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Molecular Electronics

The Next Big Development in Electronics?

MOLECULAR electronics is the name of a branch of physics which will probably be unfamiliar to the majority of today's electronic engineers—but it could well prove to be the basis of a development as significant as the solid-state electronics on which microelectronic integrated circuits depend. Molecular electronics is the ultimate stage of the thin-film and its potential importance is such as to have persuaded the UK Science and Engineering Research Council to promote research into what are now generally known as Langmuir Blodgett films (LB films).

Langmuir and Blodgett carried out their original work on the technique of depositing thin films one molecule thick in the United States at the Schenectady Laboratories of General Electric over 60 years ago. Irving Langmuir's name will be familiar to many radio engineers from his work on thermionics; Katharine Blodgett had the dual distinctions of being the first woman scientist at Schenectady and of being the first woman PhD in physics at Cambridge, where she worked at the Cavendish Laboratory under Rutherford for some years, subsequently returning to Schenectady.

The films are an interesting example of ordered organic systems: they may be assembled one monolayer at a time to form a planar two-dimensional sheet of accurately controlled thickness. The preparation process generally involves depositing a small quantity of a solution of a suitable organic material on to a liquid surface and waiting for the solvent to evaporate; the floating molecules can then be compressed until a quasi-solid, one molecule thick is formed. The container which holds the liquid subphase upon which the monolayer floats is termed a Langmuir trough. To assemble multilayer structures, a suitably prepared substrate is dipped and raised through the surface of the subphase (usually highly purified water); the thickness of the film will depend on the number of monolayers deposited and the molecular size of the material used. In order to investigate the structure of the monolayer and establish the correct dipping conditions, isotherms must first be recorded. The temperature and pH of the subphase are important variables which must be optimized for each molecule studied. The surface pressure where dipping occurs also affects the quality of the film produced. If conditions are carefully controlled and if appropriate molecules are used, one monolayer is transferred on each excursion of a substrate through the water surface. A feature of modern Langmuir troughs is the differential feedback arrangement that maintains the monolayer at the required surface pressure when dipping is in progress.

LB films are now the subject of broad and intense study in both the physical and biological sciences. It is only during the past four or five years that the extensive scientific and technological possibilities have been appreciated. This stage in the development of LB films has required an improvement in the design of the deposition system and the synthesis of novel materials with the required structure to form orientated monolayers on a substrate. Progress has been achieved only by close collaboration between solid-state physicists and innovative synthetic chemists.

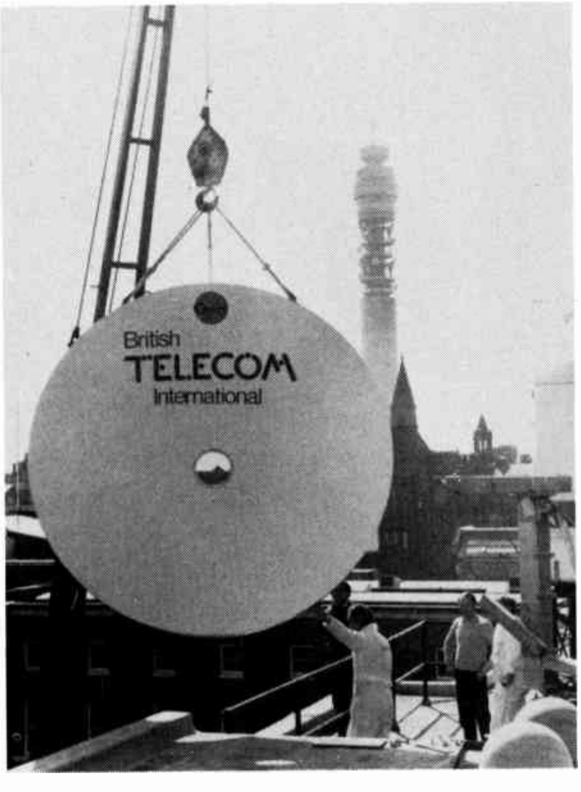
Possible technological applications of LB films are numerous but as an example the good insulating qualities of many materials has led to their incorporation in field effect devices. Thus, whereas the success of integrated circuits based on silicon is due largely to its native oxide, LB films have been successfully deposited as insulating layers on to other semiconductors. An attractive feature of the Langmuir trough technique is its large area capability which could be exploited in display and photovoltaic devices. The presence of an organic layer within a transistor structure also offers exciting possibilities in the transducer field. Other potential applications are in the areas of integrated optics (where the ability to define accurately the thickness and refractive indices is important) and electron beam lithography. The LB technique provides a means of building up organized supermolecular functional units which would be difficult or impossible to achieve by other means. This capability is likely to be capitalized upon in future investigations of high-temperature superconductors, organic semiconductors, conducting polymers, and magnetic storage devices.

With such a potential future for this exciting 'new' technique, how is development to be directed to initiate successful exploitation? The UK Science and Engineering Research Council has set up a Working Party under the chairmanship of Professor Gareth Roberts of the Department of Applied Physical Electronics at the University of Durham which is formulating recommendations for the promotion of research. Symposia for specialists and regional meetings to attract new academics and industrialists into the field are being planned and a large international conference on LB films is to be held at Durham in September 1982. Several British Universities have research projects and may be expected to participate as well as many European, US and other laboratories.

During the fifties the electronics industry entered the new field of solid-state phenomena which lead to more and more sophisticated transistors and eventually to large-scale integrated circuits. LB films, with their close relation to the relatively unfamiliar discipline of organic chemistry, must similarly be taken up by the electronics industry if their potentialities are to be fully realized. The 'vested interests' of solid-state technology must not delay what could be a step forward as significant as the invention of the transistor.

F.W.S.

Project Universe



The IERE Offices in Gower Street are now flanked by both a terrestrial microwave tower and a satellite Earth Station! The picture shows the dish for one of the terminals for Project Universe being lifted on to the roof of University College London recently. In the background is the well-known outline of British Telecom Tower; 99 Gower Street lies just a few houses away to the left.

British Telecom is one of the sponsors of the £3 million project—due to start field experiments later this year—which also involves the government, industry and three universities. The project will use the European Space Agency's test satellite (OTS) to link groups of computer terminals through their Earth stations at six sites in different parts of the UK. Thereafter the method of exchanging information will be monitored as part of the experiment.

The University College Earth station, a three-metre small-dish satellite terminal, is one of three being provided and maintained by Telecom, which will also be supplying and maintaining access to the test satellite, as well as participating in the experiment from its research laboratories at Martlesham Heath near Ipswich in Suffolk.

Altogether six Earth stations will be involved in the link. As well as the one at UCL, stations will be sited later at the universities of Cambridge and Loughborough, the Marconi Research Laboratories at Great Baddow in Essex and the Science and Engineering Research Council's Rutherford Appleton Laboratory at Chilton in Oxfordshire.

The cost of Project Universe is being shared between the Department of Industry, the Science and Engineering Research Council, British Telecom, GEC–Marconi and Logica.

Project Universe is of particular interest to British Telecom as a possible application of SatStream, one of its X-Stream digital services, which is to be offered to UK businessmen in 1984 and affords specialized telecommunications to most of Western Europe. When introduced for business use, SatStream will use two European satellites and enable information to be transmitted at speeds up to 2 Mbit/s.

Notice to all Corporate Members of the Institution of Electronic and Radio Engineers

NOMINATIONS FOR ELECTION TO THE 1982–83 COUNCIL OF THE INSTITUTION

In accordance with Bye-law 49, the Council has nominated the following members for election at the Annual General Meeting to be held on Thursday, 28th October 1982.

President

Under Bye-law 45, the President retires each year but may be re-elected provided he does not serve thereby for more than two years in succession.

For Re-election: H. E. Drew, C.B., Hon.C.G.I.A.

Vice-Presidents

Under Bye-law 46, all Vice-Presidents retire each year but may be re-elected provided they do not serve thereby for more than three years in succession.

For Re-election: Colonel W. Barker; L. A. Bonvini; Major-General H. E. Roper, C.B.; D. L. A. Smith, B.Sc. (Eng.); Group Captain J. M. Walker, R.A.F.

For Election: R. Larry; R. W. Wray, B.Sc.

Honorary Treasurer

For Re-election: S. R. Wilkins

Ordinary Members of Council

Under Bye-law 48, Ordinary Members of Council are elected for three years and may not hold office for more than three years in succession.

FELLOWS

The following must retire: R. Larry

For Election: Professor D. P. Howson, D.Sc., M.Sc.; G. A. Jackson, B.Sc.

MEMBERS

The following must retire: D. J. Kenner, B.Sc., M.Sc.; B. Mann, M.Sc.; K. R. Thrower

For Election: A. F. Dyson, Dip.El.; Professor P. A. Payne, Ph.D.; M. M. Zepler, M.A.

ASSOCIATE MEMBER

The following must retire: P. J. Hulse

For Election: Commander A. R. B. Norris, R.N. (Retd)

ASSOCIATE

The following must retire: M. W. Wright

For Election: D. R. Caunter

The remaining members of Council will continue to serve in accordance with the period of office laid down in Bye-law 48.

Within twenty-eight days after the publication of the names of the persons nominated by the Council for the vacancies about to occur, any ten or more Corporate Members may nominate any one other duly qualified person to fill any of these vacancies by causing to be delivered to the Secretary a nomination in writing signed by them together with the written consent of the person nominated undertaking to accept office if elected, but each nominator shall be debarred from nominating any other person for the same vacancy (Bye-law 50).

By Order of the Council
S. M. DAVIDSON

17th May 1982

Secretary

Members' Appointments

CORPORATE MEMBERS

Dr R. Benjamin, C.B., D.Sc. (Fellow 1975) has had conferred on him the Fellowship of the City & Guilds of London Institute 'in recognition of the eminence he has attained in the engineering profession since graduating from the City & Guilds College'. Dr Benjamin is Chief Scientist and Superintending Director of the Government Communications Headquarters.

Sir Robert Clayton, D.Sc. (Fellow 1977), Technical Director of the General Electric Company, has been appointed to the University Grants Committee for a period of five years.

F. Grimm (Fellow 1973, Member 1952, Graduate 1949), who has been Technical Co-ordinator of the Philips Mobile Radio Management Group based in Cambridge since 1977, has recently retired; he has completed over 30 years with the Group and its predecessor, Pye Telecommunications, of which he was Technical Director. Mr Grimm is to take up a part-time consultancy with the Science and Engineering Research Council, co-ordinating a programme on terrestrial radio communications systems.

Professor D. R. Towill, D.Sc. (Fellow 1970) has been appointed Head of the Department of Mechanical Engineering and Engineering Production at the University of Wales Institute of Science and Technology. He has been at UWIST since 1970 and founded the Dynamic Analysis Group. Professor Towill currently serves on the Professional Activities Committee and on Group Committees. He has contributed numerous papers to the Journal, several of which have received premiums.

J. R. Ashman (Member 1978) has been appointed Technical Director of McMurdo Instrument Company, Portsmouth where he will be responsible for all research and development associated with electronic components, defence and marine safety equipments. Mr Ashman was previously with AB Electronics where he held the post of Technical Director, Controls Division, having previously been Managing Director of the Company's Switching Division.

Brigadier J. F. Blake (Member 1969) has taken up an appointment as Deputy Assistant Chief of Staff, Communications and Electronics Division, SHAPE, in Belgium. He previously held the post of Director of Military Communications Procurement in the Ministry of Defence. He is a member of the Aerospace, Maritime and Military Systems Group Committee.

N. G. Brown (Member 1962) has joined Management in Action, a consortium of independent consultants specializing in management development, as a principal. He was formerly Director of an export marketing consultancy.

I. P. Bunyard (Member 1980, Graduate 1972) who was with the East Midlands Gas for nine years, latterly as Telephony Engineer, has taken up an appointment as a Telecommunications Engineer with Shell UK.

Lt Col W. J. Crouch, B.Sc. (Eng.), REME (Member 1971), Commanding Officer, REME Technical Services BAOR since 1979, has been appointed Commanding Officer, 33 Central Workshop REME, Newark.

D. J. Doughty (Member 1965, Graduate 1954) who recently retired from the Electrical Quality Assurance Directorate in the Ministry of Defence (PE), has taken up a part-time appointment as Principal Project Officer with Lonsdale Technical. While at EQD he was responsible for the BS9000 and CECC systems in the Directorate's role as National Supervising Inspectorate. His new post is mainly concerned in providing management support services to MOD's electronics components authorities. Mr Doughty contributed a wide-ranging survey paper to the Journal some years ago covering waveguide component manufacturing and inspection methods.

A. F. Flynn (Member 1971, Graduate 1968) has been appointed Technical Director of Salter Industrial Measurement, West Bromwich. He was previously Technical Services Director of Penlon, Abingdon.

D. Haley, B.Sc. (Member 1966) has been appointed Senior Lecturer and Head of the Telecommunications Section at West Kent College of Further Education, Tonbridge. He joined the College as Lecturer II in 1975 following some thirty years with Kelvin Hughes.

Gp Capt. A. G. Hicks, M.A., RAF (Member 1972, Graduate 1970) has recently taken up an appointment as the Senior Air Staff Officer at Headquarters, Northern Maritime Air Region RAF, Pitreavie Castle, Dunfermline. Prior to the past two years when he was at the RAF Staff College, Gp Capt. Hicks was Officer Commanding 42 Squadron, RAF St Mawgan.

H. Jackson, M.Sc. (Member 1973, Graduate 1970) has been appointed Head of the Electrical Engineering Department, Longlands College of Further Education, Middlesbrough. Mr Jackson previously held the post of Senior Lecturer in Computing at Darlington College of Technology.

H. C. Jamieson, O.B.E. (Member 1967) has been appointed Managing Director of System Production, based in Farnborough. Mr Jamieson was previously Northern Director of Afa-Minerva, Leeds.

G. J. Le Jehan (Member 1969, Graduate 1966) is now Programme Progress Officer with Plessey Defence Systems, Christchurch.

A. S. Omosule (Member 1972, Graduate 1971), Director of Engineering Services, Ondo State Television Corporation, Nigeria since 1979, has been appointed General Manager of the Corporation.

H. J. H. Perry, B.Sc., M.B.A., P.Eng. (Member 1960), for the last four years Quality Control Manager at the Manati plant of Davis & Geck, Puerto Rico, has returned to company headquarters in Danbury, Connecticut, as Quality Assurance Manager, New Medical Devices.

Lt Cdr J. B. Sadler, RN (Member 1973, Graduate 1974) has been posted to the Admiralty Underwater Weapons Establishment, Portland, as Naval Application Officer for Submarine Weapon Handling and Discharge Systems. He was Staff Weapons Engineer Officer with the Second Submarine Squadron from October 1979 to February 1982.

G. E. Turner (Member 1969, Graduate 1966) has been appointed Director, Technical Services USA, with Mitel in Washington D.C. Previously Eastern Regional Manager for Plessey Communications in New York since 1979, Mr Turner was with that company for some 25 years for much of the time as Installation Manager in Brazil.

Lt Col P. B. Webster, B.Sc. (Eng.), R Signals (Member 1977) has been appointed Commanding Officer, 10th Signal Regiment, Hounslow, Middlesex.

Sqdn Ldr R. M. Webster, RAF (Member 1973, Graduate 1969) who was Officer Commanding Electrical Engineering Squadron at RAF Leuchars, has been posted to the Wright-Patterson Air Force Base in Ohio on exchange duties with the US Air Force.

T. C. Wright (Member 1969, Graduate 1963) is now Product Manager with GEC Telecommunications in Chiswick, New South Wales. Before going to Australia in 1981 he was a Systems Engineer with Marconi Space and Defence Systems, Camberley.

NON-CORPORATE MEMBERS

Cdr A. R. B. Norris, RN (Ret) (Associate Member 1959) recently retired from the Royal Navy after completing a total of 35 years non-commissioned and commissioned service, his final posting being at the Admiralty Surface Weapons Establishment. He is now a Project Manager with Vickers Design and Projects, based at Eastleigh, Hants. Cdr Norris, who has served for a number of years on the Aerospace, Maritime and Military Systems Group Committee, has been nominated as a member of the council, representing the class of Associate Members.

E. Jones (Associate 1975) who was a Project Officer with MOD(PE), is now a Senior Telecommunications Officer with the Ministry of Defence (Air) in Hanover.

Obituary

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

Major Bertram Norman Greville Cooper, REME (Ret) (Member 1962, Associate 1958) of Wokingham, died on 7th December 1981, aged 60. After completing his apprenticeship he worked in the aircraft industry as an experimental draughtsman on the design of jigs and tools until 1947 when he was commissioned in the Royal Electrical and Mechanical Engineers. During the next 24 years he held appointments at REME Workshops at home and overseas and also was on the staff of the Training School at Arborfield.

Peter Ernest Noel Fuller (Member 1973, Graduate 1970) of West Kirby, died on 20th February 1982 aged 53 years. After service in the Royal Navy as Radio Electrician Petty Officer, Mr Fuller worked for some three years in commerce before going to the Ministry of Civil Aviation in 1957 as a Radio Technician. In 1966 he completed the HNC with endorsements following studies at Paisley College of Technology and he was a Telecommunications Officer II at the Lowther Hill Radar Station in Lanarkshire from 1968 to 1970. Before moving to the headquarters of the National Air Traffic Services in London he was an Assistant Signal Officer concerned with the specification and siting of new surveillance and a.s.m.i. radar at London Heathrow Airport. Since 1978 Mr Fuller had been at the NATS headquarters in Liverpool as a Senior Air Traffic Engineer.

Bernard Francis Gray (Fellow 1959) died on 25th March last, following a heart attack. His death was sadly premature: he would have been but 58 years old on May 15th. His career was very much linked with Hatfield Polytechnic to which, while it was still Hatfield College of Technology, he was appointed as Head of Department of Electrical Engineering in 1956. Previously, as a graduate in engineering from Imperial College, he had worked for Marconi Instruments, for George Kent of Luton and, between 1949 and 1956, had taught his subject at Luton College of Technology, latterly as a Senior Lecturer.

It was the expressed intention of its Principal that Hatfield should become 'the pinnacle of technical education in Hertfordshire'. In the undoubted achievement of that objective Bernard played an emphatic part during his 25 years with the College. He was made Dean of Engineering in 1975; following incipient heart trouble and the need to respect it, he became Associate Dean in 1979 until his early retirement in 1981.

Bernard's extra-mural work for the academic world, the engineering profession, and the Institution in particular brought him national acknowledgment. He served as a member of the Institution's Education and Training Committee from 1960 until his death, including a five year period as its Chairman. He was a member of Council between 1968 and 1970, and represented the Institution on many important bodies, including the CNAA Electrical Engineering Board; the Joint Committees for the ONC in Engineering and for HNC/D in Electrical and Electronic Engineering; on Programme Committee A2 of T.E.C.; and on bodies concerned with the foundation of the CEI examinations. At the time of his death he was Chairman of the Beds and Herts Section of the Institution.

One piece of work in which he rightly took great pride was his initiation of the Ordinary National Diploma in Technology. To lead the development of this course he was seconded for a year from the Polytechnic. The product was a splendid grouped course, broadly based, and explicitly intended as an alternative to the A-level route into professional engineering courses: a model of what a course of its kind should be. The format reminds one that Bernard was a proponent of the grouped course. Like many of us, he did not go along with 'the Green Shield stamp' approach to engineering education. Indeed one of his several publications is entitled 'Who wants modular courses?' (*The Radio and Electronic Engineer*, January 1976.) Echo answers, Who?

Many another of Bernard's professional activities—his visits overseas, his textbooks on Electronics, his work for the Institution of Electrical Engineers, his written and spoken contributions to many aspects of engineering education—are perforce omitted from this tribute. An actively creative life is not to be compressed into a few hundred words. The essential fact is that Bernard had given much and had yet more to offer; a keen analytical mind associated with the essential quizzical good humour which was Bernard's will be sorely missed by his very many friends and colleagues in the profession. His widow and his three children should take comfort in that, and it is to them that all of us in the Institution offer our sympathy.

D.L.A.S.

Leslie Hale (Member 1961, Associate 1960), of Stevenage died on 4th March 1982, aged 54 years.

Leslie Hale served in the RAF as Radar Fitter up to 1947 when he joined Standard Telephones and Cables, North Woolwich. Initially an Inspector, he subsequently became a Circuit Investigation Engineer before moving in 1957 to English Electric Aviation, Stevenage (now part of British Aerospace) as a Senior Design Engineer. He worked first in the Precision Products Group and later in the Automatic Test Systems Group, eventually becoming Engineering Manager. In 1975 Mr Hale moved to the Personnel and Training side of British Aerospace Guided Weapons as Manager for Engineering, the position he held up to his death.

Since 1976 Leslie Hale had served on the Institution's Electronics Production Technology Group Committee and became its Chairman in 1978.

William Charles Henshaw (Associate Member 1973, Associate 1951) of Chipping Sodbury, died on 25th March 1982, aged 65 years. During the War 'Bill' Henshaw served in the Royal Air Force as a Signals Officer being Mentioned in Despatches in January 1945 while serving in South East Asia. In 1949 he was appointed a Lecturer and subsequently a Principal Lecturer in the School of Management Studies at Bristol College of Commerce where he had special responsibility for statistical studies which included operational research and scope and applications of electronic computers. He obtained the MSc. degree in mathematical statistics during this period. After the formation of Bristol Polytechnic, Mr Henshaw was appointed Director of Studies in the Department of Computer Studies and Mathematics. Both before and after his retirement he was an industrial consultant in statistics and operational research. Bill Henshaw was a founder member and from 1958 to 1962 Honorary Secretary of the Institution's South Western Section. His apparently unbounded energy and enthusiasm in anything he undertook were in a large measure responsible for the sound foundations of this Section.

Benjamin James Lambert (Graduate 1963) of Malvern, died in February 1982 aged 52. Mr Lambert joined the Technical Group, REME at the Royal Radar Establishment, Malvern, in 1960 following service in the Royal Signals. He was granted exemption from the Institution's Graduateship examination as a result of a Higher National Certificate with endorsements gained at the College of Electronics Malvern, and in 1976 he obtained the B.A. degree from the Open University. At the time of his death Mr Lambert was a Professional and Technology Officer Grade 2 with the Telecommunications and Radar Branch REME at Malvern.

Joseph Polonsky (Fellow 1960), of Issy-les-Moulineaux, near Paris, died on 10th January 1982 at the age of 68. Joseph Polonsky was well known to television engineers in Great Britain and indeed throughout the world for his numerous contributions to the development of the transmission side of television including SECAM and he also made important contributions in the field of biological electronics. He was for many years Director and Technical Manager of the Television Department of Thomson-CSF and he came on several occasions to IERE Conferences, presenting a paper on a mobile microwave outside broadcasting link in 1959. He also took an active part in the organization of the technical programme of the Montreux Convention. M. Polonsky was a graduate of the Ecole Polytechnique de Grenoble and of the Ecole Supérieure d'Electricité (Radio Section). In 1958 he was appointed a Professor at ESE de Grenoble; he also was a group leader of biophysics research at CNRS.

Conference Report

Fibre-Optics

IERE Conference held in London on 1st–2nd March 1982

An audience of some 170 assembled at the City Conference Centre (Institute of Marine Engineers) to hear 20 papers presented.* The conference was opened with a keynote address by Professor Alec Gambling of Southampton University, who, as usual, set the scene magnificently for the subsequent papers.

The conference was all-embracing, covering as it did sessions on components and devices, civil communications systems, defence systems, standards and testing. The papers were generally all of a very high standard and an international flavour was obtained with 25% of the papers coming from abroad, notably USA, France, Germany and Italy.

Papers in the first session on components and devices concentrated on the development of light-emitting diodes and particularly noteworthy was the paper by A. C. Carter *et al.* on devices operating at 1300 nm.

Expanded beam optical connectors were covered by J. C. Challans *et al.* and the final paper in the session discussed the use of fibre-optics in gyroscopes.

The second session of the first day commenced with a valuable overview of the principles and applications of optical fibre sensors by J. P. Dakin. Further papers considered the use of lasers and their military applications, followed by a paper on transmitter and receiver packages also developed for military applications. The final paper of the first day, excellently presented by M. Ramsey, discussed rugged fibre optic cables for military use.

The second day commenced with three papers on civil communications. The paper by Hensel and Hooper described a cost-effective short-haul link for System X and the following paper by Stern and Garrett gave the latest developments in monomode fibre technology. The third paper discussed the design and application of one particular long haul system.

Applications of fibre optics in defence systems was the subject of the next session, which included papers on fibre-optic systems for *Ptarmigan*, *Gina*, and for the cruise missile. A fourth paper discussed the effects of nuclear radiation.

The final session of the conference covered standards and testing with some most informative papers giving the latest situation in the IEC and CCITT.

There is no doubt that the conference was highly successful and fulfilled a very real need. Could it be that this conference will become an annual event?

C. J. LILLY

*The papers are contained in IERE Conference Proceedings No. 53, price £27 from the Publications Department.

Electronics and Safety

While the emphasis of the annual report of HM Factory Inspectorate of the Health and Safety Executive on Manufacturing and Service Industries* is largely concerned with the industries using chemical processes or involving mechanical hazards, several passages in the report relate to the application of electronic equipment.

R.F. Hazards

Following extensive investigations led by the Factory Inspectorate of the possibilities of radio-frequency ignition hazards at major gas processing plants in Scotland, the HSE co-ordinated a unique series of tests carried out by RLSD, ERA Technology, Shell Research and the BBC at a large radio transmitting station to simulate conditions in which incendive sparks could be caused in large structures by radio frequency transmissions. The results from these successful tests are being used to produce a practical revision of the British Standard Guide (BS 4992) on this subject. (The project formed the basis of a special issue of *The Radio and Electronic Engineer* on Radio and Radar Ignition Hazards, published in April 1981.)

Microprocessor-based Control Systems

New uses for microelectronically-based control systems are continually being found in factories and the work of the Microprocessor Technical Committee, described in the 1979 report, has continued. Its initial work aimed at producing broad guidance on safety implications and resulted in the preparation of a booklet which is now published. The booklet, which is aimed at the general reader rather than the specialist, is not intended to be a complete technical treatment, but attempts to point out some of the changes to traditional thinking on safety that may become necessary when considering microelectronic devices. A draft code of practice is also being developed. The committee is now turning its attention to specific uses of these devices.

Over the last few years the Chemical National Industry Group has examined a number of computer/microprocessor controlled process plants, mainly in the continuous process field in major hazard plants. The purpose, primarily, has been to determine the degree of control retained by the human operator, the need for safety systems independent of the computer control which enable the plant to be moved into the safe mode should an emergency arise, the relative importance of safety back-up systems built into the computer and the importance of operator training.

This work has shown that while some computer/microprocessor control system designers might claim that conventional back-up safety devices are rendered somewhat superfluous by the back-up system built into the controller, responsible manufacturers have, wisely, continued to design process plant for computer control with conventional back-up safety features, often retaining the facility for direct plant control at times when the computer is not functioning.

