.s.k. coded signals. The extension of these matched filters to accommodate the longer codes (10 000 chips) and integration times (several milliseconds) used in current spread-spectrum systems is reviewed, and their importance in reducing synchronization acquisition is highlighted.

Surface acoustic wave device approaches to synthesize and detect the alternative f.h. spread-spectrum waveforms are also discussed. The realizations of digitally-controlled coherent frequency synthesizers using s.a.w. chirp filters and bandpass filterbanks are highlighted. Finally the paper identifies the problem in many spread-spectrum systems, where the input-signal-to-interference ratio is so low that code matched filtering alone is not sufficient to detect the transmitted data, and it investigates the use of adaptive signal processing techniques for further suppression of wide and narrowband interference.

Editorial Note: The author introduces the term 'chip' for bits of spreading code as this is common practice in spread-spectrum systems. Readers should not confuse it with the more familiar usage to describe an individual microelectronic integrated circuit die.

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1 Introduction to Principles of Spread Spectrum

A spread spectrum (s.s.) communication link,[1,2] (Fig. 1) employs a transmitted signal bandwidth many times larger than the digitized information bandwidth in order to achieve some or all of: anti-jam capability; multipath rejection; low detectability; and an accurate ranging capability. Each transmitter incorporates a wideband (1 to 500 MHz) synthesizer, normally at i.f., which spreads narrowband data (300–2400 baud) from the subscriber into the common communication channel. The remote receiver employs an identical synthesizer, synchronized to the incoming signal, and it integrates over the bit interval to demodulate the wideband transmission for information recovery. Interference protection is provided by the spreading ratio, that is, transmitted signal bandwidth divided by the encoded data bandwidth. In order to distinguish between the wideband spreading code and the narrowband data the bits of spreading code are called chips so that the term bits applies only to the narrowband traffic.

Some examples of systems employing spread-spectrum techniques are the JAGUAR-V[3] land mobile communications system, the ground and airborne Joint Tactical Information Distribution System (JTIDS),[4] the SKYNET[5] satellite communication and Navstar Global Navigation System[6] (Table 1). Another example is the Packet Radio System.[7] These systems are further described in this paper highlighting their associated spread-spectrum coding techniques and the potential impact of analogue signal processors. In general these systems do not offer the same channel utilization or efficiency[57] of conventional frequency division multiple access (f.d.m.a.) and time division multiple access (t.d.m.a.) systems but the excessive bandwidth is used for protection against intentional and unintentional interference.

Table 1

Examples of spread-spectrum systems

Scenario	Frequency allocation	Coding format	System
Ground-ground	v.h.f./u.h.f.	f.h.	JAGUAR SYNCGARS
Ground-air-ground	L band	d.s./f.h.	JTIDS
Ground-satellite-ground	L band	d.s.	SKYNET Navstar/GPS TDRSS

There are two basic types of spread-spectrum signal transmissions. Subscribers may transmit *repetitively* when the repeat period of the s.s. synthesizer is several milliseconds or seconds. Here, the correlation properties (orthogonality) of the s.s. synthesizer waveforms must be used to separate individual subscriber transmissions. This uncoordinated code division multiple access (c.d.m.a.) is attractive for large geographic coverage systems, such as those employing a satellite repeater. These are normally referred to as direct sequence spread-spectrum systems. Alternatively, subscribers may transmit information in short *bursts* when accessing may

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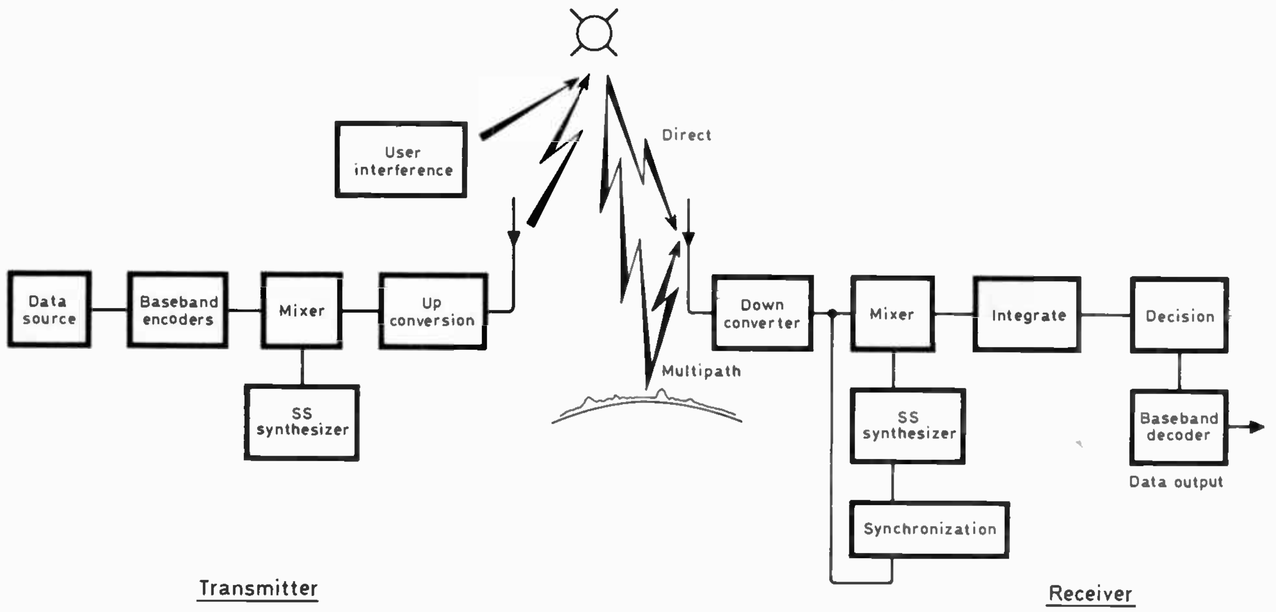


Fig. 1. Typical satellite repeater spread-spectrum system.

either be uncoordinated or conventional t.d.m.a. techniques may be employed, dependent on the precise operational requirements.

Two basic techniques, phase shift keying (p.s.k.)—biphase or quadriphase—and frequency hopping (f.h.), both under the control of a pseudo-noise (p.n.) code,[8] are favoured for the s.s. synthesizer waveforms (Fig. 2). Current equipments predominantly use p.n. p.s.k. s.s. waveforms due to the ready availability of wideband phase modulators. However, coherent p.s.k. or minimum shift keying (m.s.k.)[9] is gaining in popularity as its spectral properties make better use of the communications channel. With p.s.k. or m.s.k. the transmitted r.f. bandwidth is controlled directly by the p.n. code chip (clock) rate. With f.h. techniques the s.s. communication system can be designed, in principle, with any desired number of frequency slots, several hundred being typical. Individual frequencies are selected in the synthesizer by detecting several adjacent chips of code and decoding the word to identify the transmit frequency. Figure 2 shows a trivial example where one of four frequencies is decoded by selecting code chips in pairs. Orthogonality between adjacent frequency slots is ensured if the dwell time on each frequency equals the reciprocal of the slot separation. Frequency hopping offers much flatter transmitted spectrum than p.s.k. or m.s.k. and for a given r.f. bandwidth, the f.h. dwell time is much longer than the equivalent chip time when p.s.k. techniques are used, which eases synchronization acquisition in the f.h. system. For slow f.h. systems coherence between hops is not required but in systems which hop within the bit interval the f.h. synthesizer must be both fast switched and phase coherent from hop to hop. Other coding techniques are time hopping, and hybrids, e.g. p.s.k./f.h. systems. The linear frequency modulated (chirp) techniques used in radar waveform coding[10] have not received significant application in s.s. communications.

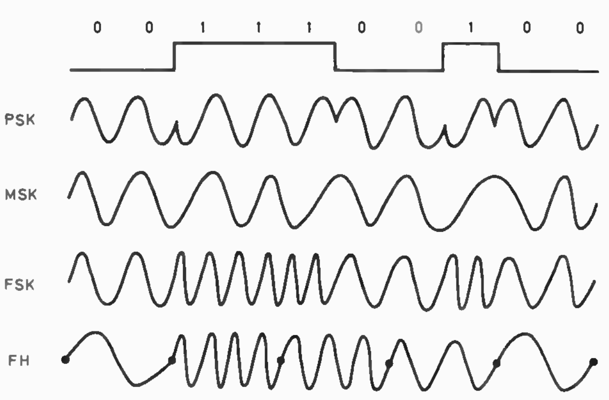


Fig. 2. Examples of spread-spectrum modulation techniques.

This paper surveys the potential application of surface acoustic wave (s.a.w.)[11,12,13] and charge-coupled device (c.c.d.)[14,15] matched filters for generating and detecting these s.s. signal waveforms. As i.f. processors the s.a.w. devices are initially very attractive but they can only handle intrinsically short duration, e.g. <50 μs, waveforms. Hence, they are not capable of directly processing the millisecond and longer waveforms employed in typical repetitive transmission systems. However, it has been shown that device cascading,[16] signal recirculation[17,18] and input correlation[19] techniques now permit s.a.w. devices to process the signals transmitted in these systems. In comparison the baseband c.c.d. requires complex I and Q demodulation but this disadvantage is offset by the ease with which the code and chip rate can be altered.

The paper is laid out by first discussing, in Section 2, the uncoordinated, continuous transmission, direct sequence s.s. system and comparing it with a coordinated t.d.m.a. burst transmission system. Section 3 reports the recent development of s.a.w. and c.c.d. matched filters and Section 4 discusses f.h. synthesizers and investigates their possible impact on s.s. signal processing. Section 5 summarizes recent developments in adaptive signal processors for use in adaptive antennas and adaptive transversal filters which are used in the suppression of wide and narrowband interference respectively.

2 Spread-spectrum Communications

2.1 Direct Sequence Systems

As stated in Section 1, a common s.s. system employing repetitive transmissions is the direct sequence spread-spectrum or direct p.n. system,[1] which is illustrated schematically in Fig. 1. One example of such a system is SKYNET.[5,20] In these systems, the transmitter synthesizer generates a long, repetitive p.n. code, typically at 1–10 MHz rate, which is added modulo-2 with the input data (300–2400 baud) prior to i.f. phase modulation. Up-conversion to the microwave carrier is employed before this characteristic signal is transmitted. After relay of all signals through a satellite, the characteristic signal now corrupted by user interference and multipath arrives at a receiving station. The intended receiver contains a similar s.s. wideband synthesizer, a mixer, an integrator and a synchronizer. The receiver synthesizer is set to provide an output identical to that of the transmitter synthesizer and synchronized to include a delay equal to the propagation time. The synthesizer output is mixed with the incoming corrupted characteristic signal and integrated over the information bit period—the active correlation process—which is achieved in today's equipments with micro-electric circuitry. The polarity of the baseband voltage then determines whether an encoded data bit '1' or '0' was transmitted.