High component costs are leading plant design away from comprehensive mainframe computers to discrete microprocessor devices which control certain parts of the reaction and which work in conjunction with one another. In some cases the replacement with microprocessor-controlled equipment has resulted in the duplication of analogue and digital read-out and the retention of manual control. However, it is likely that the costs of future plant will mean the convention for control systems will consist of computer/microprocessor equipment with built-in safeguards, a keyboard for initiating the start of the process and obtaining read-out for process efficiency and a visual display unit. Already such units are in use, some with the facility for producing mimic plant diagrams showing the stage of reaction in terms of temperature, pressures, etc., at various points, in addition to the now familiar read-out prints.

Although the need for computer/microprocessor plant control is greatest in those areas where continuous operation within close limits is essential, high labour costs and material costs will accelerate the introduction of this type of equipment

*'Health and Safety: Manufacturing and Service Industries 1980', HM Stationery Office or from booksellers, price £5 plus postage.

into smaller manufacturing operations. High energy reactions and those with high inventories of flammable solvents at elevated temperatures will require conventional back-up safety systems, but there are many mild reactions which use relatively innocuous materials where back-up safety devices may be of a rudimentary nature, and control may be left to the computer.

The need for operator training, particularly in the major hazard plants using computer control, is important. In general computer/microprocessor control is more efficient than manual control, and operators have little to do beyond watching visual read-out devices and listening for alarms. However, sensing devices and transducers have limitations and often the human nose is the most sensitive detector for small tell-tale leakages, which would not be noticed by automatic devices. Some manufacturers have instituted what can only be termed as 'walk abouts' for operators, who would never normally need to venture from the control room. This system consists of placing analogue or digital read-out devices at various locations throughout the plant and requiring data logging at set times. The 'walk abouts' are developed from in-plant training schemes to familiarize operators with the plant hardware and reaction sequence, their purpose being to enable operators to conduct orderly shutdown procedure during an emergency or to operate the plant if possible using manual methods should the computer fail.

Robotics

In 1980 there was an upsurge in the interest shown by industry in the use of industrial robots and the Technical Committee has necessarily to consider the foreseeable safety problems.

The question of whether a particular robot installation is safe is dependent on many factors including the size and power of the robot, the nature of the task, the degree of interaction with personnel and the methods used for programming and maintenance. For this reason, the safety of any installation is likely to depend far more on an individual assessment of the factors involved than on adherence to a uniform standard. Some broad principles can, however, be established and HSE presented a paper on the subject at the International Conference on Robotics in 1981. The paper described a method of analysis which, it is hoped, will highlight the issues which need to be considered in relation to any particular installation.

Following a recent initiative from the Department of Industry there have been discussions between representatives of France, Germany and the UK on robotic developments. Safety was identified as a possible area for co-operation and these discussions are to continue with the Inspectorate's participation.

The discussions have led to the consideration of the possibility that a UK liaison body be formed which would be representative of the principal robot manufacturers, users and researchers so that informed advice may be available to HSE. The Inspectorate is also involved with the Machine Tool Research Association in its proposal to produce codes of practice for the safe use of robots.

The use of robots in industry may produce benefits for companies in terms of safety as well as production. Robots are being used in fettling, paint spraying, stacking applications, loading and unloading and many other processes where some degree of risk may be inevitable. It would be regrettable, therefore, if the potential benefits were offset by the failure to recognise the potential hazards from the robot itself.

Many robots are taught the required work cycle by being moved through the action pattern under the control of a programme using a portable pendant control (not unlike the controls for an overhead crane). The machine positions are recorded in a memory and, once taught, the robot can reproduce the action pattern as required. The teach method often involves close approach to a machine under power and there is evidence that this is the area where accidents are most likely to occur. Causes may vary from inadvertent operation of control buttons and malfunctions of equipment to lack of awareness of the exact movement of the machine. Unlike conventional machines robots typically have a working volume which may be quite large and their movements are not as predictable as devices constrained by slideways. This aspect of robot use is being considered with parties interested in robotic development.

Machine Tools

Computer numerically controlled (CNC) machine tools are becoming widely used in the engineering industry and a change in the pattern of accident causation on such machines is beginning to show itself. In the nature of things it would be expected that with programmed systems the risk of an operator coming into contact with dangerous parts of the machinery would be diminished and that the introduction of such systems would be entirely beneficial from the safety point of view. It is becoming evident, however, that the potential hazards may have been altered rather than removed. There is evidence to suggest that the typical incident with CNC machines is likely to be the ejection of components or tools consequent upon some aberration in the control system — possibly because of faulty programming, the effects of stray electrical fields or other similar external causes.

The Microprocessors Technical Committee is proposing to look at these problems early in its programme of developing activities.

Standard Frequency and Time Service

(Communication from the National Physical Laboratory

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

MARCH 1982	MSF 60 kHz	GBR 16 kHz	Droitwich 200 kHz
1	−10·4	31·9	67·1
2	−10·3	34·0	67·0
3	−10·4	33·3	66·9
4	−10·6	33·2	66·8
5	−10·7	33·5	66·6
6	−10·7	33·7	66·4
7	−10·8	33·4	66·1
8	−10·5	33·4	65·8
9	−10·5	33·7	65·6
10	−10·5	34·5	65·4
11	−10·3	34·0	65·4
12	−10·3	34·5	65·4
13	−10·3	34·5	65·3
14	−10·3	34·7	65·0
15	−10·3	34·5	64·8
16	−10·3	35·5	64·6
17	−10·4	35·3	64·5
18	−10·4	35·2	64·3
19	−10·3	35·2	64·1
20	−10·4	35·6	64·0
21	−10·3	35·4	63·7
22	−10·6	35·0	63·5
23	−10·6	35·1	63·3
24	−10·6	34·5	63·1
25	−10·6	34·4	62·9
26	−10·7	34·3	62·7
27	−10·6	34·3	62·4
28	−10·6	34·2	62·1
29	−10·4	34·5	61·8
30	−10·3	35·4	61·5
31	−10·4	35·0	61·2

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 1·2 part in 10^{11} per day.
(d) It may be assumed that the relative stations on 200 kHz at Westerglen and Burghead will follow the day to day changes in these phase values.

Contributors to this issue*

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Martin Ackroyd graduated from Birmingham University in 1966, with first class honours in electronics and electrical engineering. He is now Reader in Telecommunications at Aston University, where he is primarily concerned with research on the analysis and control of traffic in communication systems. Before joining the University in 1975, he was in charge of a section in the Central Research Laboratories of EMI concerned with image processing and pattern recognition. His first academic appointment was as a lecturer in the Department of Electronic and Electrical Engineering at Loughborough University of Technology, where he had done research on signal theory for his Ph.D. degree.

*See also pages 282 and 296

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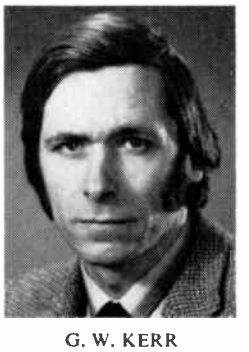

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A survey of methods for digitally encoding speech signals

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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

Modern communications systems are increasingly using digital transmission, at various digit rates, for the many advantages thereby obtained. There are many methods that can be used for converting analogue speech signals to digital form. The aim is to produce very good speech quality using simple equipment at a low digit rate, but any application inevitably involves choosing the appropriate compromise between these three conflicting attributes. The paper reviews the most important of currently known methods, including systems already in widespread use and those still in the research stage. The problem of assessing speech transmission performance is also discussed.

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1 Introduction

There are many advantages that can be gained by representing speech signals in digital form. In transmission the use of digital regenerative repeaters can make the degradation in long-distance transmission negligible. Digital signals are easy to handle in modern electronic switching systems. The use of digits for speech enables telegraphy, speech and data to be integrated into one transmission network. In military communication the digital representation greatly eases the problem of providing communication security. In speech storage and retrieval applications the ability to use large random-access digital stores in computers gives great flexibility.

To achieve these numerous benefits in practice for the various applications there are widely varying requirements. Ideally all users would like to be offered very good speech quality and very low digit rate, with reliable compact low-cost equipment. However, there is at present no way of satisfying the users on all of these points, and there must inevitably be a compromise between the three conflicting attributes of speech quality, digit rate and equipment complexity. The relative importance of these attributes depends on the application.

In civil telephony the most important aspect of digital coding is that it should not appreciably degrade the speech quality as perceived by the users of the system, who comprise a very wide range of talkers. This requirement is made more stringent by the fact that in existing networks there are always some links where, with old equipment and poor lines at local ends, the speech transmission is already only marginally acceptable, and if the addition of digital transmission to part of the network were to make the speech quality any worse such links would be unusable. On the other hand the fact that at present civil telephony is limited to the band 300–3400 Hz removes the requirement for coding speech signal components outside this band. Because of the large numbers of equipments that could be involved, cost and size must be major factors in the choice of coders for civil telephony, and will become even more important if digital coding is in future provided on a per-channel basis at the users' telephones. For this application transmission digit rate, although important, is the least vital of the three attributes, and has to be chosen high enough for the system to meet the other needs.

For some military radio communication requirements, such as during difficult h.f. radio propagation or in the presence of intentional jamming, it may be impossible to achieve a transmission rate of more than 1000–2000 bit/s. If speech communication is essential, both speech quality and simplicity of equipment may then have to be sacrificed, but there will be some choice in the compromise between these two factors. However, in a military situation some difficulties are reduced by the fact that there is more control over the users, who can be trained in communications procedures and, if necessary, selected for suitable voices. It is also possible to choose microphones and other equipment to be the most suitable for the specific conditions of use.

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In systems requiring a large range of digitally-coded speech messages for subsequent retrieval, such as in announcing machine applications, both low digit rate and high quality are likely to be important. However, provided the decoder is not too complicated, it is not serious if the complexity at the coding end is very high, as a very small number of coders will be required. It might also be acceptable if the coding process could only work more slowly than real time, using high-digit-rate temporary storage of the digitized speech waveform for the input. In the final coded form it is no disadvantage for this application if the number of digits stored per unit speech time is highly variable, so giving scope for further digital storage savings during short silences or highly predictable parts of the signal.

The aim of this paper is to review the most important of currently-known methods of speech coding for any of the above uses, pointing out how the three-way compromise between quality, bit rate and complexity can be made. The extent to which the systems exploit the properties of human speech production and perception to achieve digit-rate economy will also be explained. There is a vast range of published literature on speech coding, and it is therefore not practicable to present more than a small proportion of typical references to the various methods. Several of the cited papers do, of course, give further references to the subject, although many are the original papers abour particular methods.

2 Possibilities for Reducing Digit Rate

For any ordinary audio communications channel it is possible to calculate its communication capacity in terms of the bandwidth of the channel and the degree of accuracy to which the signal can be specified within that bandwidth, using principles formulated by Shannon.[1] The accuracy is usually limited by random noise in the channel. For a typical telephone channel the bandwidth is about 3 kHz and the signal/noise ratio might be 40 dB. The capacity in this case is about 40 000 bit/s. For a high-fidelity monophonic sound reproducing system the bandwidth would be about five times greater, and the noise would probably be 60–70 dB below peak signal level. In this case a rate of about 300,000 bit/s is required to specify any of the possible distinct signals that could be transmitted by such a system. For digitization based on quantizing to a set of equally spaced levels the number of bits per sample needs to be slightly more than the number implied by the signal/noise ratio if there is to be negligible additional degradation to the signal, so the total digit rates needed might be 60 000 and 400 000 bit/s in the two examples quoted.

In contrast to these very high figures, it is known that human cognitive processes cannot take account of an information rate in excess of a few tens of bits per second, (e.g. see Ref. 2), thus implying a ratio of information transmitted to information used of between 1000 and 10 000. This very large ratio suggests that the full information capacity of an audio channel is not necessary for speech transmission. Unfortunately for the communications engineer the human listener can be very selective in deciding what aspects of the signal are chosen for attention by the few tens of bits per second available for cognitive processing. Usually the listener concentrates on the message, which, with its normal high degree of linguistic redundancy, falls well within the capacity available. However, the listener may pay attention specifically to the voice quality of the speaker, the background noise, or even to the way certain speech sounds are reproduced. In these latter cases the actual word content of the message might be largely unnoticed. The more usual situation would be for the message to take up a fair proportion of the total attention, but for the remaining capacity available to be spread round the other factors. Any sounds that do not fit in with the subconsciously or consciously expected quality of the signal will attract the listener's attention, and thus occasional clicks, non-linear distortion, or an unusual way of producing a particular speech sound might all be noticed by a listener.

There are, however, two properties of speech communication that can be heavily exploited in speech coding. The first is the restricted capacity of the human auditory system. The inner ear can be regarded as an approximation to a very large number of fairly broad bandpass filters, with extensively overlapping pass bands. On the output of each filter is a highly non-linear 'hair cell' connected to a nerve fibre carrying the response, coded in the form of frequency and timing of nerve pulses, to the higher auditory centres for further processing. The effective selectivity of the ear's analysis varies with frequency, and it is established that the frequency-domain structure of complex sounds cannot be resolved in a bandwidth less than the so-called 'critical bands',[3] which are of approximately constant width up to about 1 kHz, and have a width of about one-sixth of their centre frequency above 1 kHz. Pitch perception of periodic sounds is, of course, much more precise than would be expected from the filtering action of the inner ear, and is probably connected with the fact that nerve pulse intervals tend to synchronize to multiples of the repetition period of the sound. The ear does not seem to be appreciably affected by phase distortion in its input if the pressure wave pattern applied to the individual hair cells is not much disturbed. To a first approximation this means that group-delay distortion is not noticed if it varies little within one critical bandwidth. However, human speech is normally heard in the presence of significant time-dispersion caused by reverberation in the acoustic environment, and so a much larger random group delay distortion is not detected as unnatural if it is within the range that occurs naturally (i.e. a small number of tens of milliseconds). Another consequence of the characteristics of the ear's filtering action is the ability of high-intensity sounds in one frequency region to mask lower-level sounds in other regions. This masking is very great within the width of a critical band, but even outside a critical bandwidth significant masking occurs, mainly of high-frequency sounds by low-frequency ones. This property is of great value in making the listener insensitive to low-level background noise or distortion when the wanted signal is of higher level.

When designing speech coding systems it can also be advantageous to make use of the fact that the signal is known to be produced by a human talker. The physiology and habits of use of the speaking mechanism put strong constraints on the types of signal that can occur, and this fact can be exploited by making the receiving end of a speech link model the human speech functions to some extent. The potential reduction in digit rate that can ultimately be achieved from this approach is much greater than is possible from exploiting auditory restrictions alone, but such systems are only suited to auditory signals that are speech-like.

In the following discussion of individual coding methods some mention will be made of the extent to which properties of perception and production are exploited. The systems presented are divided into three classes, referred to as waveform coders, analysis/synthesis systems and intermediate; members of each class exploit aspects of production constraints and of perception tolerance.

In all the methods of speech coding described there are many design parameters that affect the performance to a greater or lesser extent, and other features, perhaps not fully evaluated, that have been suggested as likely improvements. A comprehensive discussion of all of these points would be useful, but in a paper of this length there is no possibility to do more than outline the general principles of each system, and comment briefly on speech transmission performance, digit rate and complexity.

3 Waveform Coders

Waveform coders, as their name implies, attempt to copy the actual shape of the waveform produced by the microphone and its associated analogue circuits. If the bandwidth is limited, the sampling theorem (explained by E. T. Whittaker in a 1915 Edinburgh University Report, and included in many modern textbooks, such as Ref. 4) shows that it is theoretically possible to reconstruct the waveform exactly from a specification in terms of the amplitudes of regularly-spaced ordinate samples taken at a frequency of at least twice the bandwidth. In its conceptually simplest form a waveform coder consists of a band-limiting filter, a sampler and a device for coding the samples. The sampler operates at a rate higher than twice the cut-off frequency of the filter. The amplitudes of the samples are then represented as a digital code (normally binary) with enough digits to specify the signal ordinates sufficiently accurately. There is obviously no point in making the specification much more accurate than can be made use of for the given input signal-to-noise ratio. This principle of coding, known as pulse code modulation (p.c.m.), was suggested by Reeves in 1938[5] and is now widely used for feeding analogue signals into computers or other digital equipment for subsequent processing (in which case it is known as analogue-to-digital (a/d) conversion). The process is not normally used in its simplest form for speech transmission, because the required digit rate for acceptable quality is too high. Simple p.c.m. does not exploit any of the special properties of speech production or auditory perception except their limited bandwidth.

The distortion caused by p.c.m. can be considered as the addition of a signal representing the successive sample errors in the coding process. If the number of bits per sample in the code is fairly large (say >5) this 'quantizing noise' has properties not obviously related to the structure of the speech, and its effect is then subjectively equivalent to adding a small amount of flat-spectrum noise to the signal. If the number of digits in the binary code is small or if the input signal level exceeds the permitted coder range, the quantizing noise will have different properties, and will be highly correlated with the speech signal. In this case the fidelity of reproduction of the speech waveform will obviously be much worse, but the degradation will no longer sound like the addition of random noise. It will be more similar subjectively to the result of non-linear distortion of the analogue signal. Such distortion produces many intermodulation products from the main spectral components of the speech signal, but even when extremely distorted the signal usually contains sufficient of the spectral features of the original signal for much of the intelligibility to be retained.

The sound pressure waveform of a speech signal has most of its total power in the frequency range below 300 Hz, even though the information content of the signal is almost entirely carried by the spectrum above 300 Hz. As quantizing noise has a flat spectrum its effect on the signal-to-noise ratio is much more serious for the weaker but more important higher frequency components. A considerable performance improvement for p.c.m. can be obtained by taking into account this property of speech production, and applying 'pre-emphasis' to the speech signal with a simple linear filter to make the average spectrum more nearly flat. After p.c.m. decoding the received signal can then be restored to its original spectral shape by 'de-emphasis', so reducing the higher frequency components of the quantizing noise. For normal communication purposes it is not, however, necessary that the de-emphasis should match the pre-emphasis, as speech intelligibility is actually improved by attenuating the low-frequency components.

The amplitude of the quantizing noise of simple p.c.m. is determined by the step size associated with a unit increment of the binary code. During low-level speech or silence this noise might be very noticeable, whereas during loud speech it would to some extent be masked by the wanted signal. For a given subjective degradation in p.c.m. it is therefore permissible to allow the quantizing noise to vary to some extent with signal level, so exploiting a property of perception. This variation can be achieved either by using a non-uniform distribution of quantizing levels or by making the quantizing step size change as the speech level varies. Both methods have commonly been adopted, and have enabled excellent quantizing-noise performance to be achieved at 8 bits/sample, and useful communications performance at 4 bits/sample. Civil telephony uses p.c.m. with 8 bits/sample at 8 kHz sampling rate, so requiring 64 kbit/s.[6] There is an instantaneous 'companding' characteristic that gives an exponential distribution of

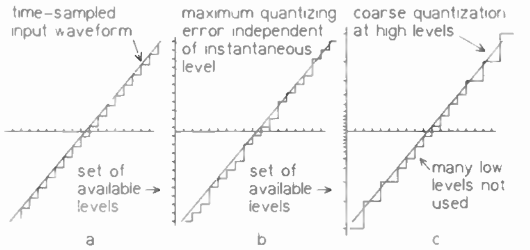

Fig. 1. P.c.m. quantizing for a large signal excursion through zero. (a) Sampled input signal. (b) Uniform quantizing adapted to the speech level. (c) Non-uniform quantizing.

quantizing intervals except at the lowest levels. The sampling rate is generous for the 300–3400 Hz bandwidth required, but this high sampling rate simplifies the requirements for the band-limiting filters.

The time resolution properties of the auditory system ensure that masking of quantizing noise by the higher level wanted signals is effective for at least a few milliseconds at a time, but instantaneous companding will give finer quantization near zero every time a large amplitude signal changes sign. It is obvious that more effective use will be made of the transmitted digits if the step size is not merely a function of waveform ordinate height, but is changed in sympathy with the short-term average speech level. In this case, however, some means must be devised to transmit the extra information about the quantizing step size. This information can be sent either as a small proportion of extra digits interleaved in the digital waveform description, or more usually it is embodied in the waveform code itself. The latter process is achieved by using a feedback loop in the coder that modifies the quantal step size slowly up or down according to whether the transmitted codes are near the extremities or centre of their permitted range. As the same codes are available at the receiver, it is in principle easy to keep the receiver quantizing interval in step with that at the transmitter, but digital errors in the transmission path disturb this process and will affect the general signal level besides adding noise to the received audio waveform. The effects of these two sorts of companding on the coding of a large waveform excursion through zero are shown in Fig. 1.

Delta-modulation[7] is a widely used alternative type of waveform coding. A delta-modulator generates a local copy of the input waveform, and successively modifies this copy, as specified by a digital code, to try to make it follow the input. The basic scheme is illustrated in Fig. 2. In its original and simplest form the quantizer uses only one bit per sample, and merely indicates whether the

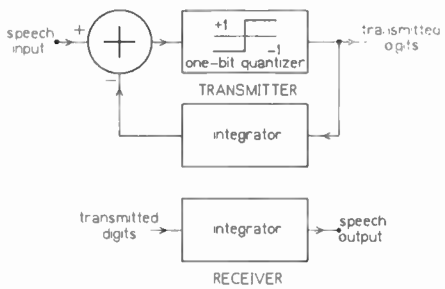

Fig. 2. Block diagram of a delta-modulator.

copy is to be increased or decreased by one quantum. Such a coder offers the possibility of extremely simple hardware implementation, and if run at a high enough sampling rate can approximate waveforms very closely. The process of waveform following in small steps makes delta-modulation work best on signals in which differences between successive ordinates are small. Thus the low-frequency dominance in speech signals is accommodated directly by delta-modulation without pre-emphasis, and it is acceptable to use a quantal step that is only a very small fraction of the waveform amplitude range. In contrast, a flat-spectrum input would cause frequent slope-overloading if used with the same step size and sampling rate. Typical waveforms in simple delta-modulation are illustrated by Fig. 3.

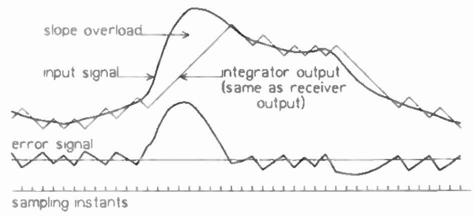

Fig. 3. Waveforms in a simple delta-modulator.

The use of a single bit per sample in delta-modulation is basically inefficient because a sampling rate much in excess of twice the highest frequency in the input signal is needed for close following of the input waveform. However, the intrinsic feedback loop in the waveform coding process gives the coder some 'memory' of coding distortion on previous waveform ordinates, for which it partially compensates later. This advantage of delta-modulation can be combined with those of p.c.m. if a p.c.m. coder is used instead of a one-bit quantizer in the feedback loop. Current terminology describes this arrangement as differential p.c.m. (d.p.c.m.).[8]

The advantages of and techniques for level adaptation apply to delta-modulation in the same way as to p.c.m., and adaptive forms of coder are normally used, so exploiting the noise masking properties of auditory perception and the slow level changes of speech production. Adaptive d.p.c.m. seems to be the most efficient of the straightforward waveform coding processes, and can give excellent quality speech of telephone bandwidth at 32 kbit/s. At lower rates, such as 16 kbit/s, the quantizing noise is noticeable, but slightly less objectionable than the noise given by adaptive delta-modulation or adaptive p.c.m. at the same digit rate.

Sometimes digit rate constraints have forced users to try these simple waveform coders at rates far below the 16–32 kbit/s needed for reasonable speech quality. In these circumstances conventional considerations of quantizing noise no longer apply, and the underloading and overloading that is an inevitable consequence of the small ratio of digit rate to bandwidth makes the coding error highly correlated with the speech signal. It is then not realistic to use signal-to-noise ratio as a performance criterion, but it is necessary instead to consider the perceptual effects of non-linear distortion. It has been demonstrated that speech communication is possible with simple waveform coders at rates below 10 kbit/s,[9] but the distortion is so severe that their applications are very limited. Modern technology has reduced the importance of the simplicity of implementation of these coders, and so for these lower rates it is possible to achieve better results with slightly more complex equipment that might still be small enough and low enough in cost to be acceptable. Some possible methods are mentioned in Section 5. For new applications it does not seem sensible to use the simple types of waveform coder below 16 kbit/s.

4 Analysis/Synthesis Systems (Vocoders)

Vocoders all make use of one basic property of speech production that gives a great saving in digit rate. Human speech sounds are generated from one of two types of sound source. The 'voiced' sounds are the result of the pulsating air flow between the vibrating vocal folds, and 'unvoiced' sounds result from turbulent air flow through a constriction of the vocal tract. Both of these sound

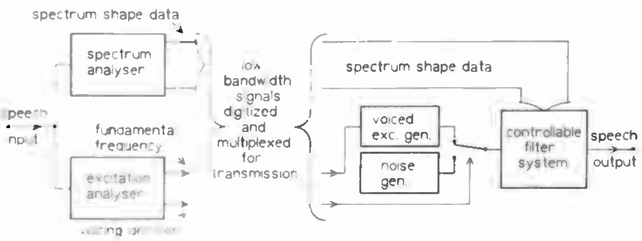

Fig. 4. Block diagram of the basic elements of a vocoder.

sources have significant power over a wide range of audio frequencies, but the spectrum of the radiated sound results from spectral shaping of these sources by the acoustic resonant system of the vocal tract. The resultant sound, analysed over a time window of a few tens of milliseconds, has either a periodic or random time structure. In the frequency domain the fine structure of the spectrum consists either of lines, resulting from harmonics of the fundamental frequency, or is continuous, during random noise excitation. (Some speech sounds use a mixture of random and periodic sound sources, but ignoring this fact has only a minor effect on the performance of vocoders.) The envelope of the speech spectrum is the result of combining the spectral trend of the corresponding sound source with the response of the vocal tract. By separating the fine structure specification of the sound sources from the overall spectral envelope description, and specifying both in terms of a fairly small number of slowly varying parameters, it is possible to transmit a reasonably adequate description of the speech at data rates of 1000–3000 bit/s. At the receiving end of a link the speech is re-synthesized by using either a periodic or random noise source feeding a dynamically controlled spectral shaping filter. The possibility of coding speech signals in this way was first described by Dudley[10] over 40 years ago. The basic elements of a vocoder are shown in Fig. 4.

There are many differences between vocoders with regard to technology of implementation, and in the principles of the device for measuring the fundamental frequency of voiced speech (usually known as the 'pitch extractor'). However, the important differences in principle between vocoders are in the means of specifying the variable spectral shaping filter. Although many types have been demonstrated in research studies, there are currently three spectral shaping filter types of much greater importance than any others. These are known as channel vocoders,[11] linear predictive coding (LPC) vocoders[12] and formant vocoders.[13] In the channel vocoder the spectrum is represented by the response of a bank of contiguous variable-gain bandpass filters. The way the desired overall response can be approximated using separate contributions from the individual channels is shown in Fig. 5. In LPC vocoders the spectral approximation is given by the response of a sampled-data filter, whose all-pole transfer function is chosen according to a least-squared-error criterion. The configuration of such a filter is illustrated in Fig. 6, which also indicates how the same type of filter system is used for adaptive predictive coding (see Sect. 5). In formant vocoders the spectrum is specified in terms of the frequencies, and usually also the amplitudes, associated with the resonant modes ('formants') of the speaker's vocal tract. The relationship between the formant control signals and the synthesized spectral shape can be seen from Fig. 7. With all these methods of description the data are coded into 'frames' associated with speech spectra measured at intervals of 10–30 ms.

By suitable choice of channel frequencies to match the distribution of critical bands the channel vocoder can take advantage of the varying frequency resolution of the ear, but unless a very large number of channels can be used (with consequent high digit rate) it is difficult to achieve a good approximation to the spectrum shapes around the formant peaks. In LPC vocoders, on the other hand, the all-pole properties of the synthesis filter make very good approximations to formant shapes possible. However, the correct analysis to achieve this result will only be obtained when the overall speech spectrum really is like the response of an all-pole filter. During vowel sounds this approximation is often very close, although there are frequent other occasions when this is not so, particularly during nasalized vowels and many consonant sounds. In these latter conditions the

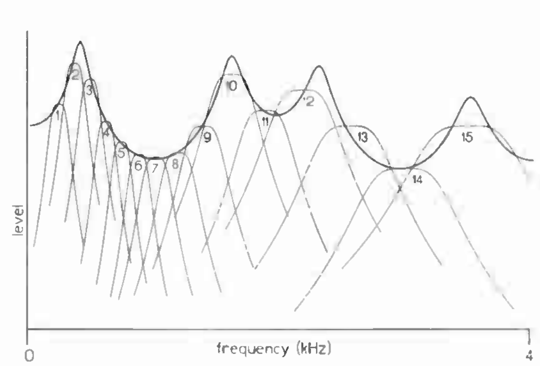

Fig. 5. Contributions of individual channels to a channel vocoder output spectrum. Thick line—desired spectrum shape; Thin lines—contributions from the channels.

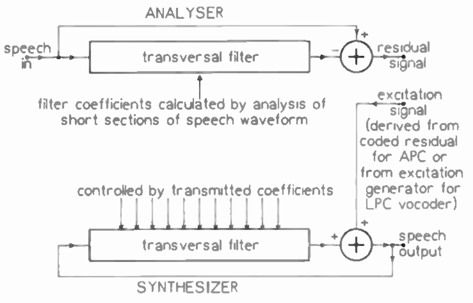

Fig. 6. Use of a predictor filter in LPC and APC systems.

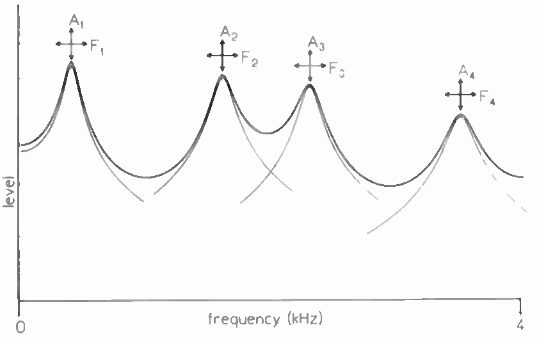

Fig. 7. Illustration of how formant control signals affect the synthesized spectrum. Thick line—desired spectrum shape. Thin lines—contributions from the individual formants.

LPC synthesis frequency produces spectral peaks whose bandwidths are too large, with a consequent 'buzziness' in the speech quality. Another inherent property of normal LPC vocoders is that all regions of the spectrum are treated equally with regard to accuracy of frequency specifications, and so no advantage is taken of the variable frequency resolution of auditory perception. Overall the advantages and disadvantages of LPC and channel vocoders are nearly equally balanced; both give usable but rather poor speech transmission performance at 2400 bit/s using fairly complex equipment.

Formant vocoders are very different from the other two types. They use a synthesizer that is much more closely related to human speech production, because not only do they use the choice of periodic or noise sources as in other vocoders, but also the spectral filter system has resonators that are explicitly related to the principal formants of the input speech. Thus the coding system can be constrained to deal only with the known frequency range and necessary accuracy of specification of each formant. The true bandwidths of the formants do not vary much during speech and such variation as does occur is fairly predictable; provided they are within the limits of natural variation, preservation of the actual formant bandwidths is not perceptually important. In consequence this property of the resonances is not normally transmitted in formant vocoders. (This contrasts with LPC vocoders where the bandwidths of the poles of the synthesis filter are intrinsic properties of the way the filter is specified by the transmitted parameters.) There is a further reduction in information rate that is more practicable with formant vocoders than in the other types, based on the time-varying pattern of the synthesizer specification. Because the responses of successive spectral frames are explicity related to the specifications of the formants, which normally move smoothly, it is possible to transmit frames specifying the formant parameters at irregular intervals and to interpolate changes between them.[14] By this technique the mean frame rate can be reduced to 25 per second or less, while still retaining the ability to have a frame rate as high as is needed during rapidly changing sounds. Such a coding system can use a total data rate as low as 1200 bit/s while retaining excellent speech quality, provided the initial formant analysis is satisfactory.

Analysis is the main difficulty with formant vocoders. When the spectral envelope of a speech sound shows a small number of well-defined peaks it is comparatively trivial to assign these to formants in a sensible way. However, there are occasions, particularly in consonants, near vowel/consonant boundaries or even in the middle of vowels if the fundamental frequency is very high, when it is not clear what is the most appropriate way to assign the parameters of the synthesizer to measurable spectral peaks of the input signal. One of the most powerful analysis methods, which is computationally extremely expensive, is to use analysis-by-synthesis.[15] In this system the analyser contains a copy of the synthesizer, and it makes its output as close as possible to the input speech, using a suitable similarity criterion. The control parameters of this synthesizer are transmitted to the synthesizer in the receiver.

There are further reductions of digit rate that are in principle possible with all three types of vocoder. One method, first suggested under the name 'pattern matching vocoder'[16] and more recently as 'vector quantization',[17] is to select from only a small proportion of the full range of possibilities for the combinations of spectral shape parameters. In all the types of vocoder described above the multi-dimensional space specified by the spectral shape parameters is very unevenly occupied, and large parts of this space are never used at all. By choosing, according to some suitable criterion, the closest to the current input of a limited number of stored spectral patterns, it is possible to specify the spectral shape resonably well in a vocoder with some 10–20 bits, instead of the more typical 40 spectrum bits per frame. Systems of this type have been demonstrated in the laboratory using digit rates around 600–800 bit/s, for channel and LPC vocoders. Performance has, however, been noticeably worse than is obtained from normal coding methods at 2400 bit/s. Although not yet demonstrated, there seems no reason in principle why a good formant vocoder should not be coded in this way; combined with variable frame rate coding a pattern-matching formant vocoder should give appreciably lower digit rate at better quality than either LPC or channel types, because of its closer relationship to speech production.

Much more dramatic savings in digit rate are potentially possible with so-called 'phonetic vocoders'.[18] In this type of vocoder the incoming speech has to be coded in phonetic units; the sequence of phonetic symbols is accompanied by timing and fundamental frequency information, so that the received speech can have the same intonation and timing as the original. However the spectral envelope of the synthesized speech is generated from a small inventory of pattern sequences, or by rules for generating such sequences from the phonetically coded data. Such a vocoder requires a transmission capacity of only 100–200 bit/s, but the best demonstrations so far still have serious difficulties with both analysis and synthesis. There is a wide range of possible methods for synthesis of speech from phonetic symbols, combined with some indication of pitch and timing. The best of these (e.g. Ref. 19) have given highly intelligible although somewhat unnatural synthetic speech, but of course do not provide the different characteristic qualities of a range of individual talkers. However, the more serious difficulty with phonetic vocoders is to obtain a sufficiently reliable automatic recognition of the phonetic content for the wide variety of speech that many different talkers might produce. The solution to this problem would seem to need a process that 'tunes in' to an individual's speech characteristics, and still seems many years away.

For sufficiently restricted applications it is possible to use automatic speech recognition and synthesis in a somewhat vocoder-like system for extremely low digit rate. This type of speech transmission system requires the range of possible messages to be very limited; in fact the messages have to be made up from only a fairly small vocabulary of isolated words or phrases. Current automatic speech recognition methods can achieve quite a high accuracy of identification of such words when the machine is set up for a specific talker. The following 'synthesis' could be technically trivial, being merely replaying corresponding recorded waveforms of natural speech for the same words.

5 Intermediate Systems

There are many ways of combining some of the detailed signal description possibilities of waveform coders with some of the signal redundancy exploitation of vocoders. The resultant intermediate systems normally give better speech reproduction in the 4–16 kbit/s range than is possible with either of the other two classes of system at these digit rates. Their complexity is, of course, always greater than for simple waveform coders, and some of the higher performance systems may be more complicated than vocoders.

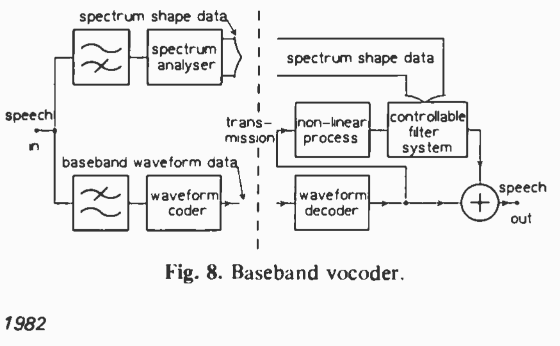

Fig. 8. Baseband vocoder.

One useful approach to gaining some of the advantages of waveform coding and vocoders is to use vocoder techniques for the perceptually less critical high-frequency regions, and to use waveform coding for the low frequencies, up to, say 800 Hz (see Fig. 8). As this 'baseband' is less than one-quarter of the total band required, only a small fraction of the normal waveform coding digit rate is needed to transmit it fairly well. The high-frequency band will need slighly less bits than the corresponding complete vocoder system. The combined digit rate of 'baseband vocoders' is usually in the 8–10 kbit/s range, which gives much improved performance over complete vocoders and over simple waveform coders at the same digit rate. An important feature of baseband vocoders is that the vocal fold source periodicity is inherently contained in the baseband signal. This fact means that the periodicity does not need to be measured, so avoiding a common practical difficulty with conventional vocoder implementations. A suitable excitation signal for the vocoder part of the system can be generated by some sort of non-linear operation on the decoded low frequency speech waveform, and therefore these vocoders are often referred to as 'voice-excited' vocoders.[20] Baseband coding with voice excitation has been mostly applied to channel vocoders, perhaps because this type is more directly amenable to use over a defined sub-portion of the total speech frequency band.

The technique of linear prediction analysis for vocoders inherently makes available in the analyser a signal, known as the residual, which if applied as input to the LPC vocoder synthesis filter will regenerate the speech waveform exactly. If some digitally coded representation of the residual can be transmitted to the receiver it can be used as excitation instead of conventional vocoder excitation (Fig. 6). There are many techniques for coding the residual, at various digit rates. Provided the voice periodicity is retained and the reconstructed residual at the receiver has an approximately flat spectral envelope even very simple methods will offer some performance advantage over normal LPC vocoders. At the other extreme, research on advanced coding methods has produced computationally expensive demonstrations giving substantially perfect reproduction of telephone bandwidth speech at moderate digit rate.[21] Residual coding methods provide a useful excitation signal for the LPC synthesis filter using only 2–8 kbit/s. The total coding system, including the LPC filter specification, is then typically in the range 4–12 kbit/s. For reasons that would require more explanation than is justified here most of these coding systems are known as adaptive predictive coders (APC), although other variants are called residual-excited linear predictors (RELP).[22]

Another way of making use of the ear's tolerance to less accurate signal specification at higher frequencies can be provided using waveform coding, by treating the speech signal in many separate bands. These systems are known as sub-band coders.[23] Each filtered sub-band is coded individually by a waveform coding process (Fig. 9), using a sampling rate equal to twice the

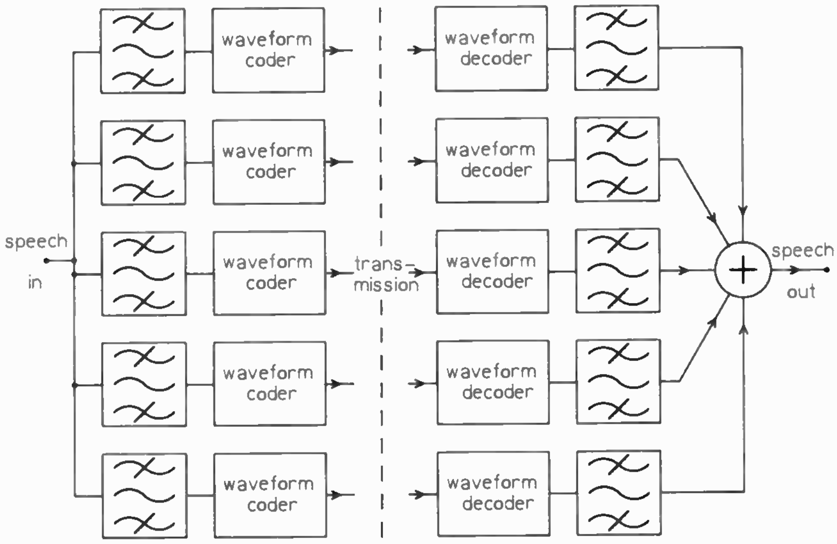

Fig. 9. Sub-band coder.

bandwidth in each case. If the bands are made as narrow as the ear's critical bands the quantizing noise in each band can be largely masked by the speech signal in the same band. In addition, the fact that most power in the higher bands is from the 'unvoiced' randomly excited sounds means that the waveform shape in these bands need not be specified so accurately, so needing fewer bits per sample. In practice, complexity of filter designs makes a choice of about five bands more attractive than the 15–20 needed for critical bands, and even with this number the speech quality in the 16–32 kbit/s range is much improved over that possible with the best conventional waveform coders at the same bit rates.

A different approach that utilizes both the ear's tolerance and some aspects of speech production is adaptive transform coding (ATC).[24] The most usual implementation of this method splits the sampled speech waveform into successive blocks of 128–256 samples, and takes a discrete cosine transform of each block. Allocation of bits to code each transform can be decided adaptively; by careful choice of the bit allocation it is possible to get a subjectively very close approximation to telephone speech with 16 k/bits, and useful communication quality with only 9·6 kbit/s.

Harmonic compression is a term used to describe a method of achieving a reduction of speech bandwidth by a small integer ratio, and this process may be applied before digital coding. The method depends on the fact that voiced speech is harmonic in the frequency domain or periodic in the time domain. A frequency-domain version was described by Schroeder, Logan and Prestigiacomo in 1962,[25] and involves analysing speech through a multichannel bank of narrow band-pass filters. Each filter output then contains at most one harmonic; the outputs are divided in frequency by a small integer ratio, filtered by filters appropriate for their divided frequency ranges, and recombined. A time-domain method of achieving a similar result was suggested by Gabor in 1947,[26] but the technology for successful implementation has only become available comparatively recently. Present-day versions segment the speech signal into successive lengths equal to one fundamental period, then omit a proportion of the periods, join up the remainder and slow down the signal output to maintain the original speech time scale. The inverse process in either domain is used for re-expansion. Harmonic compression and re-expansion by 2 : 1 produces little obvious subjective degradation of a speech signal, except a slight time-blurring of rapid transients. A factor of 4 : 1, however, causes more noticeable loss of time structure, and would not be acceptable for most applications. One great advantage of 2 : 1 harmonic compression is that it is ideally suited for combination with waveform based coding methods, such as sub-band coding or adaptive transform coding, to produce half the digit rate without a disastrous reduction of speech quality.

A much simpler method of speech coding for intermediate digit rates that seems to have potential for some improvement in performance over the basic waveform coding methods is known as time-encoded speech.[27] In this system the telephone band waveform is divided into intervals between successive zero crossings. These waveform segments are each described by one of a small inventory of waveform patterns, using perhaps five bits per pattern, and further information is sent to represent amplitudes. The possibilities for this type of system have not yet been fully investigated, so it is not yet clear what place it occupies in the compromise between digit rate, quality and complexity. However, the operations seem to require considerably less computation than most other systems aiming for good quality in the 4–16 kbit/s range.

An essential feature of time-encoded speech is that it generates symbols at a variable rate and requires a buffer delay to enable a uniform transmission rate to be achieved. This feature might also be used to advantage in other waveform coding schemes, so that obvious waveform redundancy can be exploited where it occurs, but arbitrary waveshapes can still be transmitted. Systems of this type have not been widely investigated, but they might offer a significant digit rate reduction

Table 1. Summary of the properties of speech coding systems.

Class	Examples	Bit rate Kbit/s	Quality	Complexity	Current Status
Waveform	p.c.m. deltamodulation (with companding or level adaptation)	64	excellent	low	widespread use
		32	very good		
		16	fairly good		demonstration available
		8	poor		
Analysis/synthesis	channel vocoder	2·4	fair	high	widespread use
	LPC vocoder	2·4	fair	high	some use, increasing
	formant vocoder	1·2	can be good	very high	research
	pattern-matching vocoder	0·8	poor so far	very high	research
	phonetic vocoder	0·1–0·2	poor so far	very high	research
Intermediate	baseband vocoder	9·6	good	fairly high	demonstration available
	APC or RELP	16	very good	fairly high or high	designs available—research continues
		8	good		
		4·8	fairly good		
	sub-band coding	16	good	medium	research
	SBC with harmonic compression	8	good	high	research
	adaptive transform coding	16	very good	high	research
		8	good		
		4·8	fairly good		
	ATC with harmonic compression	8	very good	high	research
		4·8	good		
	time-encoded speech	4–16	not yet proved	fairly low	research

over the simple waveform coders, without causing unacceptable amounts of transmission delay.

6 Conclusion

The characteristics of the types of speech coder described in this paper are summarized in Table 1. The entries for speech quality and complexity are meant to be only a rough guide, partly because the descriptive words used are rather vague, but mainly because there are so many possible variations of design parameters that the same principles can give a wide range of equipment capabilities. In particular, the very high quality APC demonstrations described in Ref. 21 are not properly represented. Some significant differences in performance (such as between the different types of simple waveform coder) are not great enough to be shown in the Table.

It can be seen that there is a bewildering variety of speech coding methods available, each with its own particular advantages and disadvantages, and it is very difficult for a system designer to make the best compromise between the conflicting factors which should influence the choice. Characteristics such as cost, size, weight and digit rate can be assessed using mainly engineering and economic criteria. Although such assessments are not simple, the more difficult problem is to assess how faults in the reproduction of the transmitted speech affect the communicators.

The ultimate criterion for evaluation of any speech communication system is how it satisfies its users, and ease of conversation will normally be the main factor affecting user's opinions. Although it is possible to simulate free conversation for laboratory testing,[28] such tests need to be very extensive to obtain statistically reliable results. For this reason it is normal to assess speech coders using various word intelligibility tests,[29,30] which do not need such large experiments to obtain stable results. The drawback with these methods is that they are so unrepresentative of how a telephone is really used that they are not suitable for defining absolute performance standards. They are useful, however, for comparisons between systems of similar type, as a guide to optimizing design parameters. For systems of very different type there may be factors influencing the acceptability to the users that do not cause the word intelligibilities to differ.

There is an even greater assessment difficulty that applies to most of the systems described in Table 1 as being still in the research stage. These days most speech coding research is conducted using a computer to simulate the coders, normally not in real time. It is thus not convenient to test the coders on conversation, or even on sufficient speech material for word intelligibility tests. In these cases it is usual for the research groups involved to have a small number of carefully spoken read sentences stored in the computer, and to use these to demonstrate all new coding systems. The careful speaking reduces the need for the coder to respond well to rapidly changing sounds, and the linguistic redundancy (which is almost 100% if the sentences have been heard previously!) makes intelligibility judgments very unreliable. Demonstrations of this type are very misleading, even for comparison between similar systems. However, in the early stages of research on a speech coder read sentence demonstrations are usually all that are available, and it is very difficult, even for experienced people, to resist the temptation to make judgments on such inadequate evidence.

For many years telecommunications engineers have been looking for means of rating speech links by purely objective measurement methods, and some progress has been made for application to civil telephone networks.[31] These methods have recently been extended for more widespread application by the introduction of the speech transmission index method of Steeneken and Houtgast.[32] Their method uses test signals and fidelity criteria that seem to embody most of the important features of speech communication. Results using this method appear to correlate well with subjective tests results for a large range of transmission impairments, and there is good reason to believe that the speech transmission index will be of value for evaluating many digital coding systems.

Irrespective of the methods of assessing systems, there seems little doubt that in due course digital coding will become almost universal in telephony. Its use removes so many problems associated with gradual impairment of analogue signals as they undergo successive stages of transmission. However, the digit rate needed and the users' satisfaction with the speech quality depend on having the right coding system for the particular application. As the applications are so varied it seems likely that many of the systems reviewed here (and others not yet invented) will be of practical value for several years to come, although the differences between systems are so great that compatibility requirements, where they exist, may seriously restrict the choice in some cases.

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Indexing Terms: Radio systems, Data communication systems

Multi-transmitter data systems—performance with stationary receivers

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SUMMARY

A theory is developed, based on bench measurements with two quasi-synchronous data signals, which predicts the mean bit error rate (b.e.r.) and the mean 64-bit word failure rate experienced at stationary receivers in the overlap area of two quasi-synchronous transmitters. The signals from the transmitters suffer independent Rayleigh fading. The performance with a carrier frequency difference of a few hertz is shown to be 2 dB better, and with a larger difference (18 and 180 Hz) to be 5–10 dB worse, than with a single transmitter.

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List of Symbols

p_e	probability of bit error
$s(t)$	time waveform of the envelope of the received radio carrier
$p_e(t)$	time varying bit error probability
Q and R	empirical constants defining modem performance with receiver front-end noise
S_a or S_b	envelope amplitude of the carrier received from transmitter A or B
$P(S_a)$	probability density function of S_a
$\langle p_e \rangle$	mean value of p_e
p_w	probability of a word being in error
$\langle p_w \rangle_{\text{fading}}$	mean word error rate with fading

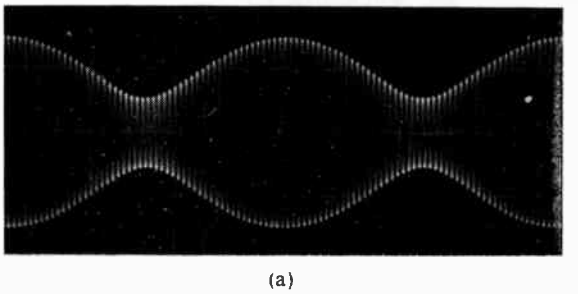
$\omega_d = 2\pi f_d$ = angular carrier frequency difference

1 Introduction

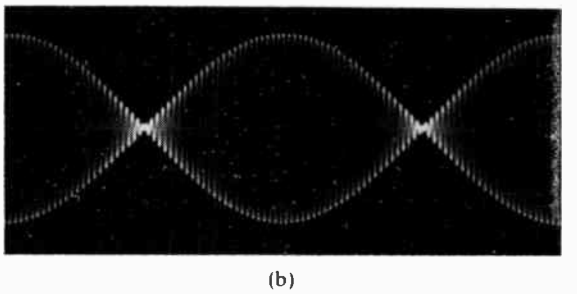
In a multi-transmitter data system, users must be provided with an acceptable performance, even in the overlap areas where signals of similar amplitude are heard from two transmitters. The mobile performance with two transmitters has been reported[1] as similar to the single transmitter case. This paper reports the performance to be expected at stationary receivers.

A stationary user will receive a steady signal from each base station site; however, due to the small difference between the frequencies of the transmitted carriers, the resultant received signal will vary with time due to carrier interference, as shown in Fig. 1. If the two signal envelopes are nearly equal in amplitude, as in Fig. 1(b), the envelope of the resultant will fall to the front-end noise level of the receiver at regular intervals and bursts of errors will occur.

The overlap area in a quasi-synchronous system is the region where the mean levels of the signals from the two transmitters are approximately equal. However, as a



(a)



(b)

Fig. 1. Interfering quasi-synchronous carriers. (a) nearly equal and (b) equal amplitudes.

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result of multipath propagation and wave interference the level of the signal envelope from each transmitter varies from point to point by 30 dB or more, relative to its mean level, and even though the means are equal, generally the signal envelopes are not. This is illustrated in Fig. 2, which shows the envelopes of two fading signals with equal mean levels as a function of distance along the street. A field measurement to find the average performance in the overlap area would involve stopping at a large number of points, so that enough combinations of signal envelopes from the two transmitters were experienced for the average performance to emerge. This was considered to be too time consuming and laborious.

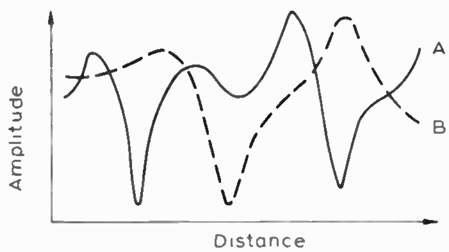

Fig. 2. Typical fading signals with equal mean levels.

The approach used here is a combination of bench measurements of bit error rates to find the probability of bit error as a function of the signal envelopes from the two quasi-synchronous sources, and numerical integration in which the signals are each given the full range of values that would result from measurements at points throughout the overlap area. Although the computer is used to simulate the range of signal envelopes caused by fading, the results are considered reliable since the performance of the receiver, which is the difficult part of a simulation, is found by actual measurements on practical receivers.

2 Bench-measured Bit Error Rate with Quasi-synchronous Signals

The first stage is to measure on the bench the bit error rate (b.e.r.) as a function of signal level, with various relative levels of the second quasi-synchronous signal present, as in Fig. 3. The signals were generated at 165 MHz by synthesized signal generators modulated by a common subcarrier data signal using the emerging standard of 1200 b/s f.f.s.k. with 1200 and 1800 Hz signalling frequencies. The measurements were made

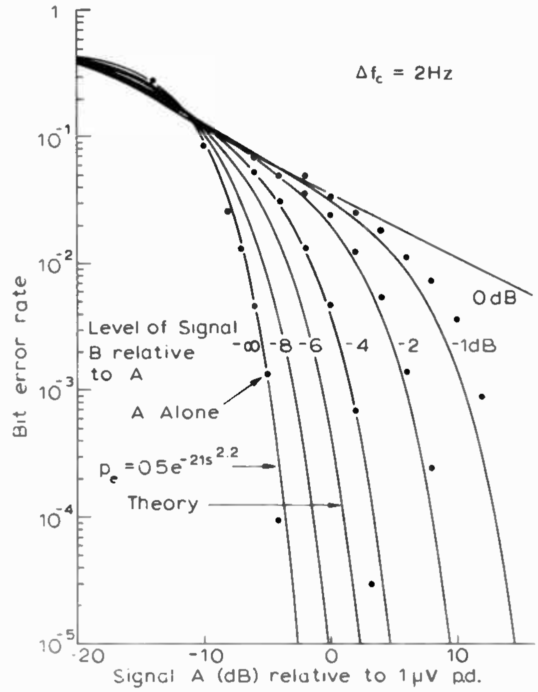

Fig. 4. Stationary a.m. quasi-synchronous error performance. (*Note.* 12 dB SINAD at −6·5 dB.)

with both amplitude and frequency modulation of the r.f. carrier, using either 50% modulation depth or 2 kHz deviation. A common data signal was used to avoid difficulties with the ambiguity in the f.f.s.k. data format, as explained in Ref. 1. The signal generator outputs were added and passed to a conventional 12·5 kHz channelled mobile receiver (a Pye Olympic M201 or M202). The demodulated data signal was then compared with a reference and the number of errors counted.