User interference and low-angle multipath, due to ground or sea reflections, are discriminated against by the active correlation process. This results in a processing gain or signal-to-noise ratio (s.n.r.) improvement which is proportional to the spreading ratio given by the transmitted signal bandwidth divided by encoded data bandwidth. Processing gain is the prime feature of s.s. communications. An intentional jammer unaware of the characteristic signature must transmit excess power to overcome this s.n.r. movement.

Receiver synchronization is achieved in three stages[20] via carrier phase locking, p.n. code vector timing and bit time synchronization. Code vector timing requires the p.n. code in the receiver synthesizer to be synchronized to the code in the received signal to within half the chip length, i.e. 50 ns if the chip rate is 10 MHz. Synchronization acquisition is achieved by conducting a number of trials where the receiver code timing is changed by about half the chip period between trials. This serial search procedure is a time consuming process, because the input s.n.r. is typically -30 to -40 dB. Hence, integration over 1 ms is necessary for each trial, to obtain a satisfactory output s.n.r. to make a decision. Generally the acquisition time of the active correlator is equal to the receiver time uncertainty multiplied by its processing gain. The key attraction of matched filters is that they require bit synchronization only. A matched filter based synchronizer searches over all the stored chips during each correlation period, improving the acquisition time by a factor equal to its time bandwidth (TB) product.

Other spread-spectrum systems also used the direct sequence technique. The Navstar[6] satellite based Global Positioning (Navigation) System (GPS) uses a continuous code transmission from each satellite in the constellation. The satellites each carry an atomic frequency standard and transmit two L-band spread spectrum signals whose carrier frequency and code epochs are synchronized with the satellite clocks. The two coherent L-band carriers are added modulo-2 with a 10 MHz chip rate precision and a 1 MHz coarse acquisition code plus the navigation data. Thus a user with an accurate knowledge of system time may use the arrival time of the signals from up to six of the satellites positioned overhead to derive a set of range measurements, and hence, knowing the satellite orbits, fix his position.

The Navstar receiver[21] has to perform a number of functions. Initially, it must select the appropriate L-band signal frequency and acquire and maintain synchronization with the transmitted codes. The synchronization control loop must cater for frequency shifts occasioned by vehicle motion. Next, the arrival times of the signals from a number of satellites must be accurately measured either simultaneously by using separate tracking loops for each satellite, or sequentially and repetitively, using the same circuitry for each satellite. This approach would be suitable in a relatively slow-moving vehicle. Once acquired, each satellite is placed in a list which is scanned sequentially by the computer. Re-acquisition of a satellite is a rapid process (typically a few hundred milli-seconds) since the previous measured code phase and Doppler shift is used to pre-set the receiver. Initial results[6] have shown that prototype equipment gives positioned accuracies of 10–50 metres, which is dependent on code chip period. The Navstar system has the added advantage over other navigation systems that it also provides accurate measurement of vehicle velocity.

2.2 Burst Transmission Systems

In comparison with the repetitive direct sequence systems which can be built with active correlators, the burst transmission systems require matched filter detectors as they must synchronize immediately to the short bursts of received signal to decode the transmitted data. One example of such a coordinated access (t.d.m.a.) system is JTIDS.[4] This is intended as an interservice communication, navigation and identification system where information is shared via the t.d.m.a. net, over line-of-sight links.

Users are each allocated a 7·8 ms time slot when they can input information, while reception is possible during all the other time slots. The system operates over the 960–1215 MHz TACAN airborne military navigation band which is already shared by IFF, SSR and DME avionic equipments. Thus it is an attempt to overlay spread-spectrum on an existing system with mutual use of a common frequency band.[58] The 7·8 ms time slot accommodates 255 message bits at $\sim$28·8 kbaud rate.

Message coding is done initially in a Reed–Solomon encoder[22, 23] to provide error detection and correction capability before being spread by a 5 MHz chip rate m.s.k. code. Frequency hopping is then used as the second level of spreading into the 255 MHz allocated band. This is an expensive system, involving a potential mix of s.a.w., l.s.i. and v.l.s.i. or v.h.s.i.c. techniques to implement the equipment. In addition, the t.d.m.a. format requires accurate clocks for each subscriber who uses the system, which are updated to alter transmit times dependent on position and time relative to t.d.m.a. net. References 24 and 25 give examples of very sophisticated s.a.w. based receivers which are applicable in this type of spread spectrum system, which is likely to be incorporated in the next generation of IFF equipments.[61]

The recent developments in bandpass filters based on s.a.w. tapped transversal filters represent an optimum i.f. design approach for many system requirements.[26] In addition to offering operating frequencies at v.h.f./u.h.f., compact construction and sharp out-of-band roll-off, a non-dispersive transversal filter implements a linear phase filter design. Alternatively the introduction of a small amount of dispersion allows the phase response to be controlled independently of the amplitude response which is particularly significant when designing filters to meet a specified group delay response.

Surface acoustic wave bandpass filters have recently been used to upgrade the capabilities of spread-spectrum systems. Conventional phase shift keyed (p.s.k.) signals result in considerable spillage of energy into the adjacent and nearby channels, since the transmitted spectrum possesses a sinc ($(\sin x)/x$) function response. The alternative m.s.k. waveforms[9] have no distinct phase transitions (Fig. 2), which reduces the power radiated in the adjacent channel from -13 dB to -26 dB.[27]

Until recently, the generation of these waveforms required the synthesis and combination of two time-interlaced p.s.k. signals. It has recently been demonstrated[27] that the arbitrary bandshape of the s.a.w. filter makes it possible to transform a p.s.k. signal into m.s.k. with a single filtering operation. In addition, the adjacent channel spillage can be suppressed beyond -26 dB by further modifying the filter bandshape.

Flexibility of s.a.w. designs also makes it attractive to use a matched s.a.w. filter to optimize the signal-to-noise ratio in the m.s.k. receiver. This development is not merely altering system specifications for spread-spectrum programs, such as JTIDS from p.s.k. to m.s.k. waveforms; it is also creating a demand for s.a.w.-filtered m.s.k. with its ultra-low adjacent channel interference.

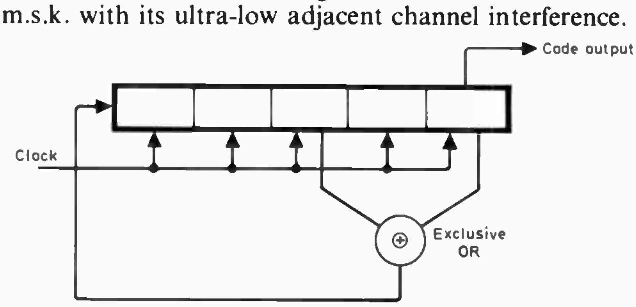


Fig. 3. Generation of 31 chip pseudo noise (p.n.) coded maximal length (M) sequence.

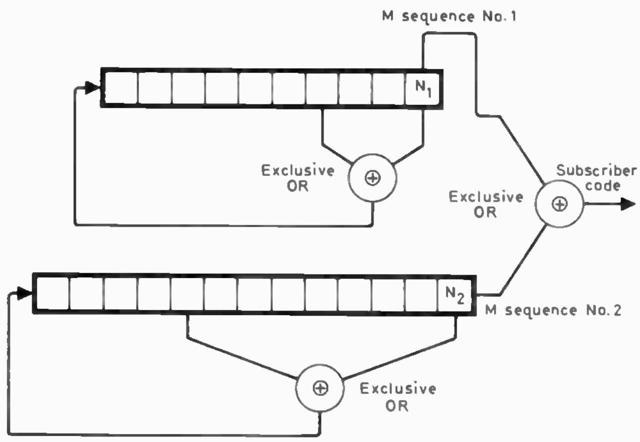


Fig. 4. Combination of 2 individual M sequences to derive a new coded sequence for use in code division multiple access systems.

2.3 Code Design for Spread-spectrum

The codes normally employed in spread-spectrum are p.n. maximal length (M) sequences.[8] Several distinct codes exist for each M-sequence of length $2^N - 1$, where N is an integer. Figure 3 shows how they can be generated by an N-stage digital shift register an appropriate feedback. The unique codes selected for use in a direct sequence system are obtained either as disjoint subsequences from a very long M-sequence or by adding, with an exclusive or gate, two separate M-sequences (Fig. 4), to obtain a Gold code.[28] In this latter case subscribers can use different relative delays between two common M-sequences, to obtain codes with auto- and cross-correlation properties which are appropriate for an uncoordinated c.d.m.a. system.

The repetitive codes used in direct sequence systems give an autocorrelation function which exhibits a sharp peak, when the codes are accurately aligned, with a set of time sidelobes whose peaks are of amplitude $1/N$ relative to the correlation peak, where N is the number of code chips in the correlator, i.e. its TB product. In comparison, burst-mode transmission also has an identical correlation peak but the sidelobes increase to relative amplitude $2/\sqrt{N}$, as the burst of code enters and leaves the matched filter (see simulations in Fig. 7). Asynchronous burst transmission systems, such as those developed for ranging,[30] dictate that both the aperiodic autocorrelation and cross-correlation functions exhibit low time sidelobes. Such performance cannot easily be synthesized from the known periodic autocorrelation performance of, for example, Gold codes and it requires extensive computer searches of comma-free codes.[29] These have shown that the best approach is to select disjoint subsequences from a longer M sequence.[30] Other codes of significance in spread-spectrum are the Barker codes[31] (up to 13 chips long) which are used for synchronization and the Reed–Solomon[22] codes which are used for deriving f.h. patterns.[59]

3 Matched Filters for Phase-coded Waveforms

To perform the optimum predetection processing of a signal, $s(t)$, corrupted by Gaussian white noise, the receiver must incorporate a matched filter. For a phase

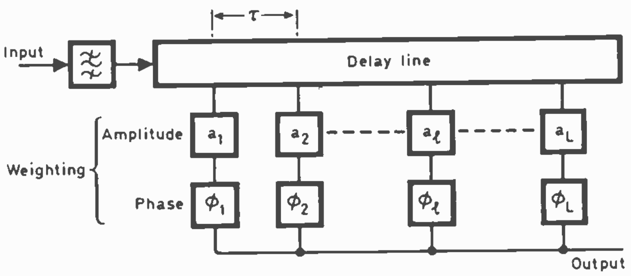


Fig. 5. Canonical form of transversal filter.

shift keyed (p.s.k.) or on/off keyed i.f. signal the matched filter can be realized with a transversal filter (Fig. 5) comprising a lossless, tapped delay line. Matching to a biphase coded signal requires each tap to be spaced apart by a delay, τ, equal to the modulation (chip) period, with equal amplitude weighting and $+1$ or -1, i.e. 0 or π tap phase weighting.