The a.m. results are shown in Fig. 4 as a plot of b.e.r. against the level of signal A, with the relative level of signal B as parameter. The measured points are plotted together with curves predicted by a theoretical model (described below), which can be seen to fit the measured points closely. With only one signal (i.e. signal B = −∞ dB) the usual curve of b.e.r. against signal level is seen in Fig. 4, with a sudden fall in b.e.r. as the signal level increases. As signal B is introduced the b.e.r. is increased, with the result that an increase of 16 dB in signal A is needed to maintain a b.e.r. of 0·001 when signal B is introduced at a relative level of −1 dB. The difference between the theoretical curve for −1 dB and

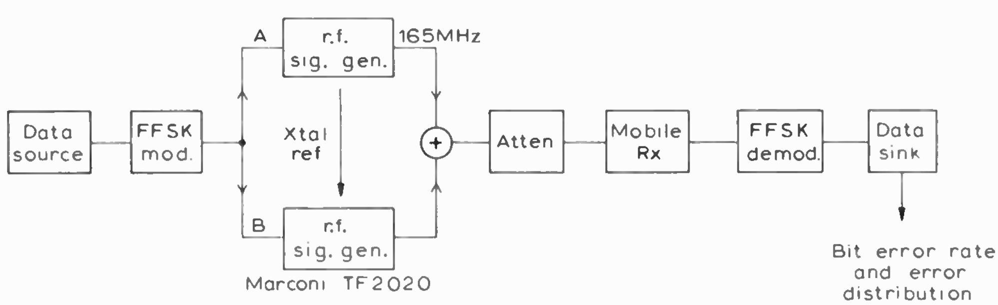

Fig. 3. Error performance measurements with steady quasi-synchronous signals.

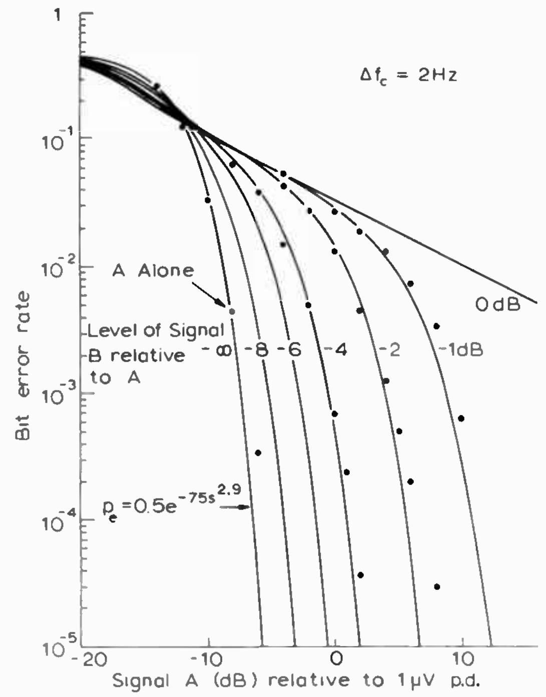

Fig. 5. Stationary f.m. quasi-synchronous error performance. (*Note.* 12 dB SINAD at −9 dB.)

the measured points represents an error in setting the signals of only 0·1 dB, which is well within experimental error.

Nearly equal signal levels clearly result in errors during the time the signals beat to a minimum. However, the time for which the resulting envelope is below the noise level, and therefore the number of errors that occur, is reduced as the absolute magnitude of the signals is increased, leading to a low b.e.r. at high signal levels, even when the signals are equal.

A similar picture was found with the f.m. results, shown in Fig. 5, and again the theoretical curves, generated from the f.m. single signal curve, are a close fit to the measured points with quasi-synchronous signals.

3 Theory for Non-fading Quasi-synchronous Error Rate

It is clear from Fig. 4 that with only one signal present (the 'A alone' points), the b.e.r. is very dependent on

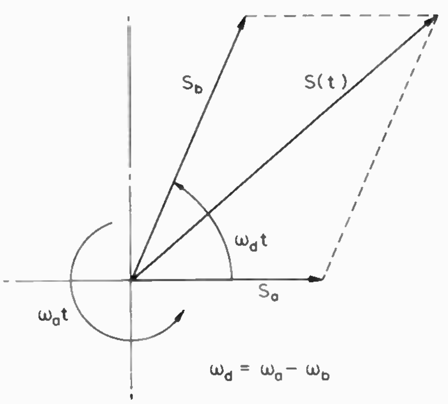

Fig. 6. The resultant of two interfering signals.

signal level, and falls rapidly as the signal level increases. With quasi-synchronous signals the amplitude of the resultant varies with time (as we have seen in Fig. 1). When the signals reinforce the resultant is large and errors do not occur; but when they cancel the resultant is small and errors may result. The theory developed here finds the error probability from moment to moment through the beat period by finding the instantaneous signal level and then using the 'A alone' points to find the error rate expected at that signal level.

The first step is to fit a curve to the measured 'A alone' points, of the form

$$p_e = 0{\cdot}5 \exp(-Qs^R) \qquad (1)$$

using the E04FBA routine[2] where s is the envelope amplitude in microvolts, and Q and R are empirical constants determined by the receiver noise figure, the modulation process and the modem. For the a.m. receiver $Q = 21$ and $R = 2{\cdot}2$.

Next an expression is needed for the instantaneous amplitude of the envelope of the beating quasi-synchronous signals. The two signals are represented in Fig. 6 as simple vectors S_a and S_b (which means we have ignored the modulation components). The frame of reference in the vector diagram is rotating at the carrier frequency, ω_a. The resultant amplitude $s(t)$ is obtained, by trigonometry, as;

$$\begin{aligned} s(t) &= \sqrt{(S_a + S_b \cos \omega_d t)^2 + (S_b \sin \omega_d t)^2} \\ &= \sqrt{S_a^2 + S_b^2 + 2S_aS_b \cos \omega_d t} \qquad (2) \end{aligned}$$

Equation (2) gives the envelope amplitude, and equation (1) gives the error probability at that amplitude; combining them gives the error probability as a function of time as it varies during the period of the beat waveform:

$$p_e(t) = 0{\cdot}5 \exp[-Q(S_a^2 + S_b^2 + 2S_aS_b \cos \omega_d t)^{R/2}]. \quad (3)$$

The average bit error rate is the average of this time varying error probability over one period of the carrier frequency difference. It is convenient to divide the beat period into bit intervals, with the relationship $f_d = 1{\cdot}875$ Hz and $f_b = 1200$ bit/s, giving 640 bits in the beat period. The mean error probability is:

$$\langle p_e \rangle = \frac{1}{640} \sum_{n=1}^{640} 0{\cdot}5 \exp\left[-Q\left(S_a^2 + S_b^2 + 2S_aS_b \cos \frac{2\pi n}{640}\right)^{R/2}\right]. \quad (4)$$

Equation (4) was used to plot the theoretical curves in Figs. 4 and 5, which can be seen to agree very closely with the measured points.

The theory is limited to carrier frequency differences which are much less than the bit rate, so that the envelope is roughly constant during the bit period. However, this is not a practical limitation since typical carrier frequency differences are less than 50 Hz and the usual bit rate is 1200 bit/s. The sensitivity to a difference in modulation timing has not been investigated, although it is unlikely to be more critical than with moving receivers, where a timing accuracy of 0·25 of a bit was adequate with direct f.s.k.[1] The two signals had approximately the same modulation depth.

Within these limitations the theory is accurate and will be used to calculate the error performance with fading signals for the amplitude modulation case. Almost identical results are obtained with frequency modulation as shown by the similarity of Figs. 4 and 5.

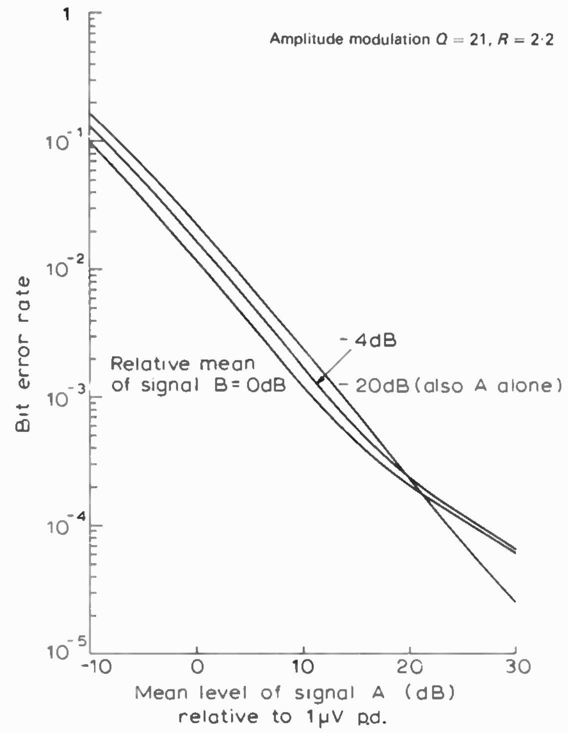

Fig. 7. The mean b.e.r. with fading quasi-synchronous signals.

4 Mean Bit Error Rate for Stationary Users in a Fading Environment

To find the mean of the b.e.r.s experienced by stationary receivers over the whole of the overlap area we must integrate over all the combinations of the two signal levels that can occur. With the Rayleigh fading normally found in mobile radio,[3] the signals S_a and S_b are Rayleigh distributed with the density function:

$$P(S_a) = \frac{\pi S_a}{2m_a^2} \exp\left[-\frac{\pi S_a^2}{4m_a^2}\right] \tag{5}$$

where $m_a = \langle S_a \rangle$, the mean of S_a (S_b has the same distribution relative to m_b). Assuming the fading in the two signals is uncorrelated the mean b.e.r. with fading will be:

$$\langle p_e \rangle_{\text{fading}} = \int_{S_a=0}^{\infty} \int_{S_b=0}^{\infty} [P(S_a)P(S_b)\langle p_e \rangle] \, dS_a \, dS_b \tag{6}$$

where $\langle p_e \rangle$ is from equation (4). It should be noted that $\langle p_e \rangle$ is a function of S_a and S_b. Equation (6) was evaluated numerically on an ICL 1904S in Algol using the numerical integration routine D01AAA.[2] The routine was used recursively for the double integral.

The results are plotted in Fig. 7, as the mean b.e.r. against the mean level of signal A with the relative mean level of signal B as parameter. Introducing the second signal has improved the error performance at low signal levels by about 3 dB but has reduced it at high signal levels.

Although the mean b.e.r. is a useful performance indicator and could be compared with field measurements, it cannot be used to find the mean message failure rate. The mean b.e.r. was found by averaging over many stationary receivers, some of which have a high b.e.r. (because they have small signals or nearly equal signals) and some a low b.e.r. Few actually experience the mean b.e.r. If the mean b.e.r. were used to find the mean message failure rate it would be like fitting a group of people with shoes of average size and then finding on average how comfortable they were. We must return to a bit-by-bit view of word failure and only then integrate over the fading signals.

5 Mean Word Failure Rate with Fading

Recently the Electronic Engineering Association proposed a preferred data transmission system for land mobile radio using a binary format. This format, which is under consideration by the Home Office, uses 64-bit codewords with error detection coding, which enables words containing errors to be discarded. With a single steady signal and constant bit error probability, p_e, the word failure rate is:

$$p_w = 1-(1-p_e)^{64} \tag{7}$$

With quasi-synchronous signals the bit error probability varies during the carrier beat period. With $f_d = 1{\cdot}875$ Hz and $f_b = 1200$ bit/s, giving 640 bits in a beat period, there will be ten 64-bit codewords. Some will be received when the carriers are cancelling and may be lost due to a burst of errors, others will benefit from the carriers reinforcing. The mean word failure rate is:

$$\langle p_w \rangle = \frac{1}{10} \sum_{m=0}^{9} 1 - \prod_{n=1}^{64} \left[1 - p_e\left(\frac{n+64m}{640}\right)\right] \tag{8}$$

where n is the bit number in the word, m is the word number, $\prod$ denotes product, and

$$p_e\left(\frac{n+64m}{640}\right) = 0{\cdot}5 \exp\left[-Q\left(S_a^2+S_b^2 + 2S_aS_b\cos 2\pi\left(\frac{n+64m}{640}\right)\right)^{R/2}\right] \tag{9}$$

Equation (9) gives the sequence of 640 discrete bit error probabilities that occurs as the beat waveform progresses through one period.

Now that we have an expression (eqn. (8)) for the word failure rate at given values of the two signal envelopes, we can proceed to integrate over the fading fluctuations:

$$\langle p_w \rangle_{\text{fading}} = \int_{S_a=0}^{\infty} \int_{S_b=0}^{\infty} [P(S_a)P(S_b)\langle p_w \rangle] \, dS_a \, dS_b \tag{10}$$

where $\langle p_w \rangle$ is from equation (8) and $P(S_a)$ and $P(S_b)$ are from equation (5).

The mean word failure rate with fading is plotted in Fig. 8 against the mean of signal A, with the relative

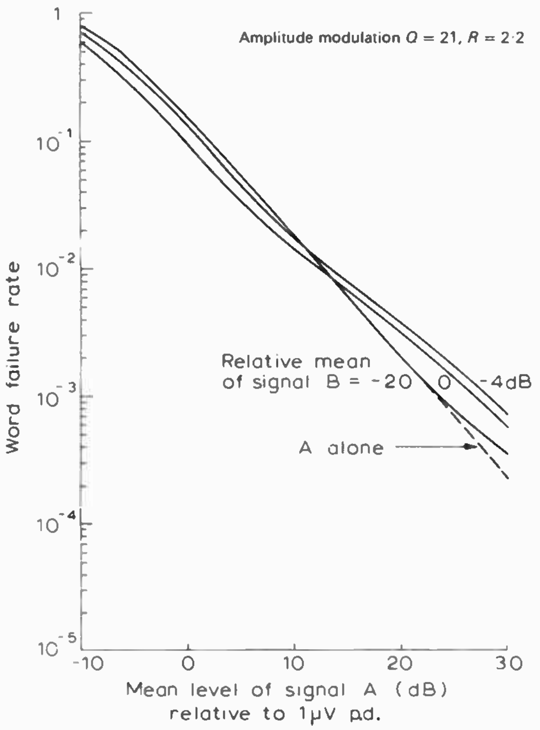

Fig. 8. Mean word failure rate. $f_d = 1{\cdot}875$ Hz, $f_b = 1200$ bit/s and 64-bit words.

mean of signal B as parameter. At signal levels below 12 dB (relative to 1 μV), introducing the second signal improves the performance by about 2·5 dB; useful but not dramatic. At signal levels above 12 dB, the second signal reduces the performance by about 4·5 dB.

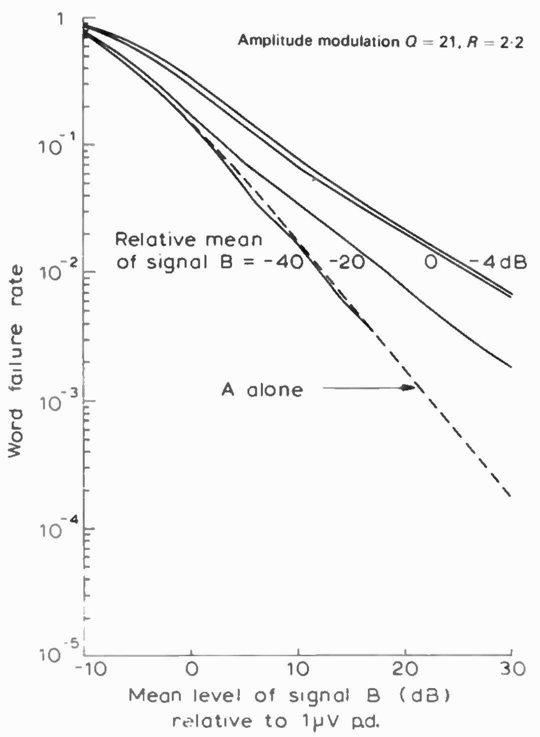

Fig. 9. Mean word failure rate. $f_d = 18{\cdot}75$ Hz, $f_b = 1200$ bit/s and 64-bit words.

However, above 12 dB the word failure rate is less than 1% and so the loss in performance is less significant. As the second signal is removed the word failure rate becomes equal to the performance found for a single signal with the method given in Ref. 4 providing a useful check.

The effect of a greater difference in carrier frequency was found by changing the number of bits in a carrier beat period. With a carrier difference of 18·75 Hz, corresponding to one 64-bit codeword in a beat period, the word failure rates with nearly equal mean levels are worse than with a single signal, as shown in Fig. 9. At low signal levels the performance loss is about 5 dB and rises to more than 10 dB at high signal levels. This is perhaps to be expected with only one codeword in each beat period, since every codeword could have an error burst in it. The performance is not specially worse because of the exact relationship of 64 bits per beat; slightly shorter or longer words could still tend to have an error burst in each codeword. Exact relationships were used to save on computing time, which even so ran into tens of hours. An even greater carrier difference of 187·5 Hz produces only a slight further loss in performance, as shown in Fig. 10.

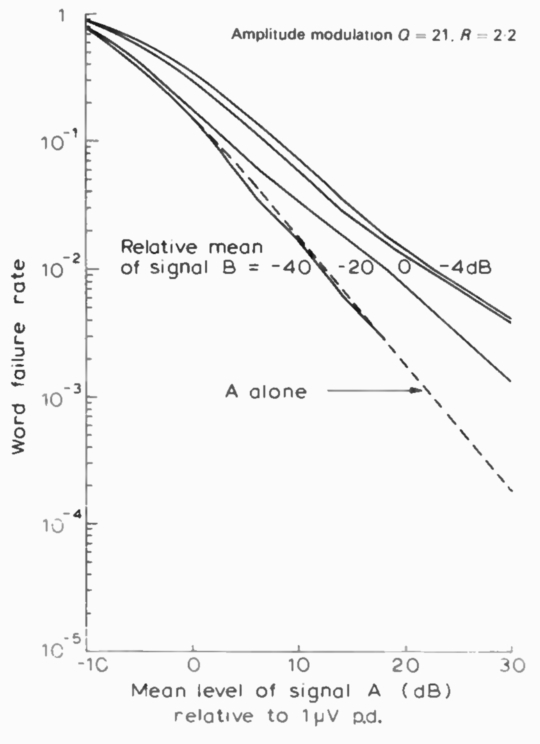

Fig. 10. Mean word failure rate. $f_d = 187{\cdot}5$ Hz, $f_b = 1200$ bit/s and 64-bit words.

6 Conclusions

Close agreement is shown between bench measurements of b.e.r. and a simple theory of error probability for quasi-synchronous data signals, with both amplitude and frequency modulation. The theory starts from the measured error performance for a single steady signal, and predicts the error performance with interfering r.f. carriers in quasi-synchronous operation.

The theory is developed to predict the mean b.e.r. and the mean message failure rate over a large number of stationary receivers in the overlap area between two transmitters, where independent Rayleigh fading of the two transmitted signals occurs.

The 64-bit message failure rate is shown to be 2·5 dB better with quasi-synchronous operation than with a single transmitter, in systems with small mean signal levels (less than 12 dB), with data signals that are accurately timed and have nearly equal modulation depth, for the case of carrier frequency differences of a few hertz. In systems with high signal levels the quasi-synchronous performance is 4·5 dB worse than with a single transmitter.

At the higher carrier frequency differences of 18 Hz and 180 Hz the performance is degraded by between 5 and 10 dB.

The message failure rate will depend on the relative length of the message codewords and the beat period. For comparable performance to single transmitter working the length of the codewords should be only one-tenth of the beat period.

At stationary receivers the error bursts come at the rate of the r.f. carrier difference frequency. This fact could be used when organizing the transmission of long messages and of repeat messages to obtain the best message throughput.

7 Acknowledgment

It is a pleasure to acknowledge the contribution of my colleagues P. J. Mabey and T. W. Whiter.

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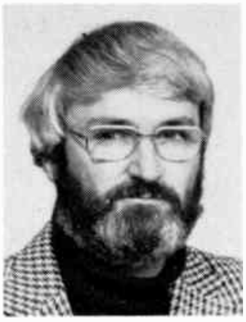
The Author



Dick French (Fellow 1980, Member 1964, Graduate 1959) who is with the Systems Division at Philips Research Laboratories, Redhill, studied telecommunications at Norwood Technical College. He received the Ph.D. degree in 1973 from the CNAA for work on speech scrambling and synchronization techniques undertaken at PRL and Brighton Polytechnic. After two years of national service in the RAF he joined Muirhead and Co. and worked on facsimile equipment. In 1960 he moved to the then Mullard Research Laboratories, Redhill, where he has worked on communications and signal processing projects in the Communications Group; his current work is on data transmission in mobile radio networks. The paper in this issue is the fifth he has contributed to the Journal and he has given several other papers at IERE Conferences.

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Indexing Terms: Fibre optics, Applications, Transducers

Optical fibre transducers

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SUMMARY

Light travelling in an optical fibre may be directly modulated by external parameters such as pressure, strain, temperature, etc. Thus the fibre may form the basis of a transducer involving neither moving parts nor electrical connections. Another class of transducers uses the fibre as a means of guiding light from a monitoring point to the measuring point where the light is externally modulated. This paper presents a review of the current status of optical fibre sensors and discusses their future potential.

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1 Introduction

The potential application of optical fibres as transducers has only been recognized within the last decade.[1,2,3] However, a considerable amount of research and development work has been completed over the last five years, and it is the aim of this paper to review this work and examine the true potential of the fibre sensor.

There are a number of potential advantages in the use of fibres as sensors. The most obvious is the lack of electrical potentials at the sensor, thereby removing electromagnetic interference problems and numerous safety considerations for operation in hazardous areas. The fibre is a low inertia transducer, which allows it to respond to stimulus changes over a very wide frequency range. It is chemically inert, and can transmit over a very long distance—of the order of a kilometre—without serious attenuation. There are other possibilities, as yet unexploited, which include the use of optical signal processing techniques to provide passive 'smart sensors' and multiplexing architectures which will allow both sensor and data transmission channels to use the same fibre path.

There are numerous applications of fibre optic sensing. Microphones and hydrophones have received much attention[4,5,6] and the fibre optic gyroscope[7,8] is probably the other principal research topic at present. Fibre sensors[9] have also been reported to measure strain,[10] temperature,[11] magnetic fields[10] and acceleration.[13] There is also considerable activity, little reported in the literature, in developing other simpler sensors for use in immediate applications.

This paper reviews the basic principles of light modulation in fibres and describes a few specific systems in more detail. It will become apparent that there is considerable scope for the development of simple but very effective transducers as well as more complex systems relying on advanced optics. Finally, some of the more advanced future concepts are described, and the applications potential is discussed. The overall conclusion is that a wide range of optical fibre transducers are feasible with current technology, but many of the more advanced devices await the arrival of suitable components (many of which are currently under investigation by PTT operators) using both fibre and integrated optic building blocks.†

2 Modulation of Light in Optical Fibres

A light beam can be modulated in five of its basic properties: optical intensity, phase, polarization, wavelength and spectral distribution. Most fibre sensors utilize the first three, though there is considerable potential in exploiting the last two. It is useful to consider briefly the optical requirements for all of these.

2.1 Intensity Sensing

This modulation technique is very simple, and uses the arrangement shown in Fig. 1. Modulation is effected either by bending the fibre[14,15] or by introducing some form of shutter which moves across the fibre. The shutter

† *Note added in proof*—Detailed report on Fibre Optic Transducers recently made available by ERA Technology, Cleeve Road, Leatherhead, Surrey.

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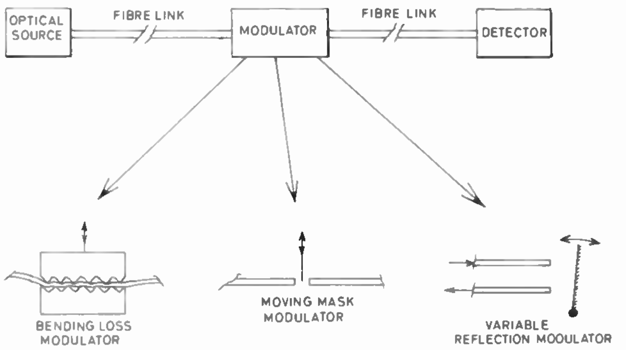

Fig. 1. The essential features of an intensity modulation optical-fibre transducer.

may be implemented either by relative motion of the input and output fibres or by motion of a mask or reflector relative to fixed fibre ends.

The principal complication in systems using this modulation technique results from the fact that optical sources are all prone to intensity drift with ageing and so a reference channel is *always* needed. This can usually be designed into the system with a little ingenuity.

2.2 Phase Modulation

Detection of optical phase changes offers a means of measuring very small variations in physical parameters, but must be performed interferometrically, which implies that a stable reference is required. The basic arrangement of one interferometer (the Mach–Zehnder) is shown in Fig. 2. In the heterodyne form, the Bragg cell provides a frequency shifted reference beam and the phase shift is detected at the difference frequency. Difference frequencies of a few tens of MHz are typical. This is usually the most convenient form of the interferometer, though a homodyne version with a complex feedback bias arrangement has been described.[16]

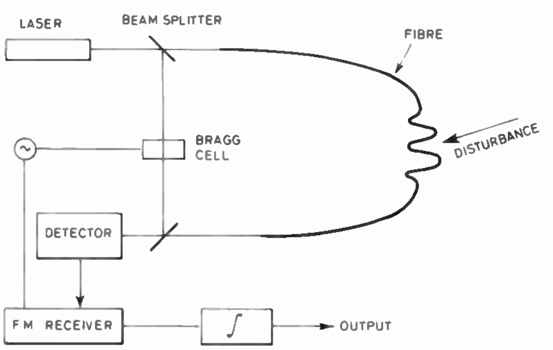

Fig. 2. The basic Mach–Zehnder (M–Z) optical-fibre interferometer.

Phase modulation schemes also require a coherent source, which must be a laser. HeNe gas lasers are the most convenient and economical, though semiconductor lasers[17,18] may be used, but with due caution. Finally, it should be mentioned that interference only occurs between waves of similar polarization, and changes in the output polarization state from the fibre will alter fringe contrast, and thus either the signal/noise ratio for the heterodyne interferometer or the calibration for the homodyne system.

The preceding discussion has also implied the use of monomode optical fibre, with consequent mechanical constraints (the core of a monomode fibre is only a few microns in diameter). The mechanical problem is eased in multimode fibres, and both heterodyning[19] and homodyning[20] interferometers have been investigated using multimode fibre in the signal arm. These systems replace the alignment requirements of single mode interferometers with a fading and signal distortion problem and a complete lack of any sensible low-frequency capability.