The digital equivalent of the transversal filter[32] would require the input to be demodulated to baseband in I and Q channels, and four real digital tapped delay lines are required to handle complex input signals. In addition the input must be converted to digital words in an a.d.c. and the shift registers expanded to 8–12 bits width. As the reference waveform is bipolar, multipliers are not required as they can be replaced by logic gates to complement the outputs when required. However, this realization is much more complex than the active correlator and hence has received little emphasis to date. For radar signals, which use linear frequency modulated chirp waveform coding, the reference weights shown in Fig. 5 are complex and it is preferable to realize the digital matched filter with Fast Fourier Transform (FFT) techniques[33] as this requires considerably fewer multiplier operations, giving a greater bandwidth capability.

3.1 C.C.D. Matched Filters

In comparison the sampled analogue c.c.d. programmable matched filter[14] is gaining emphasis for spread spectrum applications.[15] The c.c.d. operates by temporarily storing a quantity of charge, representing the analogue signal, in a potential well created by pulsing a metal-oxide-semiconductor capacitor. The transfer of charge between capacitors is achieved by clocking respective electrodes in a linear array. The importance of the *programmable* transversal filter is well reflected in a wide variety of schemes which have been proposed to realize it.[14] No single realization is suited to all applications, each striking a definite compromise between bandwidth, power dissipation, accuracy and size.

A specific advantage of the use of purely analogue techniques is the feasibility of direct and analogue–analogue multiplication using one or two m.o.s. transistors which is much simpler than the equivalent digital multiplier. The spatial economy of these elements allows the integration of dense parallel form processors with corresponding bandwidth advantages. Furthermore analogue coefficient storage is implied, and m.o.s. capacitors offer a compact solution. Figure 6 shows the periodic autocorrelation response of a 256-tap c.c.d. programmable correlator when matched to a 255-chip p.n. sequence. The upper trace shows the input bipolar signal and the lower trace the c.c.d. autocorrelation response, which has a measured peak to sidelobe ratio of >35 dB (theoretical $= 48$ dB). This result was obtained with a full analogue–analogue correlator, which can be matched to any amplitude- or phase-modulated waveform and hence is an overdesign for many spread spectrum applications. The consequent penalty is that its maximum bandwidth is in the range 100–500 kHz which is well suited to sonar applications but is insufficient for spread spectrum.

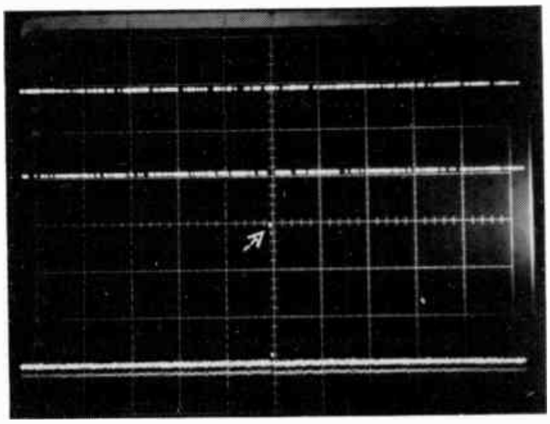


Fig. 6. Lower trace shows autocorrelation response of c.c.d. matched filter (with correlation peak arrowed) when receiving the periodic 255 chip p.n. coded signal shown in the upper trace. (a) Input signal (b) autocorrelation response (horizontal scales 2·5 ms/div, vertical scales linear). (C.c.d. developed by Wolfson Microelectronics Institute under contract to Ministry of Defence).

Charge-coupled device filters designed for matching to the p.n. codes used in spread-spectrum can be implemented with an analogue-binary design. Integrated devices have been reported[15] which can handle up to 512-chip waveforms at 8 MHz chip rates. These low-cost modems still require I and Q demodulation, but they have been shown to exhibit only 1 dB degradation from the theoretical peak-to-sidelobe ratio when detecting 512-chip aperiodic coded sequences. The loss in correlator amplitude that is associated with sampling[34] has been evaluated to assess how this relates to the number of c.c.d. delay stages that are employed per correlator point. The c.c.d. approach also compares favourably with one-bit and two-bit quantized digital correlators in terms of chip area and power consumption.[34]

3.2 S.A.W. Matched Filters

Surface acoustic wave i.f. tapped delay line structures can be designed for p.s.k., f.s.k. or linear f.m. waveforms, using the transversal filter approach (Fig. 5). The preference for phase modulation in spread-spectrum has resulted in several designs utilizing this coding technique. Early device designs, incorporating simple co-linear tap arrays with fixed amplitude and 0 or π phase weighting, highlighted severe performance degradations due predominantly to reflections within the tapping

Table 2

Comparison of s.a.w. p.s.k. matched filter capabilities with charge-coupled device and digital matched filter

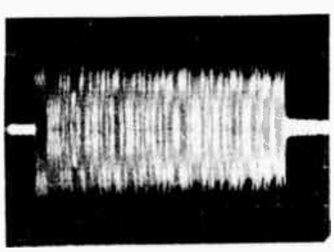
	S.A.W. fixed coded filter	S.A.W. acoustic convolver	S.A.W. semiconductor Convolver		Digital filter	C.C.D.
			Separated	Integrated		
Centre frequency (MHz)	500	500	500	250	baseband processor	baseband processor
Sequence length (μs)	50	60	40	15	arbitrary	arbitrary
Chip rate (MHz)	20	100	150	25	10	10
Time-bandwidth	128	1000	1000	256	256	512
Dynamic range in compression (dB)	60	55	55	45	70	60
Level of spurious responses (dB)	40	40	40	35	50	40
C.w. insertion loss (dB)	40	50	45	45	n/a	n/a

structure. With the alternative reflection-compensated co-linear tap array designs, employing either split electrodes[35] or dual tap geometries,[36] it is possible to overcome these problems and extend these devices to *TB* products of ~ 512 in the laboratory, but 128 is more common in production (Table 2).

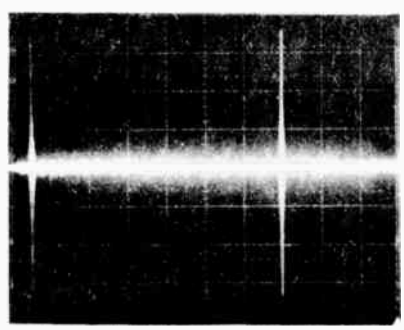
Figure 7 shows the typical burst-mode auto-correlation response for 127-chip p.n.-p.s.k. fixed coded s.a.w. matched filter. Trace (a) shows the impulse response with a mean loss to each tap of ~62 dB and a roll-off of ~2 dB. Trace (b) shows the response to a continuously generated matched signal exhibiting a peak-to-sidelobe ratio of 29 dB in comparison with the theoretical level of 42 dB. Finally, trace (c) shows the response to an aperiodic signal. Here the peak to maximum sidelobe ratio is within 2 dB of theoretical, but the sidelobe pattern does not exhibit the expected symmetry about the peak. Spurious signal levels were measured > −37 dB below the peak.

Surface acoustic wave matched filters have mainly been employed for synchronization in spread-spectrum systems. Many systems use short-length, fixed coded synchronization sequences, such as those based on the Barker codes,[31] and here the fully asynchronous operation of the s.a.w. i.f. matched filter makes it an attractive cost effective component. For systems which use continuously changing codes the s.a.w. matched filter must be electronically programmable. Programmable matched filters,[12] incorporating diode,[37] transistor or m.o.s.f.e.t. integrated tap switches[38] have been developed but there is a preference to use s.a.w. convolvers for this signal processing function, as they offer greater time-bandwidth product (Table 2).

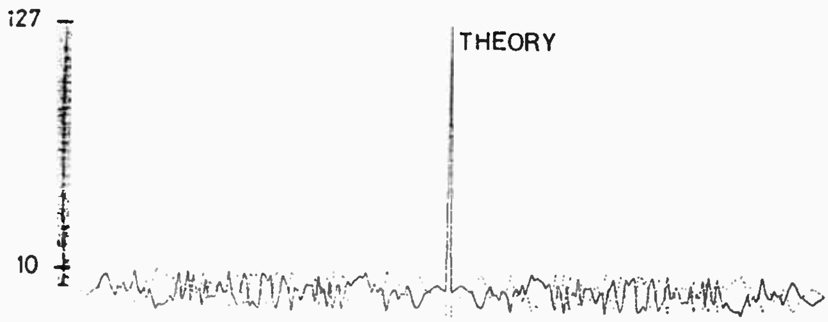
In the simplest device, the degenerate acoustic

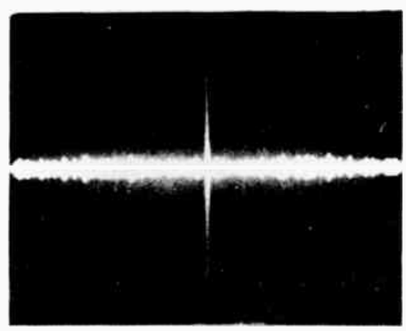


(a)



(b)





(c)

Fig. 7. Surface acoustic wave 127 chip fixed coded matched filter. (a) Impulse response, (b) autocorrelation response with periodic (continuous) input waveform, (c) theoretical and experimental aperiodic autocorrelation response (horizontal scales 2 μs/div, vertical scales linear).

convolver,[39] two input waveforms, the signal and reference, are applied to the two input ports to launch two surface waves towards the centre of the device. The lithium niobate substrate provides a parametric interaction between the two contradirected co-linear s.a.w.s to give the convolution of the two input signals. Thus, the device behaves as a highly programmable linear filter whose impulse response is determined by the reference waveform. Matched filtering is achieved by arranging the reference input to be a time-reversed replica of the signal input. The simple fabrication permits time-bandwidth products of $\sim 10^3$ to be achieved (Table 2). Hence, this matched filter has considerable advantages over s.a.w. microelectronically-switched matched filters. However, the convolver is not intrinsically asynchronous since correlation is obtained only if a reference is applied almost simultaneously with the signal. Also, due to the propagation, the convolver output is time compressed by a factor of two. In addition, the s.a.w. degenerate acoustic convolver is inefficient, but bilinearity factors have been improved from the original $-95\ \mathrm{dBm}^{-1}$ values obtained in early devices to present day values of typically $-70\ \mathrm{dBm}^{-1}$ (Ref. 39).

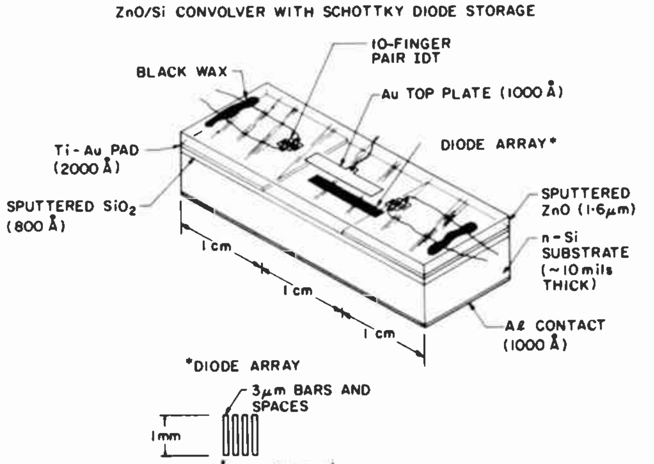


Fig. 8. Schematic of zinc oxide on silicon monolithic storage correlator (courtesy of E. L. Ginzton Laboratory, Stanford University).

Surface acoustic wave convolvers have also been designed using semiconductor layers to sense the parametric interaction. One realization[13] of semiconductor convolver is based on a silicon slice which is closely spaced to the lithium niobate substrate and propagating s.a.w. The alternative integrated realization[19] (Fig. 8), overlays the silicon substrate with a sputter-deposited layer of piezoelectric, e.g. zinc oxide, for generation and detection of the s.a.w. Such semiconductor devices offer improved efficiency at the expense of increased fabrication complexity. In addition, both these convolvers can be modified to store a reference waveform by incorporating diodes into the silicon substrate (Fig. 8), and hence implement an electronically-reprogrammable i.f. matched filter. Recent developments[40] are aimed at reprogramming these devices with a narrowband (slow) code stored in a c.c.d.

For systems which require matched filtering of large *TB* coded waveforms with $>50\ \mu s$ duration, attention has been focused on three approaches. The simplest technique is to cascade devices by connecting several individual matched filters in series through wideband delay lines.[16] The second approach, which offers considerable increase in processing gain, combines one programmable matched filter with a recirculating delay line integrator.[17] This coherently sums the train of small correlation peaks produced by the matched filter. Such a receiver[18] has correlated 2 ms of p.n.-p.s.k. coded signal at a 10 MHz chip rate to give an effective *TB* of 2×10^4. This receiver requires coarse timing information as it must possess frame synchronization to start the recirculation loop. It can be considered as an extension of the existing serial search procedure, except that it accommodates much greater timing ambiguity and hence speeds up acquisition by a factor equal to the *TB* product of the programmable correlator, i.e. 10–100.[18] Figure 9 shows a laboratory prototype module, designed to give a practical demonstration of this processor.

More recently it has been demonstrated that s.a.w. acousto-electric storage correlators can be used in the input correlation mode[19] to correlate long duration waveforms. This permits them to operate like a bank of active correlators, which are simultaneously searching over *N* chips of code, where *N* is equal to the basic convolver time-bandwidth product. When used in the input correlation mode a 10 MHz bandwidth s.a.w. convolver can integrate for >1 ms to give directly 40 dB processing gain required in a spread spectrum system. If in addition the basic device incorporates a $10\ \mu s$

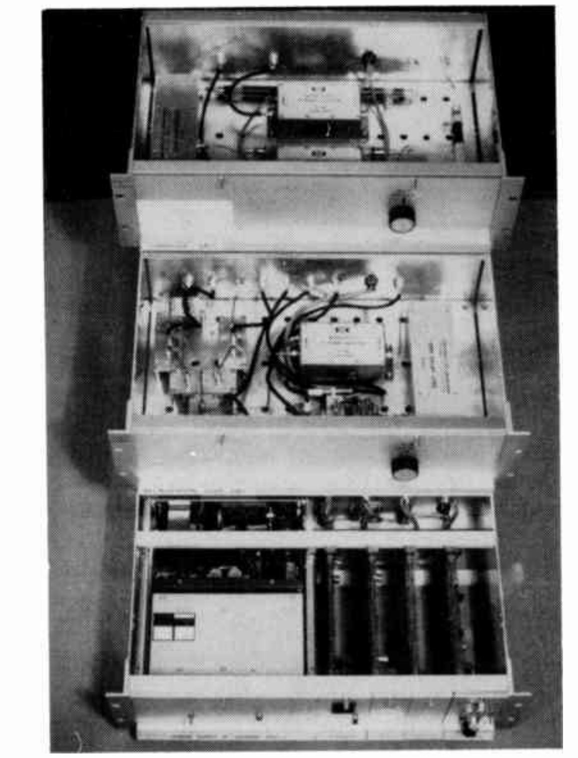


Fig. 9. Large time bandwidth product matched filter module employs s.a.w. acoustic convolver in top box, recirculating delay line integrator in centre box and power supplies and code generators in the lower box.

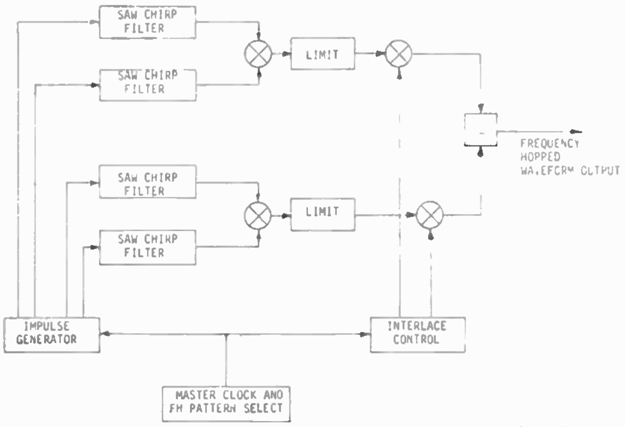


Fig. 10. Schematic of s.a.w. frequency hopped synthesizer based on chirp mixing.

interaction region, i.e. its $TB = 100$, it would offer a 100-fold improvement in acquisition time compared to current active correlators, which is comparable to the earlier modules shown in Fig. 9 but is all performed in a single integrated device. Recent results have been reported[41] on a hybrid convolver/digital module which achieves a similar processing function.

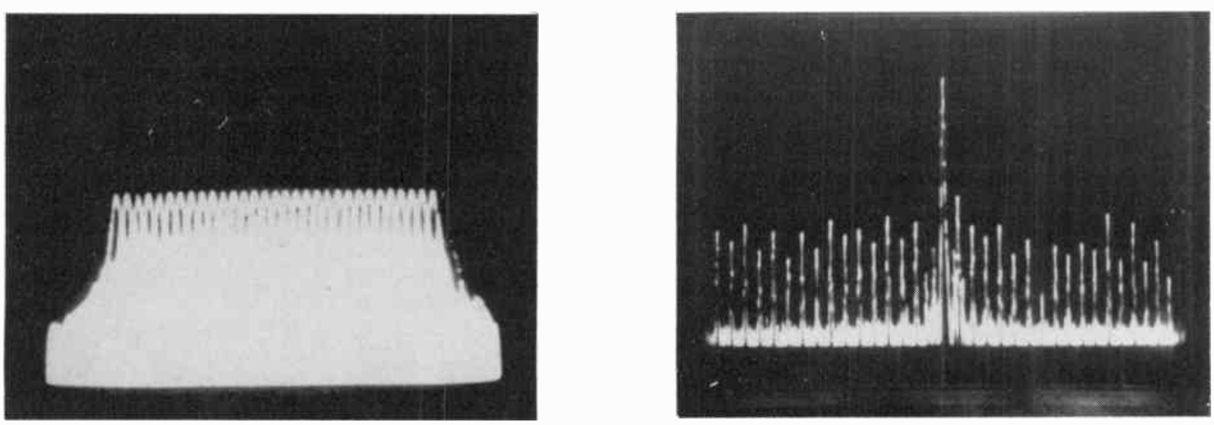


Fig. 11. Spectrum of s.a.w. frequency hopped synthesizer. (a) When (b) hopping randomly between one of 32 available frequencies every 2·9 μs (display centre frequency 127 MHz, horizontal scale 2 MHz/div, vertical scale 10 dB/div); (b) when programmed onto single frequency at 118 MHz (vertical scale 10 dB/div, horizontal scale 1 MHz/div).

4 Frequency Hopped Waveform Synthesis

Frequency hopped waveforms, comprising trains of r.f. pulses where the frequency can be varied rapidly from pulse to pulse, are an alternative modulation technique to p.s.k. for use in a wideband spread spectrum synthesizer. However, the synthesis of rapid coherent f.h. waveforms is technically challenging. Conventional frequency synthesis is based on either 'direct' or 'indirect' methods. Direct synthesis, where the output is obtained by summing harmonically related multiples and sub-multiples of a reference frequency source, provides a fast switched (ns) phase coherent output. However, the requirement for sophisticated filtering makes the equipment bulky and expensive. Indirect synthesis, where the output of a voltage-controlled oscillator is divided and subsequently phase-locked to the reference, is much simpler to implement but the switching times are typically between 100 μs and several milliseconds.

The indirect techniques are being applied to slow f.h. spread-spectrum systems for mobile radio, such as the UK Racal JAGUAR-V[3] radio, the Marconi Scimitar and the US SYNCGARS equipment. These systems, which operate on encrypted delta-modulated speech, hop the transmitted frequency in a pseudo-random pattern over a bandwidth >100 times the modulated bandwidth to gain >20 dB jammer immunity. Tests have shown that satisfactory reception can be achieved when 20% of the radio channels are blocked by jamming. These slow f.h. systems use long dwell times which exceed the bit interval to overcome the necessity for coherence between hops.