A third multimode phase sensitive system exploits the 'Fibredyne' process. This is very simple (Fig. 3) but is difficult to both linearize and stabilize. However, it has been used as the basis of both a direct[22] measuring flowmeter and a vortex shedding meter.[23]

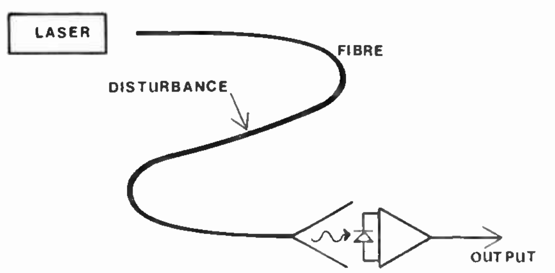

Fig. 3. The 'Fibredyne' interferometric single fibre sensor system.

Phase detection is currently used primarily for monitoring oscillatory quantities (acoustic fields, etc.). Measurement of slowly varying parameters—mechanical strain, temperature, etc.—is hampered by interference from slow drift of both the optics and the electronics. Compensating arrangements, incorporating self-correction of signal and reference arms of the interferometer are, however, feasible. Much effort has been expended on the fibre optic gyroscope, which is a particularly elegant example of a self-compensating interferometer. However, compensating systems are also currently under investigation for the monitoring of other parameters, and a number of these approaches appear to offer very promising potential.

2.3 Polarization

Variations in light polarization are difficult to interpret in optical fibre transducers, since the fibre is often itself inherently birefringent and this birefringence varies with time. However, with special fibres,[24] sensors which rely on detecting polarization variations can be built. The principal application is in measuring Faraday rotation induced by a magnetic field produced by a large (≈ 1000 A) electric current. The basic detection scheme[25] is shown in Fig. 4. Faraday rotation simply produces a polarization rotation in the fibre proportional to the magnetic field produced by the wire.

The extent of the polarization rotation is given by:

$$\phi = V \oint \mathbf{H} \cdot d\mathbf{L} = V . NI,$$

where I is the electric current, V the Verdet constant ($3{\cdot}3 \times 10^{-4}$ degrees/ampere-turn in silica), ϕ the angle of rotation and N the number of turns of fibre around the

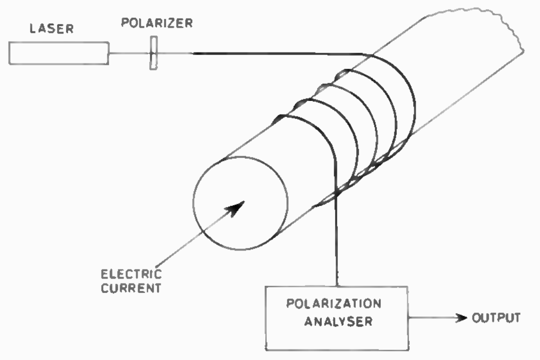

Fig. 4. A polarization modulation current detection scheme. The plane of polarization of the light leaving the single-mode fibre depends on the current through the conductor.

current carrying conductor. For a typical system, using ten turns of monomode fibre, we find that the polarization rotation is $3{\cdot}3 \times 10^{-3}$ degrees per ampere. Polarimeters with a resolution of 0·1 degrees are readily available, giving a current resolution of 30 A.

2.4 Wavelength and Wavelength Distribution Modulation

One form of wavelength modulation is simply Doppler shift which may be produced by, for instance, backscatter from moving particles at the end of the fibre. This forms the basis of the fibre optic Doppler anenometer,[25] which does of course require a coherent source (Fig. 5).

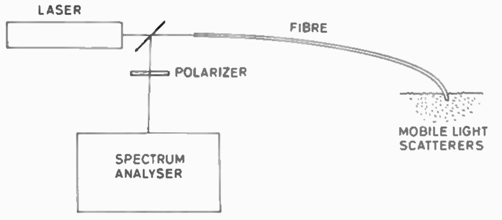

Fig. 5. Fibre-optic Doppler anemometer. The polarizer is adjusted to eliminate reflections from the launch end of the fibre, and thus the frequency of the signal on the detecting electronics depends purely on relative motion of the far end of the fibre and the mobile scattering targets near the end of the probe.

A more general form of wavelength modulation may occur when incoherent light is transmitted along a large core fibre to phosphorescent material. Light emitted from the phosphor may be at a different wavelength. The wavelength and/or intensity of the return light will be a function of the parameter to be measured, and hence remote monitoring may be performed. Clearly much depends on the stability of both the light source and the phosphor.

One example of such a system is the 'Fluoroptic' temperature sensor, now available as a commercial product. A rare earth phosphor is illuminated by white light along a short (few metres) length of large cor optical-fibre. The light excites the phosphor which emits a number of lines. Two of these lines are selected using the filters in the receiver (see Fig. 6). The ratio of the intensities of the chosen lines, which are at 540 and 630 nm, is a single valued function of the temperature of the phosphor. Thus, by measuring this ratio, an exact

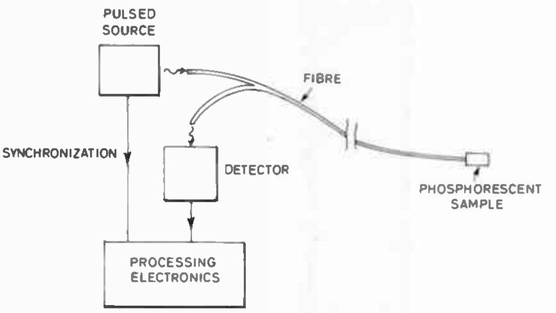

Fig. 6. Schematic of a phosphorescent sensor. The phosphor produces, for instance, a decay time dependent on temperature.

measure of the temperature—which is independent of the output light intensity—may be made. Resolution of 0·1°C over the range −50 to +250°C is possible. The instrument is, in practice, rather complex—requiring calibration during thermal stabilization since the sensitivities of even matched photodetectors do not track at the two different wavelengths as temperature varies. The device is well suited to medical applications—for instance monitoring tissue temperature—due to its small size and rapid response time.

Wavelength distribution modulation is very similar. A distinction may be made by taking as an example, radiation from a black body. This radiation varies fairly slowly with temperature, but optical fibre probes may be used to monitor the variations. One well-known form is an optical fibre pyrometer, which simply uses a fibre to guide light from a hot zone to the instrument, a normal pyrometer with slightly different receive optics.

Another example is to use the radiation itself as the signal to an electronic pyrometer. The pyrometer may take many forms, but the task is, in essence, one of simple spectral analysis. In the case of a black body temperature probe, for use at, say, temperatures in

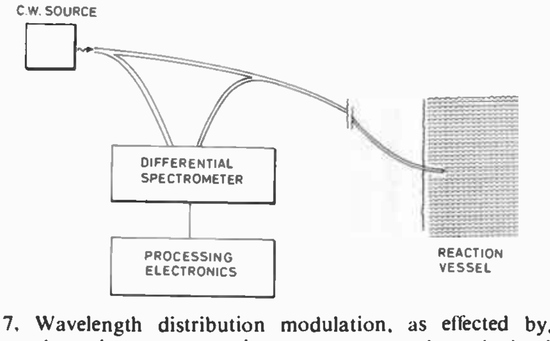

Fig. 7. Wavelength distribution modulation, as effected by, for instance, dyes whose concentration, temperature or impurity levels are to be measured.

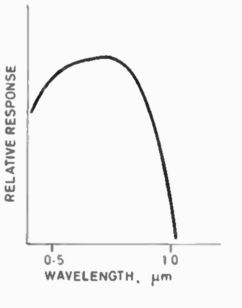

Fig. 8. Spectral response of a typical silicon photodiode indicating that wavelength to electrical output conversion may be expected, especially in regions where the output is strong function of wavelength.

excess of 200°C, it is possible to use the spectral filtering properties of the photodiode itself as the basis of the analyser. The basic probe structure is shown in Fig. 7, and the way that variations in the input spectrum to the photodiode are converted to an overall response which depends on temperature, is shown in Fig. 8.

2.5 Discussion

This Section has briefly summarized the available techniques whereby light guided by an optical fibre may be modulated and some of the means whereby that modulation may be detected. Interferometric techniques are doubtless the most complex, but offer probably the most potential. However, simple intensity and/or wavelength modulation provides the basis of a large variety of sensing functions. Some specific examples of transducers investigated to date follow.

3 Examples of Specific Sensors

This Section will indicate the diversity of functions which may be performed by optical fibre sensors. It is by no means an exclusive account!

3.1 Vernier Positional Indicator

A possible configuration for an optical fibre position sensor is shown in Fig. 9. The device is essentially a vernier device where a line of nine input fibres illuminates ten output fibres. With appropriate attention to the input and output fibre spacing, resolutions of the order of 0·05% of the fibre diameter should be attainable by suitable processing of the output intensities. Some computational facility should be available to compensate for component drifts. This device is as yet undeveloped but a much simpler device, a strain gauge 'tell-tale'[29] using an array of variable breaking strain fibres is at the prototype stage.

3.2 Optical Fibre Microphone

Numerous schemes have been suggested to realize an optical microphone. These come into two categories, using modulated reflection or transmission and using phase modulation (see Sect. 3.3).

A very simple, but effective, reflection modulation microphone—suitable for use in telephony—exploits the principles shown in Fig. 10.[30] Movement of the diaphragm modulates the amount of light reflected back along the illuminating fibre, and performance acceptable for telephony has been achieved. Modulation of the incident light intensity by a few percent may be achieved with a diaphragm displacement of fractions of a micron. It is interesting to note that sufficient optical power (a few milliwatts) can be transmitted along the fibre from an exchange to a subscriber to sound an audible alarm at the subscriber's end, so that all-optical telephone links become feasible.

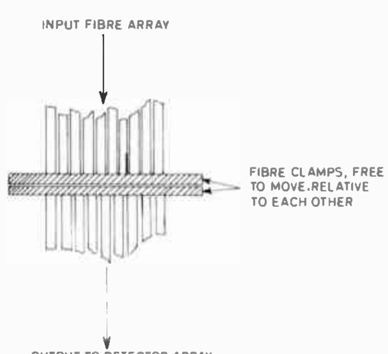

Fig. 9. A proposed fibre-optic vernier relative position monitor. The ten input fibres illuminate nine outputs. With suitable signal processing, resolution of the order of 0·05% of the fibre diameter may be expected.

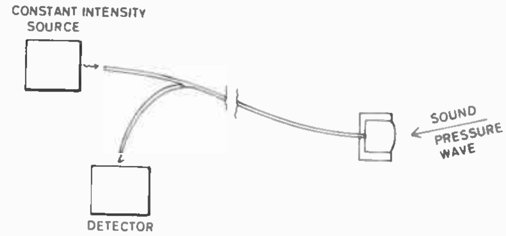

Fig. 10. A simple fibre-optic reflection microphone. A number of variations on this basic concept are described in Ref. 30.

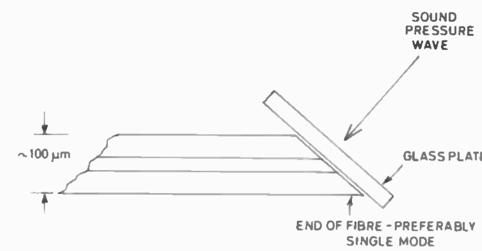

Fig. 11. One possible version of a frustrated internal reflection modulation technique. The glass plate is placed close to the angled fibre end. The coupling between the two is a function of separation, and thus the reflected power is modulated.

An alternative scheme, based on frustrated internal reflection (Fig. 11) has recently been proposed.[31] The end of an optical fibre is cut at the critical angle, and another glass plate placed almost in contact. There is evanescent wave coupling between the two components depending on the plate separation. The reflected power may be therefore modulated by varying this separation.

There are many other techniques available. One reflection technique[32] uses a central illuminating fibre surrounded by six return fibres. Sub-nanometre resolution is claimed for displacement of the reflector along the fibre axis.

3.3 Phase-sensitive Acoustic Sensors

The phase of light passing along an optical fibre may be varied by changes in environmental parameters such as pressure, temperature or strain. The phase sensitivities, per metre of fibre, are of the order 100 rad/°C, 10 rad/bar (1 bar = $1{\cdot}018 \times 10^5$ N m^{-2}) and 10 rad/microstrain.

Phase detectors can, quite readily, be made to detect 10^{-4} radians. With care, 10^{-8} radians resolution can be achieved.

Optical fibre hydrophones may be readily fabricated from a coil of single mode fibre—Fig. 12. (Multimode fibre may also be used with some penalties in noise and stability.) This fibre forms one arm of a heterodyning

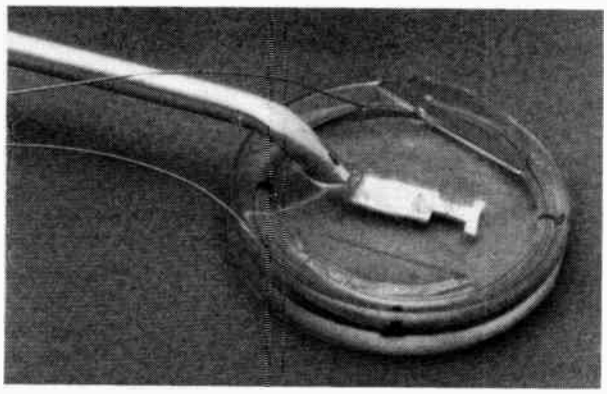

Fig. 12. Experimental optical fibre hydrophone tested at University College.

Mach–Zehnder interferometer, and sensitivities superior to piezoelectric devices have been reported.[33] The sensitivity may be varied by using specially coated fibres,[34] but there remain a number of problems. The sensitivity of the hydrophone should increase linearly with interaction length, but our own measurements (see Fig. 13) have shown a saturation effect.

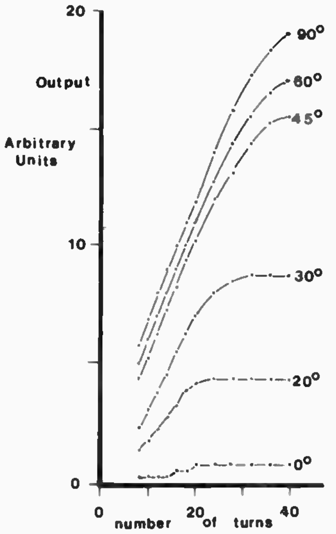

Fig. 13. Dependence of fibre hydrophone response on number of fibre turns in the sensing coil.

These measurements indicate that a simple account of the interaction between a pressure wave and an optical fibre, in which one postulates that the magnitude of the interaction increases linearly with the length of fibre in the acoustic field, is not accurate in practice. This saturation may have fundamental implications in the design of fibre-optic hydrophones. In particular, differences in these characteristics as a function of orientation of the hydrophone coil in the acoustic field and of the dimensions of the coil of fibre (in acoustic wavelengths) have been observed, and require further understanding.

There is an additional problem, in that the fibre to and from the hydrophone proper is also, inevitably, acoustically sensitive. Often, these fibre leads may be required to be longer than the fibre in the hydrophone itself. There are a number of possible solutions to both these problems, but a full evaluation of the true potential of the fibre-optic phase-modulated hydrophone is still some way in the future.

3.4 The Optical-fibre Gyroscope

The optical-fibre gyroscope is based on exploitation of the Sagnac effect, originally described more than half a century ago.[35] The basic principle is shown in Fig. 14. Light from the laser is split into two equal components launched in opposite directions into the fibre loop. If the loop is rotated, light travelling with rotation will remain in the fibre for a slightly longer time than light travelling in the opposite direction.

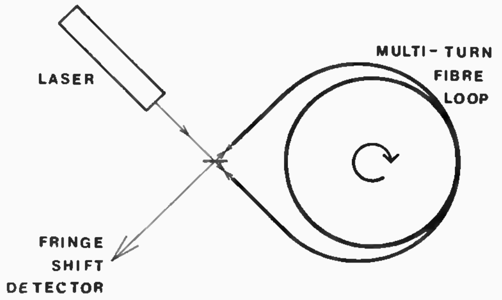

Fig. 14. The basic principle of the Sagnac effect optical fibre gyroscope.

An intuitive theoretical description of the gyroscope gives the phase difference $\Delta\phi$ between the two emerging beams as

$$\Delta\phi = \frac{4\pi LR}{\lambda c}.\Omega,$$

which is the non-relativistic limit of the device characteristics. The fibre gyroscope could be an inexpensive, fairly low quality disposable device or, perhaps in the future, a high sensitivity but complex, inertial navigation instrument. Current experimental performance figures attain a sensitivity of the order of 1 deg/h,[36,37] though it is theoretically possible to obtain significantly better performance. Detailed understanding of the system physics is still incomplete,[38] and doubtless future improvements in system configuration and component design will achieve orders of magnitude higher sensitivity levels.†

The optical fibre Sagnac interferometer is a particularly interesting device in that the 'reference' and 'signal' arms of the interferometer travel through the same length of fibre, and thus are identical, apart from any non-reciprocal effects which may occur along the path. The only non-reciprocal propagation effects are the Sagnac effect and the Faraday effect—the latter introducing non-reciprocal circular birefringence and the former non-reciprocal linear birefringence. The two effects should thus be separable (even if they do occur together), and so it becomes necessary to investigate whether there are other possible sources of non-reciprocal propagation. Lorentz reciprocity states that if the input and output from the fibre loop are in the exact same modes, then propagation in the two directions will be exactly identical. Most of the research into optical fibre gyroscopes is aimed at achieving this condition and then detecting the resulting small phase changes (due to

† *Note added in proof*—Sensitivity of 0·022 deg/h has recently been demonstrated. This is very close to the theoretical limit.[57]

rotation) of the order of one microradian. Reciprocity conditions must be achieved to better than this, and since a typical fibre gyroscope involves the order of 1 km of fibre, this implies that the paths in each direction through the loop must be identical to better than one part in 10^{17}!

3.5 Thermometric Probes

Optical fibres may form the basis of a wide variety of thermometric probes, many of which can be so tiny they may be inserted into a hypodermic needle. Interferometric probes have as yet been relatively unsuccessful due, ironically, to excessive sensitivity. Recently, single fibre interferometers using the temperature dependence of differential delay between orthogonally polarized modes in birefringent fibre,[39] as indicated schematically in Fig. 15, offer potential solutions.

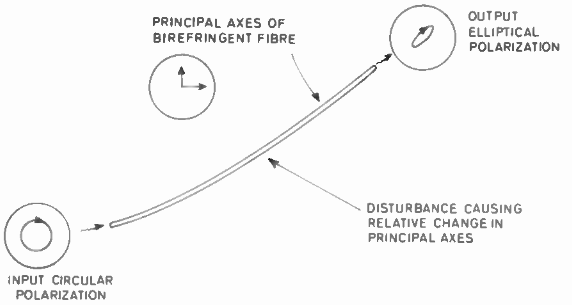

Fig. 15. Polarization modulation using birefringent optical fibre as the modulating medium. Differential phase-modulation imposed along the two principal axes of the fibre change the output polarization state.

The most promising thermometric probes are the self-luminescent and phosphorescent[40] types, shown schematically in Fig. 6. Observation of temperature-dependent delay time of phosphorescence is the one possible principle. Rugged, compact and accurate fast response thermometers may be fabricated. (See Sect. 2.4.)

3.6 Discussion

The preceding examples have barely scratched the surface of optical fibre sensors. However, the principal modulation techniques have been illustrated. The design of a real transducer reduces to the engineering problem of enhancing the modulation sensitivity to the required parameter and reducing its sensitivity to other parameters. Successful solutions to these problems will involve research into materials, optical architectures and electronic and optical signal processing.

For the wide variety of intensity modulation sensors, the required development falls more into the realm of transducer engineering development than optics research, and much has already been achieved on this basis. The general conclusion, borne out in our own experiments and by other laboratories, is that intensity sensors can themselves be remarkably sensitive. In very general terms, if an interferometric sensor may be configured with a particular sensitivity, then an intensity-modulated sensor for the same parameter may be designed to a performance level within a couple of orders magnitude of the interferometric device. Of course, the modulation principles will be totally different so that this is really nothing more than a very general rule-of-thumb. However, since interferometric sensors are very sensitive, this indicates that an adequate (at least for sensitivity) intensity-modulated device may often be designed.

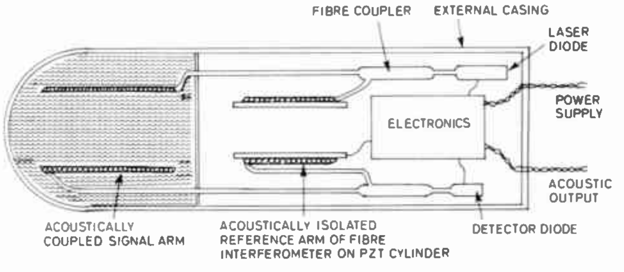

Fig. 16. Schematic diagram of a prototype optical-fibre hydrophone developed by the Naval Research Laboratory.

Most coherent (interferometric) sensors are at a relatively early stage in their development.[43] It is only recently (August 1981) that the US Naval Research Laboratory[44] tested a prototype ruggedized optical-fibre hydrophone. This device, sketched in Fig. 16, functioned very well, but note that it does require a power supply and that the output from the hydrophone is electrical. The problem with sensitivity of fibres to and from the sensor remains, and the NRL prototype is recognized as an intermediate step to the true all-fibre interferometric sensor. The Sagnac interferometer is the only presently used truly self-compensating arrangement (Sect. 3.4), and even this must be carefully adjusted with respect to launch and receive polarizations. The Sagnac configuration is clearly capable of detecting parameters other than rotation, and use of the fibre ring interferometer as an alternative current measuring device has recently been reported.[45] It should also be noted that the Faraday rotation current sensor[24] has enjoyed considerable success and prototype production devices are currently under construction.

4 Future Prospects

It will probably be some time before the full potential of optical fibre sensors is realized. There will be the inevitable engineering development of intensity-based sensors to take them into the market-place, but exciting conceptual advances remain to be made in the realm of interferometric sensors at the component level and in all optical sensors at the system level.

A number of important component developments currently under way will greatly influence the future of interferometric fibre sensors. Several investigations into enhancing the stability of single-mode semiconductor lasers using analogues to phase lock loops and similar configurations are under way.[46,47] These may be either used to lock the laser frequency or the error signal may itself provide the sensor output when the fibre forms, effectively, part of the laser cavity. Figure 17 shows one such arrangement for ensuring frequency stability of the laser.

Birefringent fibres and, more so, associated polarization selective components, will allow significant

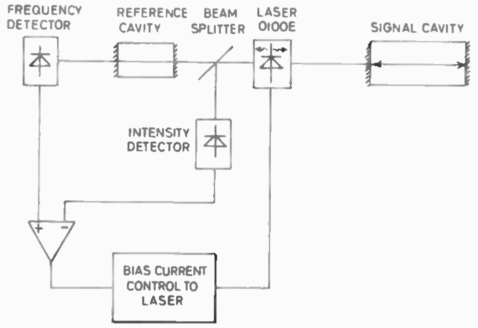

Fig. 17. A technique to perform phase locking on a semiconductor laser. The feedback signal may be used as the output of an optical phase nulling sensor.

additional flexibility in sensor configuration. In particular, polarization rotation (Fig. 15) could form the basis of a new family of lead-insensitive, compact interferometric sensors. Initial simple estimates indicate that a strain gauge of total dimensions much less than 1 mm could be readily fabricated. The device requires polarization-selective couplers, but should be within the scope of integrated optics processing.

This technology of integrated optics[48] will play a necessary role in future PTT systems and the techniques thus developed will be directly applicable to interferometric sensing. It is likely that one form of the strain gauge would be as a fully-integrated optic Mach–Zehnder (M–Z) interferometer, and indeed a conventional M–Z interferometer could be configured as a wide variety of sensors.

It is similarly likely that integrated optics will also provide essential components such as local oscillator modules for heterodyne systems[49] and splitter, combiner and phase bias components for a variety of applications, most notably at present the gyroscope.[50]

There are many components in a fibre sensor system. For interferometric devices, interface problems also require completely satisfactory solutions, particularly from semiconductor laser to waveguide, and from waveguide to fibre. Among the most promising approaches is the use of an anisotropically-etched V-groove in silicon as the locating jig from monomode fibre to monomode guide.[51] Semiconductor lasers with lens-ended single-mode fibres pigtailed permanently in place are now becoming available,[52] and it seems likely that direct planar launch from lasers into integrated optics will similarly be soon available.[53]

It is, however, at the system level that optical fibre sensors offer really significant advantages. Already, the geometrical flexibility of the fibre hydrophone allows for unique variations in polar and frequency response of underwater acoustic monitoring systems.

Numerous architectures have been suggested to allow multiplexing of many fibre sensors on to one fibre loop, and one of these is shown schematically in Fig. 18. Variations include wavelength multiplexing and sensor operational frequency multiplexing. As a final step in the multiplexing chain, it is perfectly feasible to combine this sensor multiplexing with a Fibredyne[21] data link along the same fibre thus using the fibre path and component set to its utmost.[56]

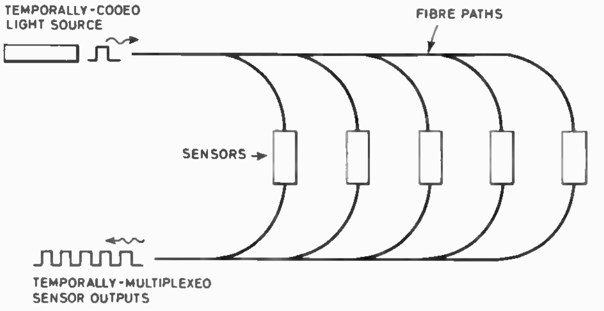

Fig. 18. One proposed multiplexing scheme for optical-fibre sensors.

Finally, the fact that the optical medium is well suited to performing instantaneous Fourier-related[54,55] signal processing on spatial signals has, to date, been completely ignored in sensing systems. This capacity can be used to encode spatial data, perhaps concerning the position of the reflector in an incoherent system, into another form—possibly digitized—by the use of suitable correlation filters. It may also be used to perform pattern recognition on to fault conditions or to change the form of the spatial data into a more readily recognized format—for instance a spectral representation.

5 Conclusions

This has been a wide ranging, and therefore often superficial, account of the prospects for optical fibre sensing. In the final reckoning, the eventual success of optical fibre transducer techniques will depend on their position in comparison to other, often well-established, measurement techniques. Comparisons may be made in terms of cost, sensitivity and applications suitability. Present-day conventional sensors are remarkably, and necessarily, economic, but it does seem that both the sensitivity of fibre sensors and the applications flexibility will prove to be attractive features. At present, fibre sensors are well suited to applications requiring spark-proof devices, for instance in petrochemical plant, even though sensitivities are probably comparable to those achievable with established techniques. In the future multiplexed transducers on to the same fibre path, especially with a parallel data route, coupled with on-line instantaneous signal processing will offer unique flexibility to the system designer.

The possibilities are remarkable. The bounds are imposed only by the designer's imagination and, in some, but not all, cases by the limits of available technology. There is, without doubt, an interesting future for optical fibre sensors and systems.

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Indexing Terms: Radio receivers, Radio broadcasting, Data transmission equipment

The impact of radio-data on broadcast receivers

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Based on a paper presented at the IERE Conference on Radio Receivers and Associated Systems in Leeds in July 1981.