The development of s.a.w. technology has revived interest in fast switched, phase coherent, f.h. waveform synthesis. The first s.a.w. f.h. synthesizers were based on sequential impulsing of individual filters within a bank of contiguous s.a.w. bandpass filters.[42] Surface acoustic wave designs ensure that the impulse responses of all filters is of exactly the same duration, permitting one impulse generator to be switched to the appropriate filter to generate the corresponding frequency. This synthesizer realization is very compact and it has low <-60 dB spurious levels, but it is limited both by the number of available frequency slots and by hop duration. An alternative approach is to synthesize coherent f.h. waveforms by routing a pair of impulses into two separate s.a.w. linear f.m. chirp filters.[43] When the impulses are separated by a time $\tau < T$, where T is the time duration of the filter impulse response, the two individual responses partially overlap. If the two waveforms have equal but opposite dispersive slope[44] then they can be mixed and the constant sum frequency selected to yield a pulse of duration $T-\tau$, whose precise frequency is controlled by the filter centre frequencies and the spacing between impulses, τ. Three prototype synthesizers, based on the schematic shown in Fig. 10, have been constructed and evaluated both in the laboratory and in field trials. Figure 11 shows their typical frequency domain outputs in both (a) f.h. and (b) c.w. modes of operation. Spectral interference in (b) can be suppressed to <-60 dB by an output tracking filter. Such equipments can presently synthesize 128 available frequencies programmable over a 52 MHz bandwidth centred at 310 MHz. Typical applications for such a f.h.

spread spectrum system is to provide protection against jamming on a remotely piloted vehicle microwave air-to-ground datalink,[45] carrying bandwidth reduced videotraffic. Prototype laboratory hardware has been designed to implement the schematic shown in Fig. 10. With repackaged devices the hardware can be considerably reduced in size for a final equipment design.

5 Adaptive Processing

There is considerable current interest in adaptive signal processing. At the research level attention is focused primarily on the development of adaptive algorithms, in particular for fast convergence, while commercial firms are actively studying the application of adaptive techniques to three distinct classes of processor. These are the adaptive antenna,[46] the adaptive transversal filter,[47] and noise canceller[47] (Fig. 12). Examples of the range of adaptive processor applications is given in Table 3.

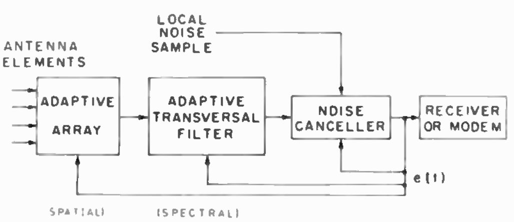


Fig. 12. Schematic of receiver showing location of different adaptive signal processing functions.

In a spread-spectrum receiver all these different classes of processor are significant. The effectiveness of an adaptive array to implement spatial processing to reduce the effectiveness of wideband interference by steering nulls in the beampattern towards interfering sources has been demonstrated.[48] Narrowband interference can be suppressed by spectral processing where a transversal filter impulse response, and hence frequency response, is adjusted to filter or notch out the interference. Noise suppression is accomplished in the canceller by adding in antiphase a separate sample of the noise without signal. In all these processors the adaption algorithm is implemented continuously to provide the best output signal-to-noise-plus-interference ratio under continuously changing input conditions. In terms of complexity the noise canceller is the simplest, as it is a single tap transversal filter with phase and quadrature weights. The adaptive transversal filter is the most complex processor as it typically incorporates 50–200 taps. The adaptive array usually falls somewhere between these two cases.

5.1 Adaptive Antennas

Adaptive antennas fall naturally into two sub-classes. One is the fully-phased adaptive array[48] where each individual antenna element is connected to a multiplier to weight the received signals prior to summation at the array output. Such an arrangement can be used for simultaneous beamforming and null steering to maximize the desired signal and minimize the interference. The other class of adaptive antennas are sidelobe cancellers,[49] which are incorporated alongside the high-gain, mechanically-rotated, dish antenna used in radar, to minimize the reception of interference through the antenna sidelobes.

Much of the adaptive antenna work to date has been concentrated on military spread-spectrum applications. Systems such as JTIDS and Navstar are both dependent on fully-phased adaptive antennas. In Navstar the received signals are of low strength and require narrowband tracking filters to obtain satisfactory s.n.r. In addition their antenna designs are not optimum as they are forced to mount them on the exterior structure of the airframe. Hence the use of adaptive array techniques is advantageous for maximizing the signal to interference ratio in these equipments.

5.2 Adaptive Filters

In the most common adaptive filter, which is based on Widrow's Least Mean Square (LMS) algorithm,[47] the received signal is input to a non-recursive transversal filter. The tapped, weighted and summed output is then compared against a delayed conditioning signal, which represents the desired output from the filter. The resulting error signal is then used to derive updated tap weights. In the LMS algorithm the error is multiplied by a convergence coefficient and cross-correlated with samples of the input signal to give the updates to the tap weights. Figure 13 shows how this closed-loop process can be implemented in parallel for each tap in the delay line. The algorithm requires typically 100 to >1000 of these iterations to reach convergence. This requirement for tap weight updating at the full processor bandwith has been overcome in practical developments[50] where the output of the individual cross-correlators are integrated and the output is used to update the taps less

Table 3

Some examples of adaptive processor applications

	Processor operation	Applications
Adaptive filter	Echo cancellation	TV ghost or multipath suppression
	Equalization	Telephony and data modems
	Cancellation	Separation of wide and narrowband signals and cockpit noise reduction
Adaptive antenna	Coherent sidelobe cancellation	Null steering
	Adaptive phase array	Beamforming and null steering

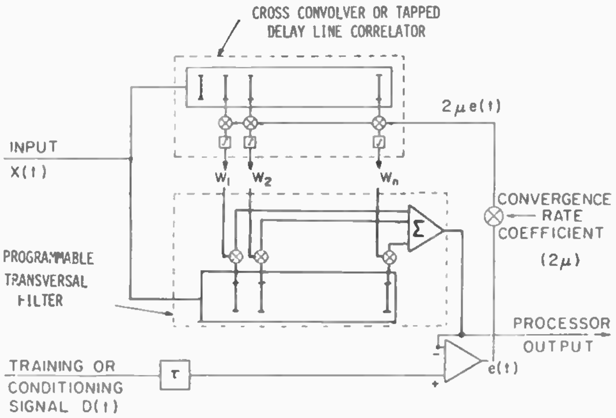


Fig. 13. Adaptive transversal filter based on least mean squares control algorithm.

frequently. In another development, the clipped LMS algorithm, the hardware is simplified by limiting or clipping the sampled input signal to the cross correlator. However, this usually degrades the convergence rate.

The significant signal processing function which the adaptive filter can perform, the separation of a wideband and narrowband signal, is very important in spread-spectrum systems, which are subject to high-level interference from both narrowband jammers and existing a.m./f.m. communication systems. Thus there is considerable interest in implementing a real time adaptive notch filter which can be automatically tuned over the 1–50 MHz channel bandwidth to remove the interference (Fig. 15). Due to the required information, bandwidth, e.g. 1 kHz, and limited available spectral allocation, e.g. 10 MHz, the system processing gain is typically restricted to <40 dB. If the receiver has to operate in an environment with high interference, e.g. interference-to-signal ratios of 60 dB, then the signal cannot be recovered in the receiver matched filter. Thus adaptive filtering must be performed on the input to reduce the amplitude of the interfering signal. When this filter is incorporated it has the benefit of reducing the dynamic range requirement on the remainder of the receiver which follows it, as only the adaptive filter itself has to possess a dynamic range suitable for both the high level interference and low level spread spectrum signals.

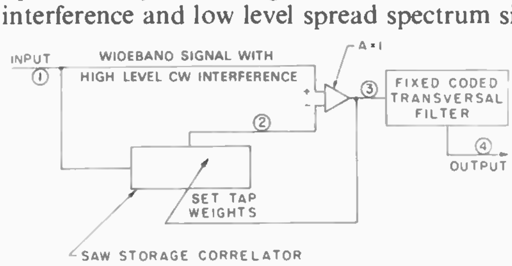


Fig. 14. Adaptive spread-spectrum receiver.

In contrast to digital l.s.i. implementations which are costly and power hungry there is interest in applying analogue c.c.d. and s.a.w. components for adaptive filtering. Charge-coupled devices are attractive for inverse filtering and equalization on narrowband speech waveform. In telephony researchers are actively trying to reduce the distorting echoes caused by the two-to-four line transformer with a c.c.d. or digital adaptive filter.[60]

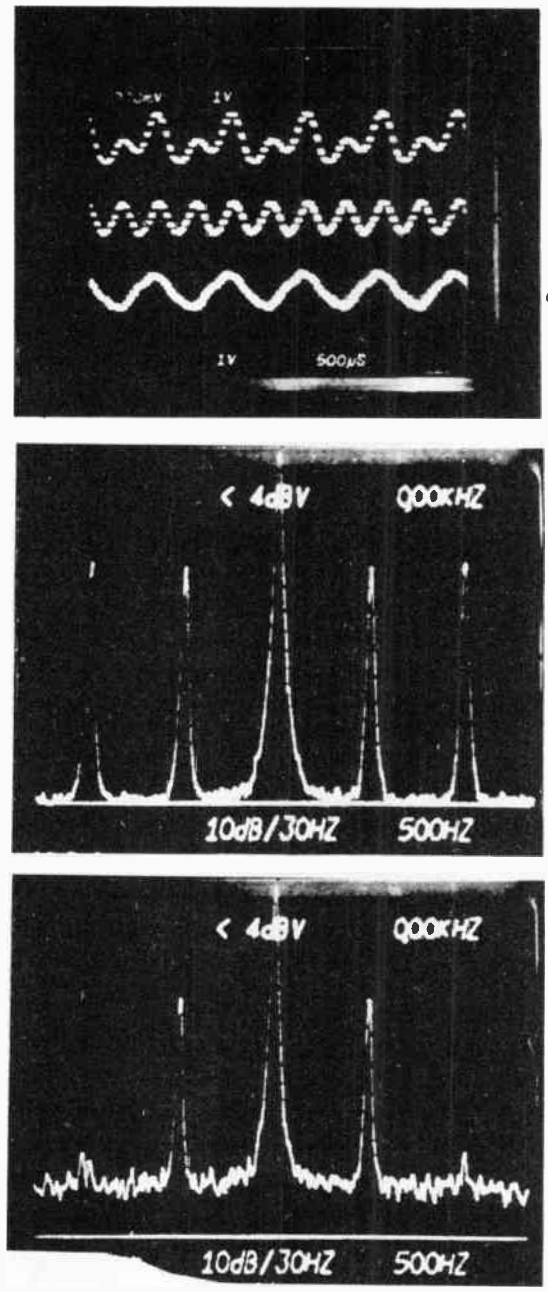


Fig. 15. Demonstration of use of c.c.d. adaptive filter for harmonic cancellation. Shows operation of a clipped LMS c.c.d. adaptive filter where the second harmonic component present in trace (a) is suppressed from trace (c) after processing in an adaptive filter. Traces (d) and (e) show spectrum of (a) and (c) respectively with cancellation exceeding 30 dB after adaptive filtering.