SUMMARY

For some years the BBC and other European broadcasters have been investigating systems for broadcasting radio-data labels from v.h.f. radio-broadcast transmitters. These data signals are primarily intended to aid radio listeners in tuning their receivers to a desired station or programme. Other applications, such as the transmission of simple messages, are also envisaged.

Although no firm specification for radio-data has yet been agreed by European broadcasters, recent international field-trials of various proposed radio-data systems were a significant milestone in the development of a standardized radio-data system. These field-trials also confirmed that reliable radio-data reception is possible with low-cost receiver/decoders even under very adverse reception conditions.

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1 Introduction

Many radio listeners find it difficult to tune their receivers to a desired station or programme and thus are unable to make best use of the available wide choice of national and local radio stations. This is especially true at v.h.f. where any given network or station may be broadcast on a large number of different frequencies. Because of this the motorist with a v.h.f. car-radio may need to retune his receiver several times during a journey as the car moves from the service area of one transmitter to another.

Receiver manufacturers are, of course, continually making improvements in their products by applying new technologies. Thus, some domestic receivers now have frequency-synthesizer tuning with a digital frequency read-out (making tuning more precise and repeatable) and the facility for storing and recalling a number of user-preselected frequencies. And at least one car-radio manufacturer now offers a receiver which, by using a stored list of possible carrier frequencies (entered by the user) for up to six networks, will automatically search and tune to the strongest signal for a particular network, automatically retuning as necessary during a journey.

But none of these receivers can completely free the listener from the chore of looking up the frequencies of the transmitters and presetting the receiver to these. This task is tedious for most listeners and impossible for some.

Radio-data was proposed[1] to help overcome these tuning difficulties by providing a positive means of identifying a broadcast station using a data label.

2 Evolution of Radio-Data

As conceived, radio-data was simply a means whereby data labels identifying the station and programme could be transmitted along with the normal broadcast programme signal. These labels would be imperceptible to listeners with existing receivers but suitably equipped future receivers would decode the data labels and display them, thus affirming the station to which the receiver was tuned.

This concept has now, of course, evolved and it has become clear that, as well as providing a means of affirming manual tuning, radio-data could also be used to automate completely the operation of tuning. Thus the listener could select the desired station or network directly by entering its name (or a simple code) via a keypad. The receiver would then search to find the strongest broadcast bearing that data label. Used in this way, radio-data could completely free the listener from the need to be concerned with wavelengths or frequencies.

Clearly radio-data could also provide a means for transmitting simple messages for display such as the title of the programme being received—there are many possibilities here. But its capacity is necessarily restricted and it could never, of course, rival the service provided by teletext.[2] Thus its primary purpose must be as an aid to tuning.

The problem for the broadcaster has been to devise a way of adding these radio-data signals so that they cause

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no impairment to normal programme reception on existing receivers and yet can be reliably received out to the fringes of the service areas, even under difficult reception conditions. A further constraint is that this must be achieved with a low-cost receiver/decoder.

The BBC[3] and other European Broadcasters[4,5] have investigated several possible systems for transmitting radio-data signals from v.h.f. radio-broadcast transmitters.† All these systems use a low-level subcarrier (added to the stereo multiplex signal at the modulation input to the v.h.f. transmitter (see Fig. 1 and Sect. 4.1)) to convey the data signal. Although no firm specification has yet been agreed, international field-trials of the various proposed systems took place in Switzerland in September 1980 (see Sect. 4.2) and some general principles have now achieved widespread acceptance amongst European Broadcasters.

This international work continues and is presently centred mainly upon the various kinds of message to be transmitted in a radio-data system.

3 Kinds of Radio-Data Message

3.1 Needs of Different Categories of Receiver

We may define three broad categories of receiver: fixed domestic receivers (including hi-fi systems and music-centres), mains/battery portables, and car-radios. Clearly each category has different requirements from a radio-data system. In fixed domestic receivers, the emphasis might be on displayed information (network name, programme title, etc.) rather than on automatic search tuning. In car-radios, a display is unlikely to be desirable (although the use of voice-synthesizers has been proposed) but the ability to tune and re-tune automatically to a particular selected network during a journey is clearly of value. In portable receivers low cost and power consumption are of paramount importance. The design of the system must permit the use of simple low-cost decoders in these applications.

† The BBC have also developed a system for broadcasting data signals from l.f. transmitters but the data channel capacity there is necessarily very restricted and it is unlikely, therefore, to be used in the same way as the v.h.f. radio-data system described in this document.

Table 1 gives a summary of the various kinds of message which are broadcast in the present (April 1981) experimental BBC system (see Sect. 4) and are thought to be representative of the kinds of message which would be broadcast in an operational system. Indication is given in the Table as to which categories of receiver are thought most likely to use the various kinds of message and estimates are given of the message repetition rates needed.

These messages may conveniently be divided into two kinds: information for control and information for display.

3.2 Information for Control

These messages are all short machine codes which are intended to be used as control signals in various automatic operations such as search-tuning. Because they are short they can be repeated frequently without taking up too much of the available channel capacity.

3.2.1. *Network identification code*

This is a label (e.g. 0134) which is used to identify a particular network (i.e. programme source); all transmitters broadcasting the programme signal of a particular network would carry the same network identification code. Thus an automatic search receiver would look for the strongest signal bearing the identification code of the selected network. A high repetition-rate (about 10 times per second) is needed to minimize search times. Search receivers with two front-ends, one searching whilst the other provides the sound programme signal, have been proposed[3] to decrease search-times.

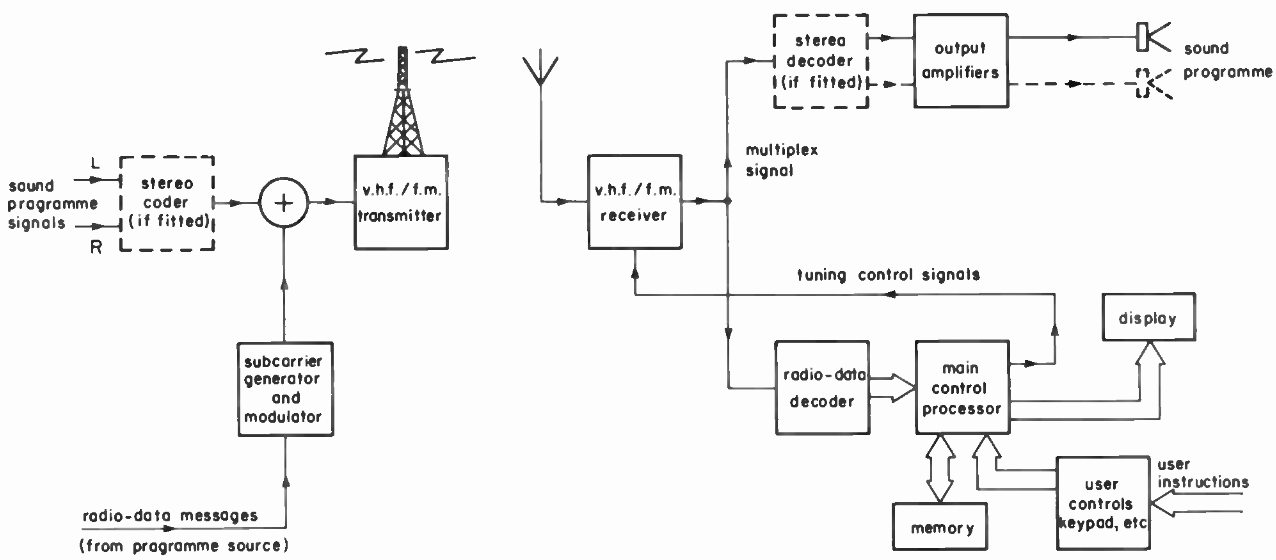

Fig. 1. V.h.f. radio-data transmission and reception.

Table 1. Kinds of radio-data message in the present BBC experimental system

Message content	Example	Application: Control (C) or Display (D)	Primary User: Fixed domestic (FD) Portable (P) Car-radio (C)	Repetition rate per second	Number of bits in code (or number of characters)	Remarks
Network identification code	0134	C	P, C	~10	16 bits	The first digit signifies the broadcasting country, e.g. 0 for U.K.
Programme type code	0 (for news)	C	C	~10	4 bits	Includes traffic announcements and an alarm code for public emergencies.
Programme item number code	31 5 14 24	C	FD	~1	20 bits	Schedule start time of programme as week-of-year day-of-week, hour-of-day, minute-of-hour.
Decoder control code	11	C	All	~1	5 bits	Includes one bit to identify whether it is speech or music being broadcast.
Network name	BBC LON	D	FD, P	~1	7 characters	7-bit ASCII characters as defined in ISO 646
Future applications yet to be defined	—	C and D	All	—		More than 60% of the available data channel capacity is at present reserved for future applications.

3.2.2. *Programme type code*

This code is used to identify the generic programme type (e.g. news, traffic announcement, etc.) and is intended to facilitate searching for a particular programme type, independent of the network which broadcasts it. Again this requires a high repetition rate.

3.2.3. *Programme item number code*

This is the scheduled start-time of a programme (specified as week-of-year, day-of-week, hour-of-day, minute-of-hour) and is used to identify uniquely (with an ambiguity of one year for a particular network) a particular programme on a given network. This gives the facility whereby the listener can pre-programme his receiver to respond to a particular programme (or several programmes) in the coming day or week. If several programmes are thus selected they may be on different networks (provided, of course, they are all broadcast at different times or on different days).

In one possible method of pre-programming,[6] a light pen attached to the receiver is stroked across a bar-code printed in a programme journal or newspaper. This method is thought to be preferable to a keyboard method of entering the data which would be tedious and could easily give rise to errors in the entries.

A repetition rate of about once per second is used for this code.

3.2.4. *Decoder control code*

At present, the main application forseen for this code is to provide identification of whether music or speech is being broadcast. Using this code, and by equipping receivers with two volume-controls, one active for speech and the other for music, the listener could set his own speech/music balance.[1] Other applications might be automatic identification of quadrophonic broadcasts if these become available in the future.

The repetition rate for this code is about once per second.

3.3 Information for Display

Messages for display are transmitted as 7-bit ASCII (ISO 646) characters. It is expected that a single line display of 32 to 40 characters would be used in most receivers having this facility, though multi-line displays may become a possibility (at low cost) in the future. Portable receivers might use a short (seven character) display for the Network Name, only.

3.3.1. *Network name*

The transmitted Network Name is a group of seven ASCII characters giving an abbreviated station or network name (e.g. BBC LON for BBC Radio London) for display on the receiver, thus affirming the result of manual tuning.

The repetition rate of the Network Name is about once per second in the present BBC experimental system; this is commensurate with the speed of manual tuning.

3.3.2. *Radiotext*

By choosing a suitable addressing structure, a number of transparent data channels have been reserved in the present BBC experimental system for the transmission of text for display. Typical applications would be the transmission of the title of the current programme or the name of an item of music.

3.4. Future Possibilities

Clearly as the radio-data system develops other applications (for both control and display) will emerge. The messages described above occupy only about one-third of the available capacity in the present experimental BBC system (the remainder is packed with dummy messages) and there is thus plenty of spare capacity available to service these new applications. A flexible message format (see Sect. 4.2) has been chosen to permit these new types of message to be added without any disruption or impairment to the transmission and reception of existing kinds of message.

One possibility which is being investigated is the transmission of messages relating to programmes on m.f. and l.f. networks, thus avoiding the need for a separate radio-data service for these broadcasts.

Another possibility is to transmit clock-time-codes (sent perhaps once per minute) to reset a free-running quartz clock in the receiver—thus providing a highly accurate but low-cost clock.

4 Design of a Standardized V.H.F. Radio-Data System

We shall now return to the problem of how these radio-data signals are to be conveyed to the receiver and in particular look at the progress which is being made towards producing a standardized radio-data system for use throughout Europe.

4.1. Choice of Subcarrier Frequency

As has been mentioned, at v.h.f. the preferred method for broadcasting radio-data signals uses a subcarrier added to the stereo multiplex signal at the modulation input to the v.h.f. f.m. transmitter.

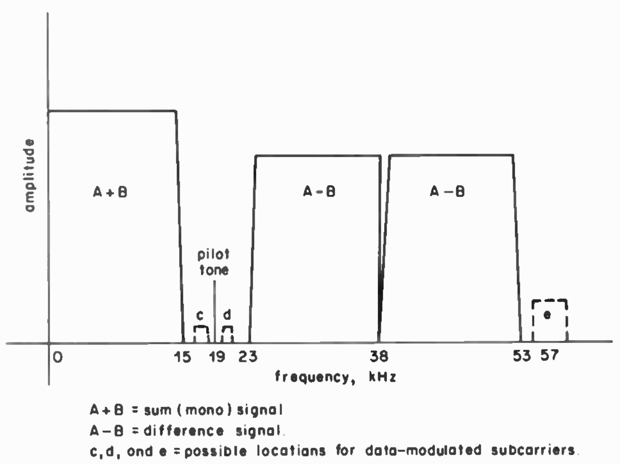

Fig. 2. Spectrum of a stereo multiplex signal showing possible locations for data-modulated sub-carriers.

As may be seen from Fig. 2 there are two regions in the spectrum of a conventional stereo multiplex signal where, in theory, a subcarrier could be located without causing audible impairment to normal programme reception. These are:

(1) In the region of the 19 kHz pilot-zone, between the upper frequencies of the sum signal ($\sim$15 kHz) and the lower frequencies of the difference signal ($\sim$23 kHz).

(2) In the region above the upper frequencies of the difference signal where, for various reasons, subcarrier frequencies around 57 kHz (three times the pilot-tone frequency) are generally preferred.

In practice, a small number of existing receivers are prone to impaired reception wherever the subcarrier is located. The severity of the impairment depends upon both the level of the subcarrier (a reduction of which reduces the reliable service area for radio-data reception) and, to a smaller extent, the transmitted bit-rate.

By 1980 four different radio-data systems (see Table 2) had been proposed and developed by different European Broadcasters.

Two of these systems (NOS (Holland) and YLE (Finland)) use subcarriers in the region of 19 kHz and two (BBC/Televerket (Sweden)† and TDF (France)) use subcarriers in the region of 57 kHz.

Clearly it is important that a standardized radio-data system should be adopted in all countries and one of the technical sub-groups of the European Broadcasting Union (EBU) has the task of attempting to standardize a v.h.f. radio-data system. This sub-group decided to conduct field-trials of all the proposed systems so that their performance could be compared under uniform conditions.

4.2. International Field-Trials of Radio-Data

The field-trials took place during September 1980 in the region of Bern in Switzerland. The source equipments for the four proposed radio-data systems described in Table 2 were installed by the Swiss PTT at one of their broadcast transmitters in the mountains and arrangements made for each system to be put on-air in turn.

It was agreed that these field-trials should test the data-channel provided by the different systems rather than complete systems including error protection (which could, of course, be applied to any system).

It was decided to concentrate mainly on mobile reception tests because it was already known that these provided the most stringent test of the ruggedness of radio-data systems. Test routes were selected which included sections of road with difficult reception conditions (due mainly to severe multipath propagation conditions because of the mountainous terrain).

The BBC provided one of the measuring vehicles (see the photograph in Fig. 3) which was equipped with a field-strength measuring receiver, and a specially developed microprocessor system for checking the received radio-data signals for errors. Decoders for the various systems were provided by their proponents. Facilities were also provided for recording the received sound-programme signals and the radio-data signal. This enabled very detailed computer analysis of the statistics of the errors (including correlations of error-

† There were, in fact, some minor differences between the BBC system and that proposed by Televerket. However, for simplicity they are treated as if they were identical here.

Table 2. Details of the radio-data systems tested in Switzerland

Broadcaster	Country	Subcarrier frequency	Peak deviation of main f.m. carrier due to subcarrier	Type of modulation of subcarrier	Intermediate coding of data	Message bit-rate (ignoring overheads)§	Remarks
BBC and Televerket	UK Sweden	57 kHz ($3 \times$ pilot-tone frequency)	± 2250 Hz	d.s.b.s.c.a.m.†	differential and biphase	1187·5 bit/s (1/48 sub-carrier frequency)	57 kHz family of systems
TDF	France	58·3314 kHz (not related to pilot-tone frequency)	± 1500 Hz	d.s.b.s.c.a.m.†	differential	607·7 (1/96 sub-carrier frequency)	
NOS	Holland	16·625 kHz (7/8 pilot-tone frequency)	± 250 Hz	d.s.b.s.c.a.m.†	none	593·75 (1/28 sub-carrier frequency)	19 kHz family of systems
YLE	Finland	19·0 kHz (pilot-tone is modulated by the data signal)	pilot-tone $= \pm 6{\cdot}75$ kHz data modulated sidebands ± 500 Hz	a.m.‡ of the pilot-tone with a modulation index of about 0·07	data frequency shift keys a 2 kHz sub-subcarrier with a peak deviation of ± 200 Hz	593·75 (1/32 pilot-tone frequency)	

†Double sideband suppressed carrier amplitude modulation
‡Amplitude modulation
§i.e. no allowance made for check bits, header codes etc. but excluding the redundancy due to intermediate coding.

rate with programme level and distortion) to be made at a later date.

It was found that under the conditions of the field-trials (which are thought to be representative of v.h.f. car-radio reception in hilly areas and towns where multipath interference is a serious problem), all four systems suffered from a high mean bit-error-rate (in the region of 10 to 20%) over all the test routes.

The principal source of errors in all the systems was thought to be multipath propagation causing programme signals to intrude into the data channels. These errors were found to occur in dense bursts and thus the proportion of radio-data blocks which were corrupted by errors was lower than would have been expected from the mean bit-error-rate had the errors occurred randomly.

Thus with the 57 kHz family of systems (BBC/Televerket and TDF) (see Table 2) about 70% of blocks of 100 bits were received error-free on each test route.

The same cannot be said, however, for the 19 kHz family of systems (NOS and YLE) which both gave very disappointing results. With the YLE system only 27% of the transmitted 100-bit blocks were received error-free; with the NOS system less than 10% of the transmitted 128-bit blocks were received error-free.

4.3. Progress since the International Field-Trials

The Switzerland field-trials represented a significant milestone in the development of radio-data systems and gave a greatly increased understanding of the strengths and weaknesses of the various systems.

It was concluded that the 57 kHz family of radio-data systems was to be preferred to the 19 kHz family although the final choice of system has yet to be made.

The TDF system is believed to be slightly more rugged under some circumstances than the BBC/Televerket system but this advantage is offset by the fact that its data rate is only half that of the BBC/Televerket system.

Fig. 3. BBC measuring vehicle during the international field trials of radio-data systems in Switzerland.

Because of the density and length of the error-bursts observed in these field-trials, it seems clear that no error-correcting code of practicable length could be used to advantage in mobile radio-data reception with any of the systems tested. A better approach would seem to be to avoid the high redundancy needed for error-correction and to use a highly reliable form of error detection. In this case, the data format would be designed so as to repeat the various items of data as often as possible, each according to the needs of the receiver (see Sect. 3 and Table 1).

A block length of about 100 bits seemed to be optimum for the BBC/Televerket system; with shorter blocks the percentage overhead for check-bits and header codes increased whilst with longer blocks the proportion received error-free decreased.

Using this experience gained in the Switzerland field-trials, the BBC have continued the development of their 57 kHz radio-data system and have now adopted a data format using 114-bit blocks. With a data-rate of 1187·5 bits there are an exact integral number (625) of 114-bit blocks in a minute which is convenient if minute-markers for clock-time transmissions are required. An important feature of this data format[7] is a modified cyclic-block-code[8] which reliably detects nearly all possible error patterns and provides block-synchronization (framing) information without the need for a separate header code for this purpose.[9]

In this proposed system, each transmitted block is a self-contained packet containing an address and each block can be decoded independently of the others. This addressing system gives flexibility to service a wide range of needs and to evolve gradually as new applications become apparent.

This system is currently (April 1981) experimentally on-air from three BBC v.h.f. radio-broadcast transmitters in the London area. Simple fixed messages of the kinds described in Table 1 and Section 3 of this document are broadcast at present. (For further details see Ref. 7.)

It remains, of course, for European broadcasters to agree a firm specification for a standardized radio data system, and this work continues within the EBU.

5 Conclusions

If a radio-data service were provided by broadcasters, the operation and tuning of all categories of receiver could be simplified. In particular, the facilities it would offer for automatic tuning and pre-programming would be significant steps forward in helping the listener to make best use of the available broadcasts.

The international field-trials of the various proposed radio-data systems were a significant milestone in the development of a standardized radio-data system and steady progress continues in that direction. These field-trials also showed that reliable radio-data reception was possible with low cost receiver/decoders even under very adverse reception conditions.

6 Acknowledgments

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UDC 621.396.674.3: 621.396.677
Indexing Terms: Antennas

Thin-wire antennas in the presence of a perfectly-conducting body of revolution

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and

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SUMMARY

A new numerical method for evaluating electromagnetic field, current distribution and input impedance of thin axial antennas in the presence of a perfectly conducting body of revolution is presented. Two particular cases, an axial dipole antenna near a conducting spheroid (prolate and oblate, including sphere) and near a solid truncated cone (including solid cylinder) are analysed. The actual body of revolution is replaced by an equivalent interior system of fictitious sources, axial arrays of Hertzian dipoles and magnetic ring currents. The moments of these sources are then adjusted to match boundary conditions on the surface of the body, while for the current distributions on the antenna conductors Hallén's type integral equations are derived. The system of integral equations obtained is solved numerically by the point matching method with the polynomial approximation for distributions of antenna currents. The solution is numerically optimized due to the number and locations of fictitious sources. Some numerical results for antenna input impedance and cimomotore force (c.m.f.) are presented.

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1 Introduction

It is well known that the properties of linear antennas can be changed by the presence of conducting bodies, so the analysis of such a problem is of great practical importance. In this paper thin-wire antennas in the presence of a conducting body of revolution are analysed.

Perfectly-conducting bodies of revolution in unbounded regions have been extensively treated in the past.[1,2] Various approaches have been used in analysis of linear antennas in the presence of conducting bodies. The conducting surface have been represented by surface patch modelling, high frequency techniques,[3] and by analytically[4] or numerically[5] known Green's functions. Also, the surface can be treated by the theory of characteristic modes for conducting bodies.[6,7] Especially, the problem of the linear antennas in the presence of a conducting sphere has been extensively treated.[8,9,10]

All the methods mentioned involve significant mathematical difficulties due to analytical and numerical evaluations and require high-speed computers with large memory. Having in mind the above-mentioned difficulties the authors suggest a new numerical method for analysis of linear antennas in the presence of a body of revolution. The method is very simple, accurate and can be applied to any arbitrary shaped body of revolution, even when a small computer is used. The accuracy of the method presented is the same as, or often better than, the accuracy of the mentioned methods.

2 Outline of the Method

Although the proposed method can be applied to an axial thin-wire antenna in the presence of an arbitrary shaped body of revolution, two particular cases are investigated: an axial dipole antenna near a perfectly-conducting spheroid (prolate and oblate, including sphere) representing a smooth body of revolution, and an axial dipole near a perfectly-conducting truncated cone (including solid cylinder) representing the bodies of revolution with edges. The radiating systems considered are shown in Fig. 1(a) and 1(b). An asymmetrical dipole having arms of lengths h_1, h_2 and radii a_1, a_2 is placed along the axis of the body (z-axis of r, ψ, z cylindrical coordinate system) and is excited by a scalar potential difference U of angular frequency ω at the point $z = d = b+g+h_2$.

According to the principle of equivalence, the conducting body of revolution is replaced by an equivalent interior system of fictitious sources, axial collinear arrays of Hertzian dipoles and magnetic current rings. Although the sources can be located arbitrarily inside the body volume, in order to simplify the analysis their locations are chosen in the following way:

Locations of Hertzian dipoles are

$$r_{hi} = 0, \quad z_{hi} = (2i-1)f_1 b/(2N_1-1), \quad i = 1, 2, \ldots, N_1, \quad (1)$$

$$r_{hi} = 0, \quad z_{hi} = (2N_1-2i+1)f_2 b/(2N_2-1), \quad i = N_1+1, N_1+2, \ldots, N_h = N_1+N_2. \quad (2)$$

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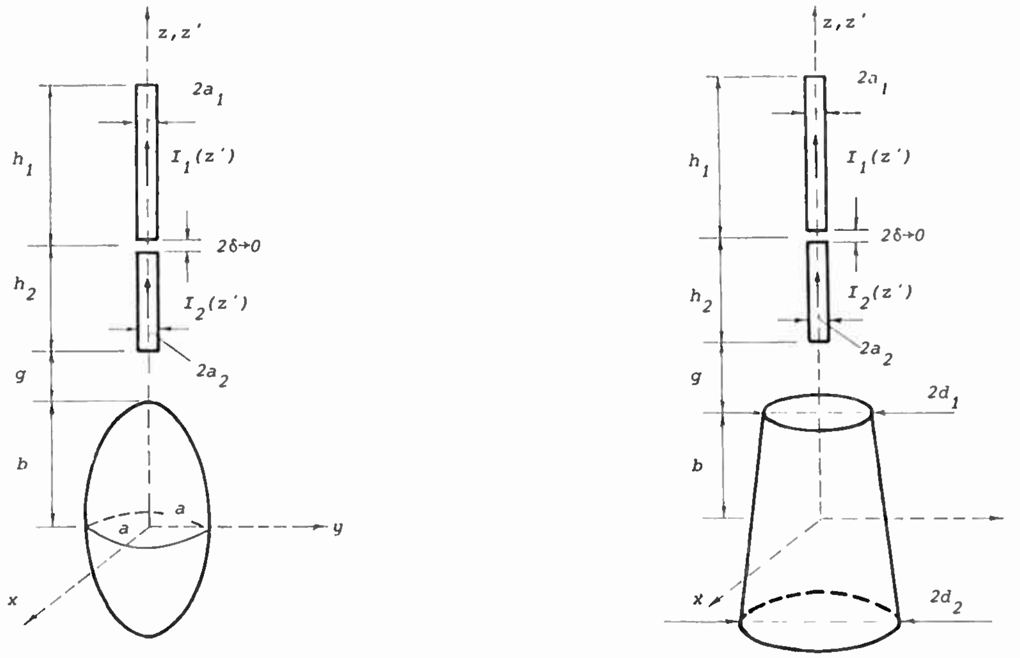

(a) Asymmetrical dipole axial to a conducting spheroid.

(b) Asymmetrical dipole axial to a conducting truncated cone.

Fig. 1.

Magnetic rings of radii r_{mn} are situated at $z = z_{mn}$ planes, where $n = 1, 2, \ldots, N_m$. For spheroid

$$r_{mn} = f_3 a \sin t_n, \quad z_{mn} = f_4 b \cos t_n, \quad t_n = n\pi/(N_m+1), \quad n = 1, 2, \ldots, N_m. \tag{3}$$

In the case of the truncated cone the additional magnetic ring currents are placed at the cone corners to account for local field singularities. So:

$$r_{m1} = d_1, \quad z_{m1} = b, \quad r_{m2} = d_2, \quad z_{m2} = -b, \tag{4}$$

$$r_{mn} = f_3[b(d_1+d_2) - z_{mn}(d_2-d_1)]/2b, \quad z_{mn} = f_4 b(N_3-2n+3)/(N_3-3), \quad n = 3, \ldots, N_3, \tag{5}$$

$$r_{mn} = f_5[2d_1+(1-f_6)(d_2-d_1)](n-N_3)/N_4, \quad z_{mn} = f_6 b, \quad n = N_3+1, \ldots, N_3+N_4/2, \tag{6}$$

$$r_{mn} = f_5[2d_1+(1+f_6)(d_2-d_1)](n-N_3-N_4/2)/N_4, \quad z_{mn} = -f_6 b, \tag{7}$$

$$n = N_3+1+N_4/2, \ldots, N_m = N_3+N_4,$$

where $0 < f_v < 1$, $v = 1, 2, \ldots, 6$, and f_v are dimensionless optimization factors.

Since h_1, $h_2 \ll a_1$, a_2 and a_1, $a_2 \ll \lambda$ (λ is free space wavelength) it is assumed that the current in the antenna is localized on the axes of the conductors, with the axial distribution $I_1 = I_1(z')$ along the upper, and $I_2 = I_2(z')$ along the lower dipole arm.

Then the magnetic and electric vector potentials of the system are:

$$\mathbf{A} = A\hat{z}, \quad A = \frac{\mu}{4\pi}\left[\int_d^{d+h_1} I_1 \frac{\exp(-jkR)}{R}\,dz' + \int_{d-h_2}^{d} I_2 \frac{\exp(-jkR)}{R}\,dz' + \sum_{i=1}^{N_h} kp_i \frac{\exp(-jR_i)}{R_i}\right], \tag{8}$$

where

$$R = \sqrt{r^2+(z-z')^2}, \quad R_i = k\sqrt{(r-r_{hi})^2+(z-z_{hi})^2}, \quad i = 1, \ldots, N_h,$$

$$\mathbf{F} = F\hat{\psi},$$

$$F = \frac{\varepsilon}{\pi}\sum_{n=1}^{N_m} I_{mn} k r_{mn} \int_0^{\pi/2} (2\sin^2\alpha - 1)\frac{\exp(-jR_{mn})}{R_{mn}}\,d\alpha, \tag{9}$$

where

$$R_{mn} = k\sqrt{(r+r_{mn})^2+(z-z_{mn})^2-4rr_{mn}\sin^2\alpha}, \quad n = 1, \ldots, N_m,$$

$\hat{z}$, $\hat{\psi}$ are sorts of cylindrical coordinate system r, ψ, z; $k = 2\pi/\lambda$ is the free-space propagation coefficient; μ is magnetic permeability; ε is dielectric permittivity; p_i is the electric dipole moment of the ith Hertz dipole and I_{mn} is magnetic current of nth magnetic source.

The unknown moments, p_i, $i = 1, \ldots, N_h$, and currents, I_{mn}, $n = 1, \ldots, N_m$, of the fictitious sources are adjusted to match boundary conditions on the surface, S, of the body of revolution:

$$\mathbf{E} \times \hat{n} = 0, \quad \text{on } S, \tag{10}$$

where $\mathbf{E}$ is the resultant electric field on considered radiating system (see Appendix) and $\hat{n}$ is the unit outward normal to the body surface.

Satisfying the boundary value conditions on the surface of the dipole arms, the following Hallén's type integral equations are derived:

$$A(r=a_1, d \leqslant z \leqslant d+h_1) = C_1 \cos k(z-d) + C_2 \sin k(z-d) + \frac{1}{jv}\int_{s=d}^{z} E_{mz}\Big|_{\substack{z=s\\ r=a_1}} \sin k(z-s)\,ds, \tag{11}$$

Table 1

Input impedance/admittance of half-wave dipole ($h_1 = h_2 = 0{\cdot}25\lambda$, $a_1 = a_2 = 0{\cdot}007022\lambda$, $M_1 = M_2 = 3$) axial to the prolate spheroid having semiaxes a and b for various values of distance g

g/λ	f_1	f_2	Y[mS]	Z[Ω]
a=0.1875λ, b=0.25λ,		N_1=N_2=6,		N_m=0
0.00*	0.80	0.75	7.186 - j2.558	123.5 + j44.0
0.05	0.75	0.75	8.905 - j3.036	100.6 + j34.3
0.10	0.75	0.75	9.191 - j3.346	96.1 + j35.0
0.25	0.75	0.75	9.176 - j3.548	94.8 + j36.7
0.50	0.75	0.75	9.153 - j3.559	94.9 + j36.9
0.75	0.75	0.75	9.154 - j3.547	95.0 + j36.8
1.00	0.75	0.75	9.154 - j3.554	94.9 + j36.8
a=0.125λ, b=0.25λ,		N_1=N_2=6,		N_m=0
0.00*	0.85	0.80	7.902 - j2.821	112.2 + j40.1
0.05	0.85	0.80	9.155 - j3.423	95.8 + j35.8
0.10	0.80	0.80	9.163 - j3.597	94.6 + j37.1
0.25	0.80	0.80	9.119 - j3.520	95.4 + j36.8
0.50	0.80	0.80	9.165 - j3.565	94.8 + j36.9
0.75	0.80	0.80	9.150 - j3.545	95.0 + j36.8
1.00	0.80	0.80	9.156 - j3.554	94.9 + j36.8

* When g/λ=0 there is no conducting junction between the antenna and the body

Table 2

Input impedance/admittance of half-wave dipole ($h_1 = h_2 = 0{\cdot}25\lambda$, $a_1 = a_2 = 0{\cdot}007022\lambda$, $M_1 = M_2 = 3$) radial to the conducting sphere of radius $a = 0{\cdot}25\lambda$ ($N_1 = N_2 = 5$, $N_m = 0$) for different g/λ

g/λ	f_1	f_2	Y[mS]	Z[Ω]
0.00*	0.80	0.60	6.851 - j2.287	131.3 + j43.8
0.05	0.75	0.60	8.771 - j2.674	104.3 + j31.8
0.10	0.50	0.50	9.270 - j3.099	97.0 + j32.4
0.25	0.50	0.50	9.242 - j3.600	94.0 + j36.6
0.50	0.50	0.50	9.135 - j3.547	95.1 + j36.9
0.75	0.50	0.50	9.161 - j3.551	94.9 + j36.8
1.00	0.50	0.50	9.151 - j3.552	95.0 + j36.8

Table 3

Input impedance/admittance of half-wave dipole ($h_1 = h_2 = 0{\cdot}25\lambda$, $a_1 = a_2 = 0{\cdot}007022\lambda$, $M_1 = M_2 = 3$) axial to the oblate spheroid having semiaxes a and b for various values of distance g

g/λ	f_1	f_2	Y[mS]	Z[Ω]
a=0.25λ,	b=0.1875λ,		N_1=N_2=2, N_m=8,	f_3=0.67, f_4=0.142
0.00*	0.50	0.50	6.781 - j2.483	130.0 + j47.6
0.05	0.50	0.50	8.696 - j2.819	104.0 + j33.7
0.10	0.50	0.50	9.187 - j3.175	97.2 + j33.6
0.25	0.50	0.50	9.207 - j3.572	94.4 + j36.6
0.50	0.50	0.50	9.150 - j3.551	95.0 + j36.9
0.75	0.50	0.50	9.155 - j3.551	94.9 + j36.8
1.00	0.50	0.50	9.155 - j3.552	94.9 + j36.8
a=0.25λ,	b=0.125λ,		N_1=N_2=1, N_m=10,	f_3=0.87, f_4=0.166
0.00*	0.50	0.50	6.494 - j2.559	133.3 + j52.5
0.05	0.50	0.50	8.585 - j2.805	105.2 + j34.4
0.10	0.50	0.50	9.150 - j3.132	97.8 + j33.5
0.25	0.50	0.50	9.217 - j3.578	94.3 + j36.6
0.50	0.50	0.50	9.147 - j3.546	95.0 + j36.8
0.75	0.50	0.50	9.156 - j3.553	94.9 + j36.8
1.00	0.50	0.50	9.154 - j3.550	95.0 + j36.8

Table 4

Input impedance/admittance of half-wave dipole ($h_1 = h_2 = 0{\cdot}25\lambda$, $a_1 = a_2 = 0{\cdot}007022\lambda$, $M_1 = M_2 = 3$) axial to the conducting solid cylinder ($d_1 = d_2 = b = 0{\cdot}25\lambda$, $N_1 = N_2 = 0$, $N_3 = N_4 = 10$) for different values of distance g

g/λ	f_5	f_6	Y[mS]	Z[Ω]
0.00*	0.80	0.90	2.326 - j4.846	80.5 + j167.7
0.05	0.80	0.80	4.987 - j5.196	96.1 + j100.2
0.10	0.80	0.80	6.388 - j4.450	105.4 + j 73.4
0.25	0.80	0.80	8.355 - j3.002	106.0 + j 38.1
0.50	0.80	0.80	9.409 - j3.571	92.9 + j 35.2
0.75	0.80	0.80	9.069 - j3.566	95.5 + j 37.5
1.00	0.80	0.80	9.191 - j3.539	94.8 + j 36.5

Table 5

Input impedance/admittance of symmetrical dipole ($h_1 = h_2 = h$, $a_1 = a_2 = 0{\cdot}007022\lambda$, $M_1 = M_2 = M$) radial to the conducting sphere having radius $a = 0{\cdot}25\lambda$ ($N_1 = N_2 = 5$, $N_m = 0$) for different values of h/λ and g/λ

h/λ	M	g/λ	f_1	f_2	Y[mS]	Z[Ω]
0.100	2	0.00*	0.80	0.60	0.239+j3.914	15.5-j254.3
0.100	2	0.25	0.50	0.50	0.090+j3.632	6.8-j275.2
0.100	2	1.00	0.50	0.50	0.090+j3.635	6.8-j274.9
0.100	2	∞	-	-	0.090+j3.635	6.8-j274.9
0.250	3	0.00*	0.80	0.60	6.851-j2.287	131.3+j 43.8
0.250	3	0.25	0.50	0.50	9.242-j3.600	94.0+j 36.6
0.250	3	1.00	0.50	0.50	9.151-j3.552	95.0+j 36.8
0.250	3	∞	-	-	9.154-j3.551	94.9+j 36.8
0.375	3	0.00*	0.80	0.60	1.626-j0.017	614.8+j 6.6
0.375	3	0.25	0.50	0.50	1.531-j0.267	633.8+j110.5
0.375	3	1.00	0.50	0.50	1.535-j0.260	633.3+j107.4
0.375	3	∞	-	-	1.535-j0.260	633.3+j107.4
0.500	3	0.00*	0.80	0.60	1.038+j1.686	264.8-j430.1
0.500	3	0.25	0.50	0.50	0.961+j1.577	281.7-j461.2
0.500	3	1.00	0.50	0.50	0.963+j1.582	280.9-j461.2
0.500	3	∞	-	-	0.963+j1.581	280.9-j461.2
0.750	4	0.00*	0.80	0.60	5.339-j1.412	175.0+j 46.3
0.750	4	0.25	0.50	0.50	6.832-j2.117	133.6+j 41.4
0.750	4	1.00	0.50	0.50	6.732-j2.097	135.4+j 42.2
0.750	4	∞	-	-	6.747-j2.095	135.2+j 42.0

Table 6

Input impedance/admittance of half-wave dipole ($h_1 = h_2 = 0{\cdot}25\lambda$, $a_1 = a_2 = 0{\cdot}007022\lambda$, $M_1 = M_2 = 3$) radial to the conducting sphere having different radii a ($N_1 = N_2$, $N_m = 0$) for various distance g

a/λ	g/λ	N_1	f_1	f_2	Y[mS]	Z[Ω]
0.50	0.00*	5	0.80	0.60	6.985-j2.345	128.7+j43.2
0.50	0.05	5	0.75	0.60	8.843-j2.534	104.5+j30.0
0.50	0.10	5	0.50	0.50	9.448-j3.051	95.8+j31.0
0.50	0.25	5	0.50	0.50	9.212-j3.697	93.5+j37.5
0.50	0.50	5	0.50	0.50	9.147-j3.506	95.3+j36.5
0.50	0.75	5	0.50	0.50	9.157-j3.572	94.8+j37.0
0.50	1.00	5	0.50	0.50	9.153-j3.541	95.0+j36.8
0.75	0.00*	6	0.50	0.50	7.911-j1.856	119.8+j28.1
0.75	0.05	6	0.75	0.50	8.817-j2.400	105.6+j28.7
0.75	0.10	6	0.60	0.50	9.495-j2.826	96.8+j28.8
0.75	0.25	6	0.50	0.50	9.312-j3.728	92.5+j37.0
0.75	0.50	6	0.50	0.50	9.114-j3.510	95.6+j36.8
0.75	0.75	6	0.50	0.50	9.169-j3.564	94.7+j36.8
0.75	1.00	6	0.50	0.50	9.149-j3.546	95.0+j36.8
1.00	0.00*	6	0.80	0.60	5.580+j0.326	178.6-j10.4
1.00	0.05	6	0.50	0.50	9.155-j2.182	103.4+j24.6
1.00	0.10	6	0.50	0.50	9.601-j2.771	96.1+j27.8
1.00	0.25	6	0.50	0.50	9.317-j3.474	92.4+j37.2
1.00	0.50	6	0.50	0.50	9.121-j3.489	95.6+j36.6
1.00	0.75	6	0.50	0.50	9.167-j3.582	94.6+j37.0
1.00	1.00	6	0.50	0.50	9.147-j3.535	95.1+j36.8

Table 7

Input impedance/admittance of asymmetrical dipole (h_1, h_2, $a_1 = a_2 = 0{\cdot}007022\lambda$, $M_1 = M_2 = 3$) radial to the conducting sphere of radius $a = 0{\cdot}25\lambda$ ($N_1 = N_2 = 5$, $N_m = 0$) for various distance g

h_1/λ	h_2/λ	g/λ	f_1	f_2	Y[mS]	Z[Ω]
0.50	0.25	0.00*	0.80	0.60	1.512+j2.131	221.5+j312.2
0.50	0.25	0.25	0.50	0.50	1.457+j2.168	213.6-j317.8
0.50	0.25	1.00	0.50	0.50	1.456+j2.169	213.4-j317.8
0.25	0.50	0.00*	0.85	0.60	2.008+j2.739	174.1-j237.5
0.25	0.50	0.25	0.50	0.50	1.442+j2.146	215.7-j321.0
0.25	0.50	1.00	0.50	0.50	1.456+j2.170	213.3-j317.7

$$A(r=a_2, d-h_2 \leqslant z \leqslant d) = C_3 \cos k(z-d) + C_4 \sin k(z-d) + \frac{1}{jv} \int_{s=d}^{z} E_{mz} \bigg|_{\substack{z=s \\ r=a_2}} \sin k(z-s)\, ds, \quad (12)$$

where E_{mz} is the z-component of electric field due to the magnetic ring sources and C_1, C_2, C_3, C_4 are unknown constants to be determined.

Solution of integral equations (10–12) can be obtained only numerically, for example by the point-matching technique with the following polynomial approximation for the current distribution functions along the antenna conductors

$$I_1(z') = \sum_{m=1}^{M_1} B_m \left(\frac{d+h_1-z'}{h_1}\right)^m, \quad I_2(z') = \sum_{m=1}^{M_2} D_m \left(\frac{z'-d+h_2}{h_2}\right)^m. \quad (13)$$

Matching points are taken equidistantly on the surfaces of the dipole arms and body of revolution.

Knowledge of current distributions and moments of fictitious sources enables us to compute other quantities of interest, such as input impedance, currents on the body of revolution and radiation patterns (cimomotore force—c.m.f., or field strength).

Since too large a number of fictitious sources becomes costly, the solution is numerically optimized due to the number and the locations of fictitious sources. (The most appropriate values of optimization factors, f_v, are determined.) For that purpose the mean error in satisfying the boundary conditions

$$\delta_s = \frac{1}{S} \int_S \delta \, dS \quad (14)$$

is minimized over the conducting body surface, S, where $\delta = |E_t/E_n|$ is the ratio of tangential, E_t, and normal, E_n, electric field component on the surface S.

3 Numerical Results

On the basis of the presented approach the general computer program for numerical solution of the axial antenna problem in the presence of a conducting body of revolution is formed. The numerous evaluations are performed and the convergence of the results and the satisfaction of the boundary conditions on the antenna and the body surfaces are investigated, so that the following conclusions can be drawn:

- Increase in number of fictitious sources or degree of the current distribution polynomials will directly influence satisfaction of boundary conditions on the surfaces of the antenna and the body of revolution.
- Investigation of numerical results has shown that the second- to fourth-order polynomials (approximating the current distributions along the antenna conductors which are less than half of wavelength) yield satisfactory results both for the antenna input impedance and radiation pattern. Also, the ratio of the intensity of tangential and normal component of electrical field at the antenna conductor surfaces is of the order of magnitude of 10^{-3} to 10^{-4} everywhere except in the small areas near antenna ends and the feeding zone, as is the case in an insulated dipole.[11]
- Concerning the body of revolution, it is noticed that with relatively small number of fictitious sources (about twenty), the mean error, δ_s, in satisfying the boundary conditions on the body surface can be made less than 1% by appropriate choice of number and locations of fictitious sources.

In Tables 1, 2 and 3 the numerical results for the imput impedance/admittance of an axial half-wave dipole near a perfectly-conducting prolate spheroid, sphere and oblate spheroid, respectively, when the dimensions of the spheroid and the distance between the spheroid and the dipole are varied, are presented.

In Table 4 the corresponding results for the axial dipole near a perfectly-conducting solid cylinder are presented.

In Table 5 the values for input impedance/admittance of an axial symmetrical dipole near a perfectly-conducting sphere are presented. The dimensions of the sphere are kept constant and the dipole arm length and distance between the sphere and the dipole are varied.

In Table 6 the values for input impedance/admittance of axial half-wave dipole near perfectly conducting sphere with different radii and for various distances between the sphere and the dipole, are presented.

In Table 7 the corresponding results for axial asymmetrical dipole are presented.

In Fig. 2 the results for the c.m.f. of half-wave dipole 0·1 λ distant from the conducting spheroid (oblate and prolate) and solid cylinder are presented.

In Fig. 3 the numerical results for c.m.f. of half-wave dipole near solid cylinder, when the distance between the dipole and the cylinder is varied, are presented.

The numerical results presented point out that the influence of the body on the antenna input impedance is almost insignificant when the shortest distance between

Table 8

Comparison between authors' results and results presented in Ref. 10 for input admittance of half-wave dipole ($h_1 = h_2 = 0{\cdot}25\lambda$, $a_1 = a_2 = 0{\cdot}007022\lambda$, $M_1 = M_2 = M$) radial to the conducting sphere having different radii a ($N_1 = N_2 = 6$, $N_m = 0$) for various distance g

M	a/λ	g/λ	Y [ms] authors' results	Y [ms] results in Ref. 10
2	0.25	0.15	9.351 -j3.402	9.35 -j3.40
3	0.25	0.15	9.353 -j3.386	9.35 -j3.39
2	0.25	0.50	9.129 -j3.564	9.13 -j3.56
3	0.25	0.50	9.135 -j3.547	9.14 -j3.55
2	0.50	0.15	9.484 -j3.472	9.48 -j3.47
3	0.50	0.15	9.485 -j3.456	9.48 -j3.46
2	0.50	0.50	9.143 -j3.522	9.14 -j3.52
3	0.50	0.50	9.147 -j3.506	9.15 -j3.51
2	1.00	0.15	9.671 -j3.312	9.63 -j3.32
3	1.00	0.15	9.673 -j3.300	9.63 -j3.30
2	1.00	0.50	9.112 -j3.496	9.11 -j3.50
3	1.00	0.50	9.121 -j3.489	9.12 -j3.49

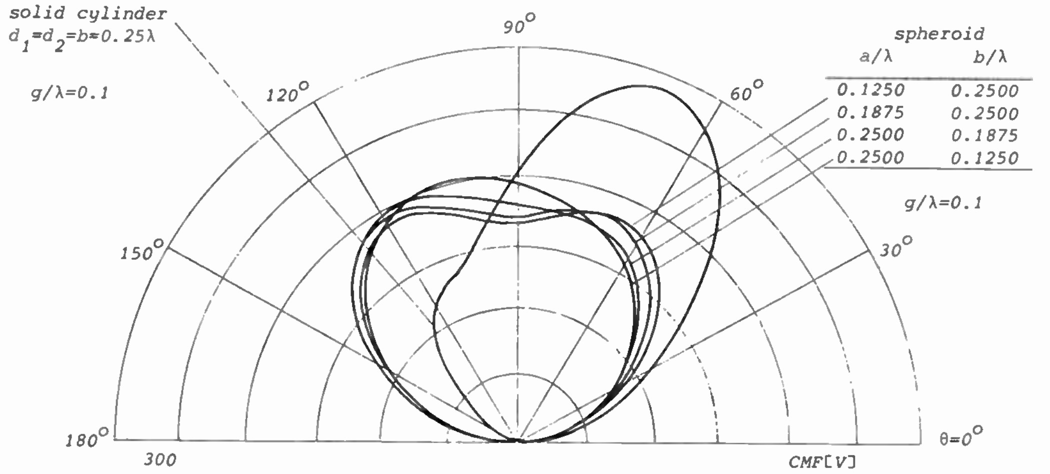

Fig. 2. C.m.f. of half-wave dipole ($h_1 = h_2 = 0{\cdot}25\lambda$, $a_1 = a_2 = 0{\cdot}007022\lambda$) $0{\cdot}1\lambda$ distant from conducting spheroid (having semiaxes a and b) or solid cylinder ($d_1 = d_2 = b = 0{\cdot}25\lambda$) versus angle θ between the direction of radiation and z-axis.

the antenna and the conducting body is greater than 0·1 wavelength and when antenna and body dimensions are at the same time of the order of a half wavelength. For shorter distances this influence is more significant. At the same time the influence of the conducting body on the antenna c.m.f. is considerable.

The results for input impedance of radial dipole near perfectly conducting sphere are in excellent agreement with the values presented in Ref. 10, as can be noticed from Table 8, and the same is seen for the c.m.f. The method described in Ref. 10 is based on Green's function for the Hertzian dipole above the sphere and on integral equations of Hallén type, which are numerically solved by the point-matching method combined with the polynomial approximation of the current distribution on the antenna conductors.

4 Experimental Results

Since the constructed solution for the problem of wire antenna axially oriented near a perfectly-conducting body of revolution is approximate, owing to the various assumptions previously made, the validity of the theory was checked experimentally.

For impedance measurements a modified mathematical model was used with a dipole antenna between two symmetrically placed spheres. In fact the image theory was used and the monopole antenna (coaxially fed) was mounted on a 5 λ square ground plane which was in turn vertically erected. Axially to the antenna a sphere was placed so that the distance between the sphere and the antenna, g, could be varied. The monopole was a simple protrusion of the inner coaxial-line conductor of length $h = 100$ mm and radius $a_d = 3$ mm. Excitation was performed by means of 50 Ω coaxial line of outer conductor $c = 6{\cdot}9$ mm. The radius of the sphere was $a = 102{\cdot}5$ mm (Fig. 4). Measurements were performed using a slotted coaxial line at 750 MHz and the v.s.w.r. method was used.

In Fig. 4 the theoretical and experimental results for the input resistance and reactance of a dipole antenna versus distance g between the antenna and the sphere are presented. Numerical results were obtained using the

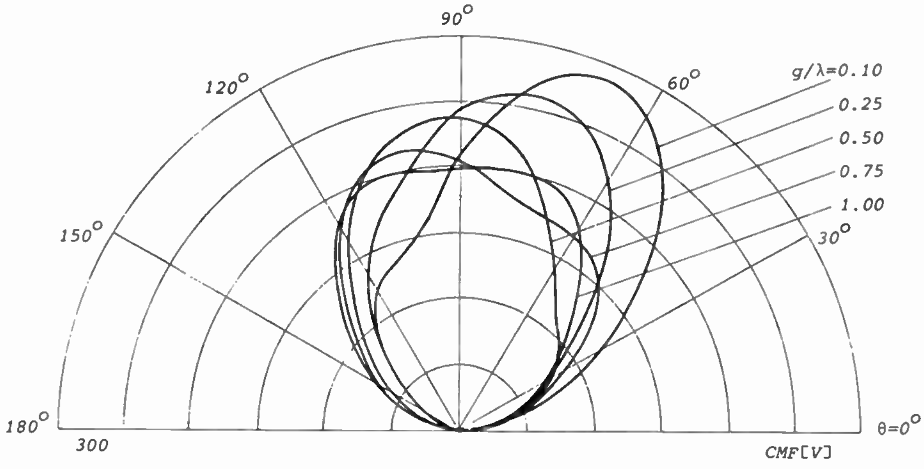

Fig. 3. C.m.f. of half-wave dipole ($h_1 = h_2 = 0{\cdot}25\lambda$, $a_1 = a_2 = 0{\cdot}007022\lambda$) axial to conducting solid cylinder ($d_1 = d_2 = b = 0{\cdot}25\lambda$) versus angle θ between the direction of radiation and z-axis for g/λ as a parameter.

third-order polynomial approximation for antenna current distribution, while the sphere was replaced by an axial array of sixteen Hertzian dipoles ($N_1, N_2 = 8$) for corresponding optimization factors. Also, the effect of antenna-flat ends was taken into account, which became significant when the distance between the sphere and the antenna was small (about a few antenna radii). Decreasing of the distance g caused increase of the input resistance and reactance, as well as the discrepancies between theoretical end experimental results and the error is about 10% when g is equal 4 mm. Also, there was an almost constant difference between measured and calculated values of antenna reactance. This error is mostly inherent in the mathematical model adopted for the feeding zone (scalar potential difference). The same behaviour of the antenna input impedance was observed at other frequencies about 100 MHz around the frequency of 750 MHz. As is seen, after comparison with experimental results, the agreement between the theory and measurement is quite good, thus substantiating the validity of the theory.

5 Conclusion

A simple method is presented for numerical analysis of wire antennas in the presence of a perfectly-conducting body of revolution having arbitrary shape (with smooth surface or with edges). Numerical results obtained according to the proposed theory are in very good agreement with experimental ones. Owing to the satisfactory convergence relating the number of fictitious sources replacing the body of revolution, the proposed method is convenient for small computers.

Also the method can be applied to the axial antenna with lumped impedance loadings, collinear array of dipoles and collinear array of impedance-loaded dipoles in the presence of a single conducting body of revolution or array of axial conducting bodies of revolution. The same approach can be used when the problems of scattering by conducting bodies of revolution are treated. So the method can be used for solving problems of mobile and satellite communications.