Figure 15 shows results obtained in an 8-tap clipped LMS parallel update c.c.d. adaptive filter when implementing a canceller for suppression of the second-harmonic component of a signal.[51] For this test, the filter was set up with a fixed c.c.d. delay line with four stages of delay between the desired (training) signal input and the difference amplifer (Fig. 13). The sampling frequency was selected as 10 kHz (the maximum for this filter is 100 kHz). The desired signal (Fig. 15(a)), comprised a sinewave fundamental of 1 kHz with a second-harmonic signal but at different amplitude and phase. The filter corrects the phase and amplitude of the interfering harmonic at the input to correspond to the harmonic component of the delayed desired signal (Fig. 15(b)), and the two are subtracted, leaving at the error output only the fundamental sinewave (Fig. 15(c)).

Trace (d) shows the spectrum of the desired signal at the input, and trace (e) the error output spectrum. The spectra show that the second harmonic is suppressed by 32 dB at the output. This result has now been extended to cancellation depths of >50 dB in larger (64 or 256 tap) c.c.d. adaptive filters,[52] and a monolithic c.c.d. adaptive filter design is currently in development.

Although attractive, these c.c.d. adaptive filters cannot be extended to the megahertz signal bandwidths required in spread-spectrum. Here it is attractive to use s.a.w. realizations of the transversal filter. Emphasis on s.a.w. adaptive signal processors has been primarily concentrated on suppressing unwanted c.w. signals. In one demonstration, an analogue wideband (1–20 MHz) s.a.w. programmable storage correlator[53] was used to store directly a sample of the input signal and implement a narrowband filter, which was subsequently used to null out c.w. interference. As the narrowband c.w. interference is normally considerably larger than the wideband coded signal, the stored waveform is a burst of c.w. whose duration equals the delay in the device. In this way it implements a filter with a narrowband $(\sin x)/x$ frequency response automatically set on the interfering frequency. When incorporated as in Fig. 14 the s.a.w. device filters out only the interfering signal at point 2, which is subsequently amplified to overcome the loss in the filter and added in antiphase with the input signal. This implements a wideband filter with a notch centred on the interference frequency. Provided that the notch is narrow with respect to the wideband coded signal, then there is minimal distortion in the wideband signal. After interference cancellation, the coded signal can be detected in the receiver matched filter.

Figure 16, trace (a), shows the signal input to this s.a.w.-based adaptive processor. It comprises a +10 dBm 120·9 MHz c.w. signal plus a −20 dBm p.n. coded waveform modulated at 5 MHz rate onto a 120 MHz carrier. The direct correlation of this composite signal in the 31-chip matched filter, without the adaptive processor, is shown in trace (b). The adaptive processor stores a 4 μs sample of the c.w. signal, and sums the amplified filtered output with the input after ensuring that the signals are in antiphase. This implemented a flat amplitude characteristic filter with a 30 dB deep, 250 kHz wide notch at the interference frequency. A comparison between the input signal trace (a) and the cancelled output signal trace (c) clearly shows the 30 dB suppression of the 120·9 MHz interference, leaving the wideband coded signal. The adaptive interference canceller permitted the correlation peak from the final matched filter (Fig. 14) to be optimized. It can simultaneously notch out one or more interfering frequencies anywhere within the 8 MHz bandwidth of the s.a.w. storage correlator.

This demonstration implemented an adaptive filter without the LMS feedback algorithm. Its notch depth was therefore constrained by component drift. Closed-loop adaptive filters have been reported which perform both the processing functions shown in Fig. 13 (transversal filtering and complex tap weight update and storage) within a single storage correlator, such as that shown schematically in Fig. 8. When used for echo suppression this sophisticated processor[54] gave 15 dB attenuation after only 10 iterations of the LMS based adaptive filter which was accomplished in 200 μs. Other advanced processors[50] have been realized with 200 tap s.a.w. delay lines incorporating complex weighting which can adaptively insert 50–55 dB deep, 10 kHz wide notches anywhere over a 2·5 MHz processor bandwidth. They operate with a conditioning signal, comprising either noise or randomly modulated carrier, whose spectral power density approximates the received signal. Thus they are capable of simultaneously adapting to the spectral characteristics of the received signal and notching out undesired narrowband interference. Adaptive notch filters can also be designed by cascading two Fourier transform processors through a signal limiter[55] or gate circuit.[56] When implemented with s.a.w. chirp transform techniques these open loop processors can automatically notch out interference and provide 30 dB suppression over >50 MHz bandwidth.[55]

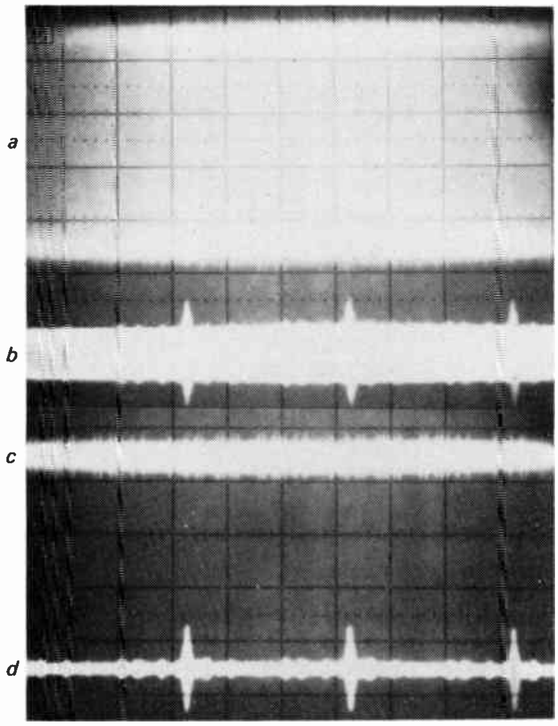


Fig. 16. Surface acoustic wave adaptive filter based on storage correlation. (a) Shows input waveform comprising a c.w. plus a 31-chip p.n.-p.s.k. coded sequence at −20 dB, both on v.h.f. carriers; (b) correlation of (a) in s.a.w. matched filter; (c) input waveform after adaptive processing to cancel out c.w. interference in s.a.w. storage correlator as shown in Fig. 14; (d) correlation of (c) in the s.a.w. matched filter.

6 Conclusions

This paper has reported several potential applications for analogue signal processors in spread spectrum equipments. Although s.a.w. matched filters initially appeared to be significant in the early 70s it is only now in the 80s that they are finding applications in these systems. The main reason for this delay is that the initial successes with fixed coded devices could not be easily extended into programmable matched filters. However, for new programmes which require wideband >20 MHz devices there is a tendency to utilize s.a.w. acoustic convolvers for programmable correlators and not adopt the more elegant but expensive acoustoelectric convolvers and correlators. Charge-coupled devices have also matured over this decade and as micro-electronic integrated circuits they are potentially much cheaper than s.a.w. convolvers and are thus likely to be most widely used in future moderate bandwidth (<20 MHz) spread-spectrum systems. The key point is that systems engineers have evaluated these analogue components and they are confident that the low power consumption makes them more attractive than current digital approaches for spread-spectrum applications. However, in the future the v.l.s.i. and v.h.s.i.c. programmes will provide strong competition with analogue techniques.

For narrowband (<1 MHz) adaptive filters the c.c.d. and digital approaches are in direct competition with c.c.d.s possessing attractive low power operation. For wideband applications s.a.w. approaches are presently complicated and expensive but if the storage correlator is developed into a commercially-available component then it has an excellent future potential as both a programmable matched and adaptive filter.

7 Acknowledgments

I wish to acknowledge the assistance of present and former colleagues in the Department of Electrical Engineering and Wolfson Microelectronic Institute, University of Edinburgh and the E. L. Ginzton Laboratory, Stanford University, California, in the preparation of this manuscript. In particular, the contributions of J. E. Bowers, J. H. Collins, B. J. Darby, J. M. Hannah, G. S. Kino, D. P. Morgan, E. W. Patterson, J. R. C. Reid, and H. S. Tuan are gratefully appreciated. Former sponsorship of the British Science and Engineering Research Council is also acknowledged.

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Peter M. Grant received the B.Sc. degree in electronic engineering from the Heriot-Watt University, Edinburgh, Scotland, in 1966 and the Ph.D. degree from the University of Edinburgh in 1975. From 1966 to 1970 he worked as a Development Engineer with the Plessey Company at both the Allen Clark Research Centre and West Leigh, Havant, designing frequency synthesizers and standards for mobile military communications. Follow a year as Senior M.O.S. Applications Engineer with Emihus Microcomponents, Glenrothes, he was appointed to a Research Fellowship at the University of Edinburgh to study the applications of surface acoustic wave and charge-coupled devices in communication systems. In 1976 he was appointed to a lectureship, teaching electronic circuits, communications and systems. During the academic year 1977–78 he was the recipient of a James Caird Travelling Scholarship and as a Visiting Assistant Professor he researched in acoustic imaging at the Ginzton Laboratory, Stanford University, Stanford, Calif.

Dr Grant serves as one of the honorary editors of the Proceedings of the Institution of Electrical Engineers, Part F, Communications Radar and Signal Processing. He has twice been awarded the IERE Bulgin Premium for papers published in *The Radio and Electronic Engineer* in 1974 and 1977.

Conferences, Courses and Exhibitions, 1982-83

The date and page references in italics at the end of an item are to issues of *The Radio and Electronic Engineer (REE)* or *The Electronics Engineer (EE)* in which fuller notices have been published.
The symbol ★ indicates that the IERE has organized the event.
The symbol ● indicates that the IERE is a participating body.
An asterisk * indicates a new item or information which has been amended since the previous issue.

Further information should be obtained from the addresses given.

MAY

Insulation *10th to 13th May*
Fourth International Conference organized by British Electrical & Allied Manufacturers Association in association with the EEA to be held at the Brighton Metropole Hotel. Information: BEAMA Publicity Department, 8 Leicester Street, London WC2H 7BN. (Tel. 01-437 0678)

Microcomputers *11th to 13th May*
Microcomputer Show, organized by Online Conferences, to be held at the Wembley Conference Centre. Information: Online Conferences, Argyle House, Joel Street, Northwood Hill, Middlesex HA6 1TS. (Tel. (09274) 282211.

Condition Monitoring Systems *12th May*
One day Symposium on The Reliability and Cost Effectiveness of Condition Monitoring Systems, organized by the British Institute of Non-Destructive Testing, to be held in London. Information: The Secretary, The British Institute of Non-Destructive Testing, 1 Spencer Parade, Northampton NN1 5AA. (Tel. (0606) 30124/5).

Security Technology *12th to 14th May*
1982 Carnahan Conference on Security Technology sponsored by the University of Kentucky, IEEE (Lexington Section and AESS) to be held at Carnahan House, University of Kentucky, Lexington, USA. Information: Sue McWain, Conference Coordinator, Office of Continuing Education, College of Engineering, University of Kentucky, 533S. Limestone Street, Lexington, Kentucky 40506. (Tel. (606) 257-3971).

Instrumentation in Flammable Atmospheres *20th May*
A short course on Instrumentation in Flammable Atmospheres, organized by Measurement Technology to be held in Luton Information: Customer Training Department, Measurement Technology Ltd, Power Court, Luton LU1 3JJ. (Tel. (0582) 23633).

Antennas and Propagation *24th to 28th May*
International Symposium on Antennas and Propagation organized by the IEEE in association with URSI, to be held in Albuquerque, New Mexico. Information: IEEE, Conference Coordination, 345 East 47th Street, New York, NY 10017.

Measurements *24th to 28th May*
Ninth Congress on Technological and Methodological Advances in Measurement organized by IMEKO to be held in Berlin. Information: IMEKO, Secretariat, P.O. Box 457, H-1371 Budapest.

Multiple Valued Logic *25th to 27th May*
12th International Symposium on Multiple valued logic sponsored by the IEEE Computer Society, to be held in Paris. Information: Michel Israel, Symposium Chairman, IIE-CNAM, 292 Rue Saint Martin, 75141, Paris Cedex 03, France (Tel. 271 24 14 ext. 511)

Electro *25th to 27th May*
Conference and Exhibition organized by the IEEE, to be held at the Boston Sheraton Hotel and Hynes Auditorium, Boston, Mass. Information: Dale Litherland, Electronic Conventions Inc. 999 N. Sepulveda Blvd., El Segundo, CA 90245 (Tel. (213) 772-2965).

Word Processing *25th to 28th May*
International Word Processing Exhibition and Conference, organized by Business Equipment Trade Association, to be held at the Wembley Conference Centre. Information: Business Equipment Trade Association, 109 Kingsway, London WC2B 6PU. (Tel. 01-405 6233).

Consumer Electronics *30th May to 2nd June*
Consumer Electronics Trade Exhibition sponsored by BREMA together with ICEA and RBA, to be held at Earls Court, London. Information: Montbuild Ltd. 11 Manchester Square, London W1M 5AB (Tel. 01-486 1951).

JUNE

***●Electrostatic Damage** *2nd June*
One day Seminar on 'Electrostatic Damage to Sensitive Devices and its Prevention' organized by the Institute of Quality Assurance in association with the IERE, will be held at The University Arms Hotel, Cambridge. Information: Conference Secretary, IQA, 54 Princes Gate, London SW7 2PG. (Tel. 01-584 9026)

Digital Audio *4th to 6th June*
Conference on The New World of Digital Audio, organized by The Audio Engineering Society, to be held at the Rye Town Hilton, Rye, New York. Information: Audio Engineering Society, 60 East 42nd Street, New York, NY 10165, USA. (Tel. (212) 661-2355/8528).

SCOTELEX '82 *8th to 10th June*
The 13th Annual Scottish Electronics Exhibition and Convention, organized by the Institution of Electronics, to be held at the Royal Highland Exhibition Hall, Ingliston, Edinburgh. Information: Institution of Electronics, 659 Oldham Road, Rochdale, Lancs. OL16 4PE (Tel. (0706) 43661).

● Reliability *14th to 18th June*
The fifth European Conference on Electrotechnics, EUROCON '82, sponsored by EUREL, to be held in Copenhagen. Information: Conference Office, (DIEU), Technical University of Denmark, Bldg. 208, DK-2800, Lyngby, Denmark (Tel. 45 (0) 882300)

Microwaves *15th to 17th June*
International Microwave Symposium organized by the IEEE will be held in Dallas, Texas. Information: IEEE, Conference Coordination, 345 East 47th Street, New York, NY 10017.

Office Automation *15th to 17th June*
Office Automation Show and Conference, to be held at the Barbican Centre, London. Information: Clapp & Poliak Europe Ltd, 232 Acton Lane, London W4 5DL. (Tel. 01-747 3131).

NES '82 *18th June*
*Annual Technical Conference of the Numerical Engineering Society, 'Manufacturing Integrated', organized by the Institution of Production Engineers, will be held at the Hotel Metropole, Brighton. Information: The Manager, Conferences & Exhibitions, Institution of Production Engineers, Rochester House, 66 Little Ealing Lane, London W5 4XX. Tel. 01-579 9411

Fisheries Acoustics *21st to 24th June*
Symposium on Fisheries Acoustics organized by the International Council for the Exploration of the Sea with the collaboration of the United Nations Food and Agriculture Organization, to be held in Bergen, Norway. Information: General Secretary, ICES, 2-4 Palaegade, 1261 Copenhagen K, Denmark

Infodial *22nd to 25th June*
International Congress and Exhibition on Data Bases and Data Banks, organized by SICOB in association with the French Federation of Data Base and Data Bank Producers, to be held in Paris. Information: Daniel Sik, IPI, 134 Holland Park Avenue, London W11 4UE. (Tel. 01-221 0998).

★ Microelectronics *29th June to 1st July*
Conference on The Influence of Microelectronics on Measurements, Instruments and Transducer Design organized by the IERE in association with the IEE, IEEE, IProdE, IOP, IMC, IQA and BES, to be held at the University of Manchester Institute of Science and Technology. Information: Conference Secretariat, IERE, 99 Gower Street, London WC1E 6AZ (Tel. 01-388 3071)

JULY

●Man/Machine Systems *6th to 9th July*
International conference on Man/Machine Systems organized by the IEE with the association of the IERE and other bodies, will be held at the University of Manchester Institute of Science and Technology. Information: Conference Department, Institution of Electrical Engineers, Savoy Place, London WC2R 0BL. (Tel. 01-240 1871)

Materials and Testing *8th and 9th July*
A Symposium on the Inter-Relationship of Materials and Testing, organized by the Institute of Physics, to be held at the University of London. Information: Institute of Physics, 47 Belgrave Square, London SW1X 8QX. (Tel. 01-235 6111).

Simulation *19th to 21st July*
1982 Summer Computer Simulation conference will be held at the Marriott-City Centre, Denver, Colorado. Information: Lawrence Sashkin, 1982 SCSC Program Director, The Aerospace Corporation, P.O. Box 92957, Los Angeles, California 90009.

Control *19th to 21st July*
Conference on Applications of Adaptive and Multivariable Control, sponsored by the IEEE in association with the University of Hull, to be held at the University of Hull. Information: G. E. Taylor, University of Hull, Dept. of Electronic Engineering, Hull (Tel. (0482) 46311 Ext 7113).

● Image Processing *26th to 28th July*
Conference on Electronic Image Processing, organized by the IEE in association with the IEEE and the IERE, to be held at the University of York. Information: IEE Conference Secretariat, Savoy Place, London WC2R 0BL (Tel. 01-240 1871).

AUGUST

★ Software *25th to 27th August*
Residential Symposium on Software for Real-Time Systems organized by the IERE Scottish Section will be held in Edinburgh. Information: Mr J. W. Henderson, YARD Ltd, Charing Cross Tower, Glasgow.

Satellite Communication *23rd to 27th August*
A Summer School on Satellite Communication Antenna Technology organized by the Eindhoven University in association with IEEE Benelux and the University of Illinois will be held at Eindhoven University. Information: Dr E. J. Maanders, Department of Electrical Engineering, University of Technology, Postbox 513, 5600 MB Eindhoven, Netherlands. (Tel. (040) 47 91 11).

SEPTEMBER

Microwaves *6th to 10th September*
Twelfth European Microwave Conference organized by the IEEE in association with URSI to be held in Helsinki. Information: IEEE, Conference Coordination, 345 East 47th Street, New York, NY 10017.

BA '82 *6th to 10th September*
Annual Meeting of the British Association for the Advancement of Science, will be held at the University of Liverpool. Information: British Association, Fortress House, 23 Savile Row, London W1X 1AB.

ICCC '82 *7th to 10th September*
Sixth International Conference on Computer Communication, sponsored by the International Council for Computer Communication, to be held at the Barbican Centre, London. ICCC '82 PO Box 23, Northwood Hills HA6 1TT, Middlesex.

Personal Computer *9th to 11th September*
Personal Computer World Show, to be held at the Cunard Hotel, Hammersmith, London W6. Information: Interbuild Exhibitions Ltd, 11 Manchester Square, London W1M 5AB. (Tel. 01-486 1951).

Remotely Piloted Vehicles *13th to 15th September*
Third Bristol International Conference on Remotely Piloted Vehicles jointly sponsored by The Royal Aeronautical Society and the University of Bristol. To be held at the University of Bristol. Information: Dr R. T. Moses, Organizing Secretary, RPV Conference, Faculty of Engineering, Queen's Building, The University, Bristol BS8 1TR. (Tel. (0272) 24161, ext 846)

***Safety** *14th to 16th September*
A three-day course on Safety of Electrical Instrumentation in Potentially Explosive Atmospheres, organized by SIRA Institute, will be held at UMIST, Manchester. Information: Conferences and Courses Unit, SIRA Institute Ltd, South Hill, Chislehurst, Kent BR7 5EH (Tel. 01-467 2636)

Wescon '82 *14th to 16th September*
Show and Convention to be held at the Anaheim Convention Centre and Anaheim Marriott, Anaheim, California. Information: Robert Myers, Electronic Conventions Inc. 999 North Sepulveda Boulevard, El Segundo CA 90245.

●Broadcasting *18th to 21st September*
The ninth International Broadcasting Convention, IBC '82, organized by the IEE, and EEA with the association of IERE, IEEE, RTS and SMPTE, will be held at the Metropole Conference and Exhibition Centre, Brighton. Information: IEE, 2 Savoy Place, London WC2R 0BL (Tel. 01-240 1871).

Non-Destructive Testing *20th to 22nd September*
National Non-Destructive Testing Conference, organized by the British Institute of Non-Destructive Testing, to be held in York. Information: BInst NDT, 1 Spencer Parade, Northampton NN1 5AA. (Tel. (0604) 30124/5).

★Electromagnetic Compatibility *20th to 22nd September*
Third conference on Electromagnetic Compatibility, organized by the IERE with the association of the IEE, IEEE, IQE and RAeS, to be held at the University of Surrey, Guildford. Information: Conference Secretariat, IERE, 99 Gower Street, London WC1E 6AZ (Tel. 01-388 3071)

***Automatic Testing** *21st to 23rd September*
Exhibition and Conference on ATE and Test Systems will be held at C.I.P., Paris. Information: Network, Printers Mews, Market Hill, Buckingham. Tel. (02802) 5226/5227

Telecommunications and Fibre Optics *21st to 24th September*
Eighth European conference on Telecommunication and Fibre Optics organized by the Electronics Industries Group (GIEL), to be held in Cannes. Information: GIEL 11 rue Hamelin, 75783 Paris Cedex 16

***Automated Assembly** *23rd September*
One day Seminar on 'Automated Assembling: Starting a Project and making it work' organized by the Institution of Production Engineers to be held at the Bowater Conference Centre, London. Information: The Manager, Conferences & Exhibitions, Institution of Production Engineers, Rochester House, 66 Little Ealing Lane, London W5 4XX. Tel. 01-579 9411

Man-Machine Systems *27th to 29th September*
Conference on Analysis, Design and Evaluation of Man-Machine Systems sponsored by IFAC in association with the IFIP/IFORS/IEA, to be held in Baden-Baden, Federal Republic of Germany. Information: VDI/VDE-Gesellschaft, Mess-und Regelungstechnik, Postfach 1139, D-4000 Dusseldorf 1. (Tel. (0211) 6214215)

Instrumentation in Flammable Atmospheres *30th September*
(See item for 20th May)

OCTOBER

Electronic Displays *5th to 7th October*
Electronics Displays Exhibition and Conference, to be held at the Kensington Exhibition Centre. Information: Network, Printers Mews, Market Hill, Buckingham. MK18 1JX. (Tel: (0282) 5226).

Defendory Expo '82 *11th to 15th October*
The 4th Exhibition for Defence Systems and Equipment for Land, Sea & Air, organized by the Institute of Industrial Exhibitions in association with the Defence Industries Directorate of The Hellenic Ministry of National Defence to be held in Athens, Greece. Information: Mrs Duda Carr, Westbourne Marketing Services, Crown House, Morden, Surrey SM4 5EB (Tel. 01-540 1101)

Internepcon *12th to 14th October*
Internepcon Conference and Exhibition, organized by Cahners Exposition Group, to be held at the Metropole Exhibition Hall, Brighton. Information: Cahners Exposition Group, Cavridy House, Ladyhead, Guildford, Surrey, GU1 1BZ. (Tel. (0483) 38083).

***CAMPRO '82** *13th and 14th October*
Conference on Computer Aided Manufacturing and Productivity organized by Institution of Production Engineers and the Society of Manufacturing Engineers USA, will be held at the Mount Royal Hotel, Marble Arch, London. Information: The Manager, Conferences & Exhibitions, Institution of Production Engineers, Rochester House, 66 Little Ealing Lane, London W5 4XX. Tel. 01-579 9411

● RADAR '82 *18th to 20th October*
International Conference on Radar, organized by the IEE in association with the IEEE EUREL, IERE, IMA, RAeS and RIN, to be held at the Royal Borough of Kensington and Chelsea Town Hall, Hornton Street, London W8. Information: IEE Conference Department, Savoy Place, London WC2R 0BL. (Tel. 01-240 1871).
(Papers by 31st May)

● Military Microwaves '82 *19th to 22nd October*
Third International Conference and Exhibition organized by Microwave Exhibitions and Publishers, to be held at The Cunard International Hotel. Information: Military Microwaves '82 Conference, Temple House, 36 High Street, Sevenoaks, Kent TN13 1JG

Multivariable Systems *26th to 28th October*
Symposium on the Application of Multivariable Systems Theory, organized by the Institute of Measurement and Control to be held at the Royal Naval Engineering College, Manadon. Information: The Institute of Measurement and Control, 20 Peel Street, London W8 7PD. (Tel. 01-727 0083).

Instrumentation *26th to 28th October*
Electronic Test & Measuring Instrumentation Exhibition and Conference, to be held at the Wembley Conference Centre. Information: Trident International Exhibitions Ltd, 21 Plymouth Road, Tavistock, Devon PL19 8AU. (Tel. (0822) 4671).

Instrumentation in Flammable Atmospheres *28th October*
(See item for 20th May)

Pattern Recognition *19th to 22nd October*
Sixth International Conference on Pattern Recognition, sponsored by the IEEE in association with the IAPR and DAGM, to be held at the Technical University of Munich. Information: Harry Hayman, P.O. Box 369, Silver Spring, MD 20901 (Tel. (301) 589-3386).

Broadcasting *19th to 21st October*
Conference on Broadcasting Satellite Systems organized by the VDE(NTG) with the association of the specialized groups of the DGLR and the IRT. Information: Herrn Dipl. Ing. Walter Stosser, AEG-Telefunken, Gerberstrasse 33, 7150 Backnang *(Papers by 28th June)*

Manufacturing Technology *26th to 28th October*
Fourth IFAC/IFIP Symposium on Information Control Problems in Manufacturing Technology organized by the National Bureau of Standards, US Department of Commerce, in association with IFAC/IFIP will be held in Gaithersburg, Maryland. Information: Mr J. L. Nevins, Vice Chairman, National Organizing Committee, 4th IFAC/IFIP Symposium Charles Stark Draper Labs, Inc. 555 Technology Square Cambridge. MA 02139 USA. (Tel. (617) 258 1347).

NOVEMBER

Computers *16th to 19th November*
Compec Exhibition, to be held at the Olympia Exhibition Centre, London. Information: IPC Exhibitions Ltd, Surrey House, 1 Throwley Way, Sutton, Surrey SM1 4QQ. (Tel. 01-643 8040).

***Safety** *23rd to 25th November*
A three-day course on Safety of Electrical Instrumentation in Potentially Explosive Atmospheres, organized by SIRA Institute, will be held at Cudham Hall, Sevenoaks, Kent. Information: Conferences and Courses Unit, SIRA Institute Ltd, South Hill, Chislehurst, Kent BR7 5EH (Tel. 01-467 2636)

Instrumentation in Flammable Atmospheres *25th November*
(See item for 20th May)

1983

FEBRUARY

MECOM '83 *7th to 10th February*
Third Middle East Electronic Communications Show and Conference, organized by Arabian Exhibition Management, to be held at the Bahrain Exhibition Centre. Information: Dennis Casson, MECOM '83, 49/50 Calthorpe Road, Edgbaston, Birmingham B15 1TH. (Tel. (021) 454 4416).

MARCH

Component Assembly *March*
Brighton Electronics Exhibition on matching components with insertion, connection and assembly aids and techniques, to be held in Brighton. Information: The Press Officer, Trident International Exhibitions Ltd, 21 Plymouth Road, Tavistock, Devon PL19 8AU. (Tel. (0822) 4671).

●Telecommunications Networks *21st to 25th March*
Second International Network Planning Symposium (Networks '83), organized by the Institution of Electrical Engineers with the association of the IERE, to be held at the University of Sussex, Brighton. Information: IEE Conference Department, Savoy Place, London WC2R 0BL. (Tel. 01-240 1871).

Inspex '83 *21st to 25th March*
Tenth International Measurement and Inspection Technology Exhibition, sponsored by *Measurement and Inspection Technology* in association with IQA and Gauge and Tool Makers' Association, to be held at the National Exhibition Centre, Birmingham. Information: Exhibition Manager, Inspex '83, IPC Exhibitions Ltd, Surrey House, 1 Throwley Way, Sutton, Surrey SM1 4QQ. (Tel. 01-643 8040).

APRIL

Engineering Education *6th to 8th April*
Second World Conference on Continuing Engineering Education, organized by the European Society for Engineering Education, to be held at UNESCO Headquarters in Paris. Information: Mr N. Krebs Ovesen, Danish Engineering Academy, Building 373, DK 2800, Lyngby, Denmark.

SEPTEMBER

Weightech '83 *13th to 15th September*
Third International Industrial and Process Weighing and Force Measurement Exhibition and Conference, organized by Specialist Exhibitions in association with the Institute of Measurement and Control, to be held at the Wembley Conference Centre. Information: Specialist Exhibitions Ltd, Green Dragon House, 64/70 High Street, Croydon, CR9 2UH. (Tel. 01-686 5741) Conference Information: IMC, 20 Peel Street, London W8 7PD. (Tel. 01-727 0083).

OCTOBER

Telecom '83 *26th October to 1st November*
Second World Telecommunication Exhibition, organized by the International Telecommunications Union, to be held at the New Exhibition Conference Centre in Geneva. Information: Telecom '83, ITU, Place des Nations, CH-1211 Genève 20, Switzerland. (Tel. (022) 99 51 11).

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EDITORIAL

IERE NEWS AND COMMENTARY

PAPERS

Speech Analysis and Synthesis

The available methods are discussed in relation to the characteristics of human hearing and three main classes of coding systems are then presented, namely waveform coders, analysis/synthesis systems, usually termed vocoders, and intermediate systems combining the desirable features of the other two classes.

Data Communication Systems

The paper examines the agreement obtained between bench measurements of bit error rate and a simple theory of error probability for two quasi-synchronous data signals. Based on the mean of 64-bit failure rate the performance with a carrier frequency difference of a few hertz is found to be significantly better than with a single transmitter.

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FORTHCOMING ISSUES OF THE JOURNAL

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