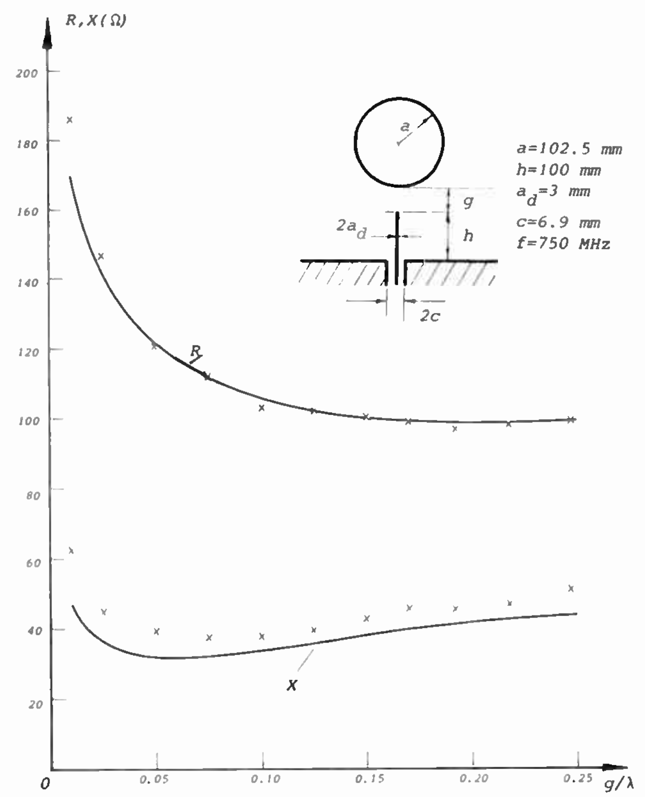

Fig. 4. Resistance (R) and reactance (X) of the dipole antenna between two spheres versus distance g
——— theory × × × × experiment

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8 Appendix

Resultant cylindrical components of electromagnetic field for considered system are:

$$E_r = E_{er} + E_{mr}, \quad E_\psi = 0, \quad E_z = E_{ez} + E_{mz}, \quad (15)$$

$$H_r = 0, \quad H_\psi = H_{e\psi} + H_{m\psi}, \quad H_z = 0, \quad (16)$$

where E_{er}, E_{ez}, $H_{e\psi}$ are electromagnetic field components due to electric currents and sources and E_{mr}, E_{mz}, $H_{m\psi}$ are electromagnetic field components due to magnetic sources:

$$E_{er} = -\frac{1}{4\pi k}\sqrt{\frac{\mu}{\varepsilon}}\left\{\int_{d}^{d+h_1}(I_1''' + k^2 I_1')\frac{\exp(-jkR)}{kr}\,dz' + \right.$$
$$+\int_{d-h_2}^{d}(I_2''' + k^2 I_2')\frac{\exp(-jkR)}{kr}\,dz' -$$
$$-[R^3 I_1'' - jkR^2(z-z')I_1' -$$
$$-jkr^2(1+jkR)I_1]\frac{\exp(-jkR)}{krR^3}\Bigg|_{z'=d}^{d+h_1} -$$
$$-[R^3 I_2'' - jkR^2(z-z')I_2' -$$
$$-jkr^2(1+jkR)I_2]\frac{\exp(-jkR)}{krR^3}\Bigg|_{z'=d-h_2}^{d} +$$
$$\left. + \sum_{i=1}^{N_h} k^5 r p_i (z - z_{hi})(j3 - 3R_i - jR_i^2)\frac{\exp(-jR_i)}{R_i^5}\right\}, \quad (17)$$

$$E_{ez} = -\frac{j}{4\pi k}\sqrt{\frac{\mu}{\varepsilon}}\left\{\int_{d}^{d+h_1}(I_1'' + k^2 I_1)\frac{\exp(-jkR)}{R}\,dz' + \right.$$
$$+\int_{d-h_2}^{d}(I_2'' + k^2 I_2)\frac{\exp(-jkR)}{R}\,dz' -$$
$$-[R^2 I_1' - (z-z')(1+jkR)I_1]\frac{\exp(-jkR)}{R^3}\Bigg|_{z'=d}^{d+h_1} -$$
$$-[R^2 I_2' - (z-z')(1+jkR)I_2]\frac{\exp(-jkR)}{R^3}\Bigg|_{z'=d-h_2}^{d} +$$
$$+\sum_{i=1}^{N_h} k^3 p_i[k^2(z-z_{hi})^2(3 + j3R_i - R_i^2) -$$
$$\left. -(R_i^2 + jR_i^3 - R_i^4)]\frac{\exp(-jR_i)}{R_i^5}\right\}, \quad (18)$$

$$H_{e\psi} = \frac{j}{4\pi}\left\{\int_{d}^{d+h_1}(I_1'' + k^2 I_1)\frac{\exp(-jkR)}{kr}\,dz' - \right.$$
$$-[RI_1' - jk(z-z')I_1]\frac{\exp(-jkR)}{krR}\Bigg|_{z'=d}^{d+h_1} +$$
$$+\int_{d-h_2}^{d}(I_2'' + k^2 I_2)\frac{\exp(-jkR)}{kr}\,dz' -$$
$$-[RI_2' - jk(z-z')I_2]\frac{\exp(-jkR)}{krR}\Bigg|_{z'=d-h_2}^{d} +$$
$$\left. + \sum_{i=1}^{N_h} k^3 p_i r (R_i - j)\frac{\exp(-jR_i)}{R_i^3}\right\}, \quad (19)$$

$$E_{mr} = \frac{k^3}{\pi}\sum_{n=1}^{N_m} I_{mn} r_{mn}(z - z_{mn})\int_0^{\pi/2}(1 - 2\sin^2\alpha)\times$$
$$\times(1 + jR_{mn})\frac{\exp(-jR_{mn})}{R_{mn}^3}\,d\alpha, \quad (20)$$

$$E_{mz} = -\frac{k^3}{\pi}\sum_{n=1}^{N_m} I_{mn} r_{mn}\int_0^{\pi/2}(r + r_{mn} - 2r\sin^2\alpha)\times$$
$$\times(1 + jR_{mn})\frac{\exp(-jR_{mn})}{R_{mn}^3}\,d\alpha, \quad (21)$$

$$H_{m\psi} = \frac{k^2}{2\pi}\sqrt{\frac{\varepsilon}{\mu}}\sum_{n=1}^{N_m} I_{mn} r_{mn}\times$$
$$\times\int_0^{\pi/2}\{(1 - 2\sin^2\alpha)(jR_{mn}^2 - R_{mn}^3 + jR_{mn}^4)\times$$
$$-2k^2(j3 - 3R_{mn} - jR_{mn}^2)[(r + r_{mn})^2 + (z - z_{mn})^2 -$$
$$-\sin^2\alpha[r^2 + r_{mn}^2 + (z - z_{mn})^2]]\}\frac{\exp(-jR_{mn})}{R_{mn}^2}\,d\alpha. \quad (22)$$

During the evaluation the following integrals appear:

$$\int_0^h z'^m \exp(-jkR)/R^n\,dz', \qquad R = \sqrt{r^2 + (z - z')^2}, \quad (23)$$

$$J_l = \int_0^{\pi/2}(A + B\sin^2\alpha)\frac{\exp(-jR_q)}{R_q^l}\,d\alpha,$$
$$R_q = R_1\sqrt{1 - F^2\sin^2\alpha},$$
$$R_1 = k\sqrt{(r + r_{mn})^2 + (z - z_{mn})^2}, \qquad F = 2krr_{mn}/R_1, \quad (24)$$

which are numerically calculated using the singularity extraction method. The procedure for calculating integral (23) is given in Ref. 12, while for (24) the following formulae can be used:

$$J_l = \sum_{m=0}^{l}\frac{(-j)^m}{m!\,R_1^{l-m}}(AQ_{l-m} + BS_{l-m}) + P_l, \quad (25)$$

$$S_m = \int_0^{\pi/2}\frac{\sin^2\alpha}{(1 - F^2\sin^2\alpha)^{m/2}}\,d\alpha = \frac{1}{F^2}(Q_m - Q_{m-2}), \quad (26)$$

$$Q_m = \int_0^{\pi/2}\frac{d\alpha}{(1 - F^2\sin^2\alpha)^{m/2}}$$
$$= \frac{1}{(1 - F^2)(m-2)}[(m-3)(2 - F^2)Q_{m-2} - (m-4)Q_{m-4}],$$
$$m = 3, 4, \ldots, \quad (27)$$

$$Q_{-2} = \frac{\pi}{2}(2 - F^2), \qquad Q_{-1} = E(F), \qquad Q_0 = \frac{\pi}{2},$$
$$Q_1 = K(F), \qquad Q_2 = \frac{\pi}{2\sqrt{1 - F^2}}, \quad (28)$$

$$P_l = \int_0^{\pi/2}\left[\exp(-jR_q) - \sum_{m=0}^{l}\frac{(-j)^m}{m!}R_q^m\right]\times$$
$$\times(A + B\sin^2\alpha)/R_q^l\,d\alpha, \quad (29)$$

where $K(F)$ and $E(F)$ are complete elliptic integrals of first and second kind.

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UDC 621.372.54:621.374:681.325

Indexing Terms: Electric filters, digital.

A single-chip non-recursive digital processor

G. W. KERR, B.A.*

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

SUMMARY

A single-chip non-recursive digital processor is being made using a 5V n-m.o.s. custom-l.s.i. approach. The processor will perform all the arithmetic required for a 48-tap non-recursive filter, with up to 30 000 samples/s. The coefficient resolution is up to 10 bits. The processor uses a special coefficient coding to minimize the time required to perform the multiplications, and for this reason the maximum sampling rate depends on the coefficients and the coefficient resolution required.

The design of the l.s.i. processor is given, and examples of filters which can be implemented using one l.s.i. processor, one 256×8 r.o.m. and two clock sources, are described.

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1 Introduction

Real-time non-recursive digital filtering usually involves a compromise between resolution required (on signal samples and coefficients), number of coefficients used, throughput (number of samples processed per second) and hardware (and hence circuit space, power dissipation and cost). For very low throughputs, a general-purpose microprocessor implementation may provide the user with a quick and reasonably flexible solution; for high throughput rates, dedicated array processors or high-speed parallel multipliers and accumulators are required, with inevitable cost penalties.

This paper describes a non-recursive digital processor which has input and output resolutions of 10 bits and processes 48 coefficients. The resolution of the coefficients is 8–10 bits but the sampling rate of the filter depends on the coefficients used, with a maximum value of 30 000 samples per second.

The rough specification initially required for the custom l.s.i. processor was:

Maximum processing clock speed	2 MHz
Minimum sampling rate for a certain set of coefficients	10 kHz
Number of coefficients	48
Signal resolution	10 bits
Minimum coefficient resolution	8 bits

2 Serial/parallel Multiplication

The most time-consuming activity in a digital processor is usually multiplication, especially when good resolution is required. This can be overcome when time is critical by using a parallel multiplier, but the penalty is considerable in terms of number of gates required. The use of a serial/parallel multiplier is far neater for many applications, but is time consuming, requiring at least the same number of clock periods as bits in the coefficient.

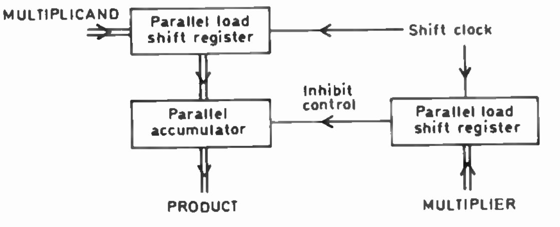

Fig. 1. Block diagram of a serial/parallel multiplier.

Figure 1 shows a simple block diagram of a serial/parallel multiplier. The multiplicand, presented in parallel, is shifted up in significance every clock period, and the multiplier, presented in serial, least significant bit first, is shifted at the same time. The shifted multiplicand is added into the accumulator whenever the multiplier bit presented is a '1', and is not added in when the bit is a '0'. In some serial/parallel multipliers, the accumulator is replaced by other logic, notably that derived from Booth's algorithm.[1] The product is shifted out from the accumulator in serial form in many implementations, but can also be read out in parallel.

From the specification given in Section 1, 48 10×8 bit

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products are required every 100 μs. This requires that, on average, one such product should be formed in every 4 clock periods, which is clearly impossible.

3 Canonic Signed Digit Coding

The conventional serial/parallel approach wastes clock periods on many multiplications, for example:

0000 0001 × multiplicand (in binary)

After the first clock period, nothing useful is done; the answer is available in the accumulator and the last seven (at least) clock periods are wasted. If the leading zeros or ones (in two's complement notation) could be avoided, then much time could be saved with small valued multipliers.

0100 0000 × multiplicand

Here only the 7th clock period is really useful, the first six merely shifting the multiplicand to the correct significance relative to the accumulator.

0111 1111 × multiplicand

At first glance all clock periods are useful, as shifting and adding takes place on each. But if we could subtract from the accumulator as well as add to it, we could reduce the multiplier (0111 1111) to (10000000–00000001), with two useful clock periods.

Canonic Signed Digit Code (CSDC) was proposed by A. Peled.[2] It is a non-unique 'signed-bit' code which minimizes the number of non-zero 'bits' in a multiplier. For example, 0111 1111 in binary becomes 1000000-1 in CSDC and 01000111 becomes 0100100-1 in CSDC. CSDC coding reduces the number of non-zero bits in an 8-bit binary word to an absolute maximum of 4 and an average of 2·8 non-zero signed bits.

4 Filter Implementation

The l.s.i. processor implementation reduces the number of clock periods required per product to the number of non-zero bits in the CSDC coefficient. Peled[2] suggested a hardware implementation of a recursive filter involving a number of r.o.m.s, r.a.m.s and a.l.u.s. The work described here has the objective of incorporating all the arithmetic processing into the custom-l.s.i. device, with the coefficients stored in an external r.o.m., to make the device general purpose. Some modifications to Peled's implementation have been made to take advantage of the restriction to non-recursive processing and the use of n-m.o.s. technology.

4.1 Signal Storage

A recirculating shift register, 48 words by 10 bits, is used, rather than the r.a.m. used by Peled.

4.2 Shifting the Signal Significance

Only non-zero bits of the coefficient are processed, so the signal is shifted to the appropriate significance by a ripple-through static shift matrix. Once all non-zero bits of one coefficient have been processed, the next product is started. A purpose-built transfer-gate array is used instead of the array of data selectors used by Peled.

4.3 Negative Bits

Instead of the shifted multiplicand being subtracted from the accumulator as in Peled's implementation, the multiplicand is pre-multiplied by −1 and the shifted multiplicand then added into the accumulator. This approach avoids the need for a large adding and subtracting accumulator; multiplying the 10-bit signal sample by −1 in two's complement is quite straightforward in m.o.s.

4.4 Accumulator

In a conventional serial/parallel implementation, complete individual products formed in the multiplier are summed into the filter accumulator. As there is no need to form the complete individual products, the intermediate accumulator in the multiplier is removed, and the subproducts summed in the one main accumulator.

4.5 Storage of Coefficients

As the device is intended to be general-purpose, the coefficients are stored in an external memory, most conveniently a r.o.m. or p.r.o.m. The coefficients have to be presented in some form of CSDC format to the processor, as outlined above. A simple desk-top computer program has been written to convert any set of coefficients to the form required in the 256 × 8 r.o.m. for this processor. All the clocks are simply derived from the r.o.m. contents. Each 8-bit word of the r.o.m. represents one processing clock period and usually one non-zero bit of a coefficient. The contents of the r.o.m. are stepped

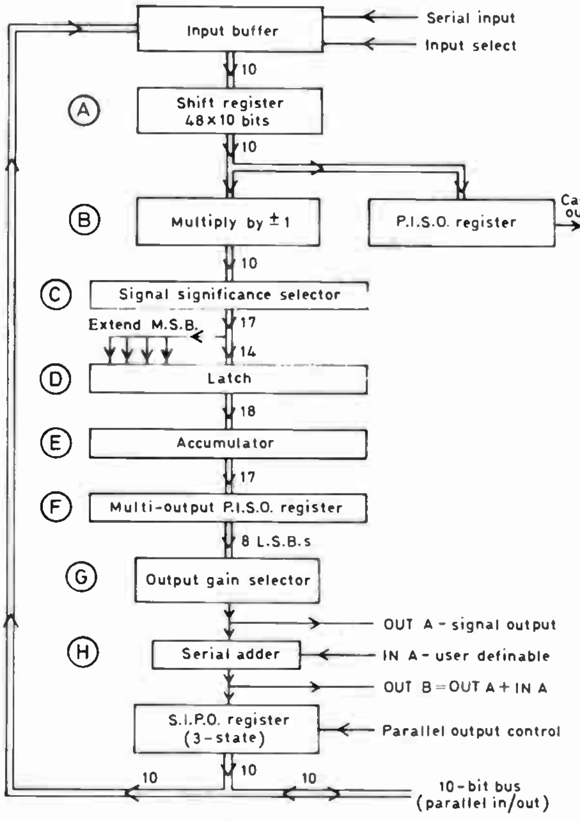

Fig. 2. Block diagram of an l.s.i. processor.

through via an on-chip 8-bit counter until an end of processing (EOP) indication is found.

5 System Description

5.1 Basic Processor

As mentioned above, the external r.o.m. controls directly the presentation of a new signal sample, the multiplication by ± 1, the shifting of signal significance and the subproduct accumulation. With reference to Fig. 2, a new signal sample is read into the signal shift register (A) at the start of processing and the oldest sample rejected.

The next oldest sample is then at the shift register output ready for multiplication by the appropriate coefficient. This requires a number of clock periods equal to the number of non-zero bits in the (non-zero) coefficient. Within one clock period the signal is multiplied by ± 1 (B), then moved in significance 0–7 bits (C) and stored in a latch (D). On the next clock period the signal stored in the latch is added into the accumulator (E) in parallel with the next set of operations as described above.

When a new signal sample is required (i.e. when multiplication by one coefficient is complete), the signal shift register is moved on prior to the operations described. At the end of processing, two extra clock periods are required to form the complete output sample, load it into the output buffering stages (F), (G), and clear the accumulator ready for the next processing cycle.

5.2 Input and Output Arrangements

Serial input and serial output are available on the device, optionally using clocks generated by the device. The internal serial input clock occurs at the start of processing, and the internal serial output clock occurs at the end of processing. Also available on the device are the oldest signal sample in the signal shift register (the 'cascade' output shown in Fig. 2) and a serial adder (H in Fig. 2). One of the inputs to the serial adder is the standard OUT A output (see Fig. 2); the other input is IN A (user definable). The output of the serial adder is OUT B.

Parallel input and parallel output are available on the 48-d.i.l. package version via a three-state 10-bit bus, for easy interfacing to microprocessor systems. Parallel input or output may be used in conjunction with serial output or input as required.

The gain of the signal at the output may be varied over 8 bits via 3 external control pins (G in Fig. 2). This allows the user to obtain maximum resolution at the output for most sets of coefficients.

5.3 Cascade Option

The device has been designed in such a way that two devices may be cascaded to provide a filter of 96 coefficients. Figure 3 shows the interconnections required: the oldest signal stored in LSI1 becomes the newest sample for LSI 2. The two accumulator outputs are added together in either processor's serial adder, and the full output is available from that processor's OUT B

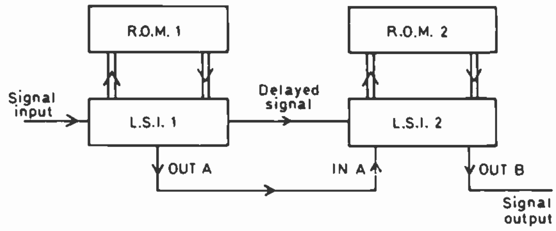

Fig. 3. Block diagram of a 96-tap filter.

(or parallel output). The maximum sampling rate of the 96 coefficient arrangement is limited only by the larger of the numbers of clock periods required to process the signals in the two l.s.i. devices.

6 Coefficient Resolution

The processor coefficient resolution is nominally 8 bits, but this resolution can be increased if a penalty in throughput can be tolerated. The increased resolution is obtained by breaking the CSDC rules at the most significant end of the coefficient, meaning that, in general, extra clock periods are required to process the coefficient.

Considering the coefficients as decimal integers, 8-bit coefficients have a range $+127$ to -127,

i.e. 0111 1111 to 1000 0001 in two's complement binary

or 1000 000$-$1 to $-$1000 0001 in CSDC

9-bit coefficients have a range $+255$ to -255

i.e. 0 1111 1111 to 1 0000 0001 in two's complement binary.

It is possible to express this extended range as

2000 000$-$1 to $-$2000 0001

where the 'bit' '2' means the signal is processed twice at the given significance. For example, with a coefficient of $+255$, the accumulator receives:

Clock period 1	$-1 \times$ signal
2	$128 \times$ signal
3	$128 \times$ signal
	$255 \times$ signal

Using this technique, coefficient resolutions greater than 9 bits can be implemented, but the throughput penalty can be considerable, and the processor accumulator may overflow. One would seldom wish to extend beyond 10 bits.

7 Applications

This Section is not intended as a filter design aid, but merely to show some of the capabilities of the hardware described. References 3 and 4 give approximate design relationships for finite impulse response low-pass filters. These relationships can be used to determine the suitability of the l.s.i. processor to a proposed task.

All of the filter designs in Figs. 4 and 5 are based on Kaiser window-functions.[5] N is the number of taps of the filter simulated, T is the number of taps per lobe of the $(\sin x)/x$ curve of the window function, K is the Kaiser factor and Q is the number of bits per coefficient. All frequency scales have been normalized to the half

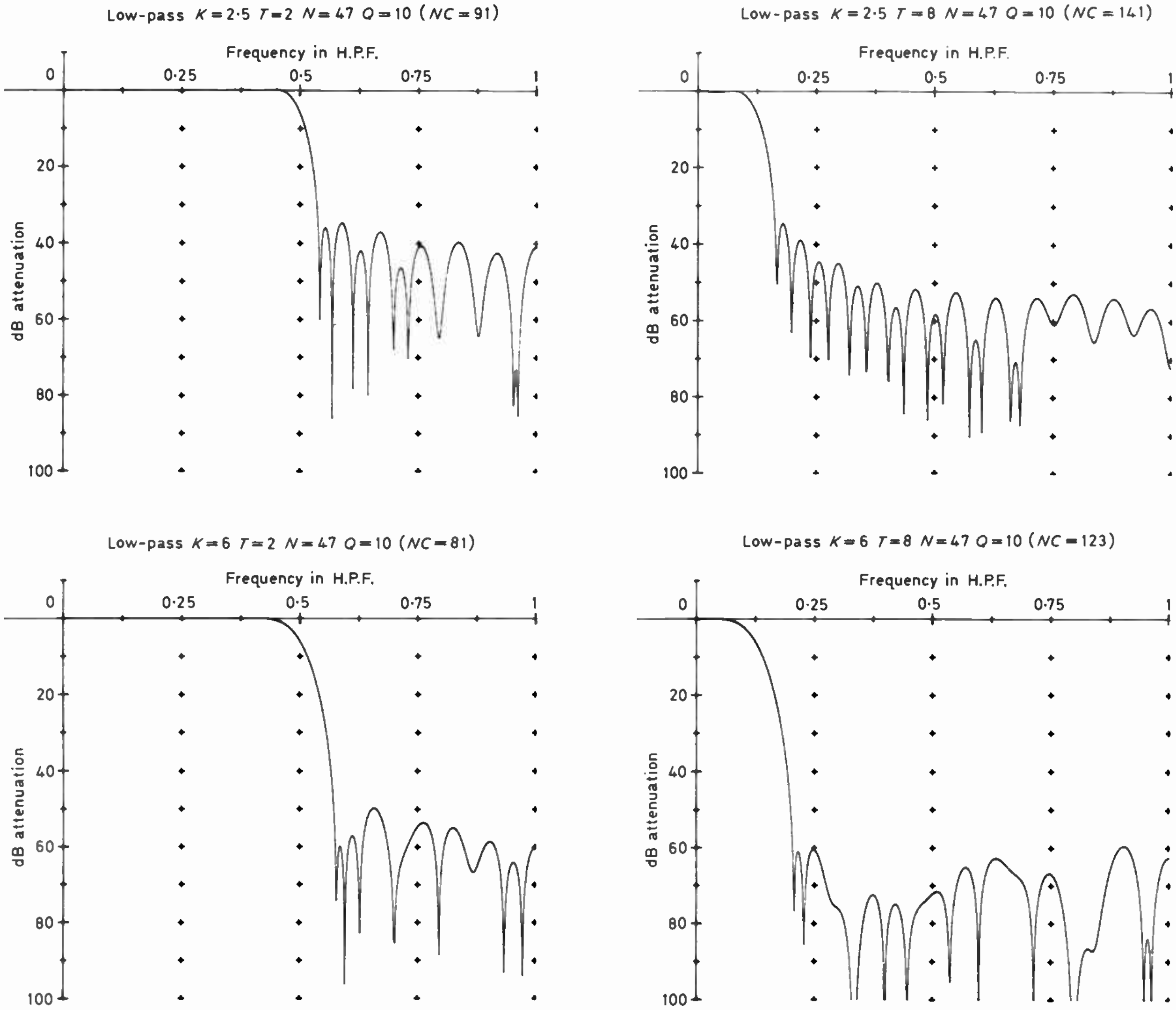

Fig. 4. Examples of responses of low-pass filters.

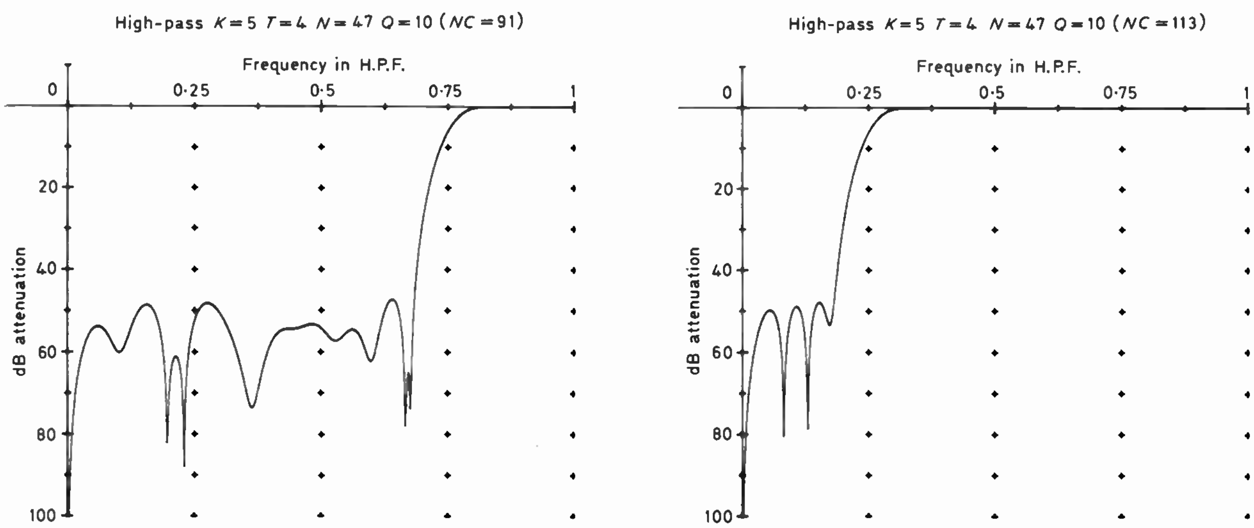

Fig. 5. Examples of responses of high-pass filters.

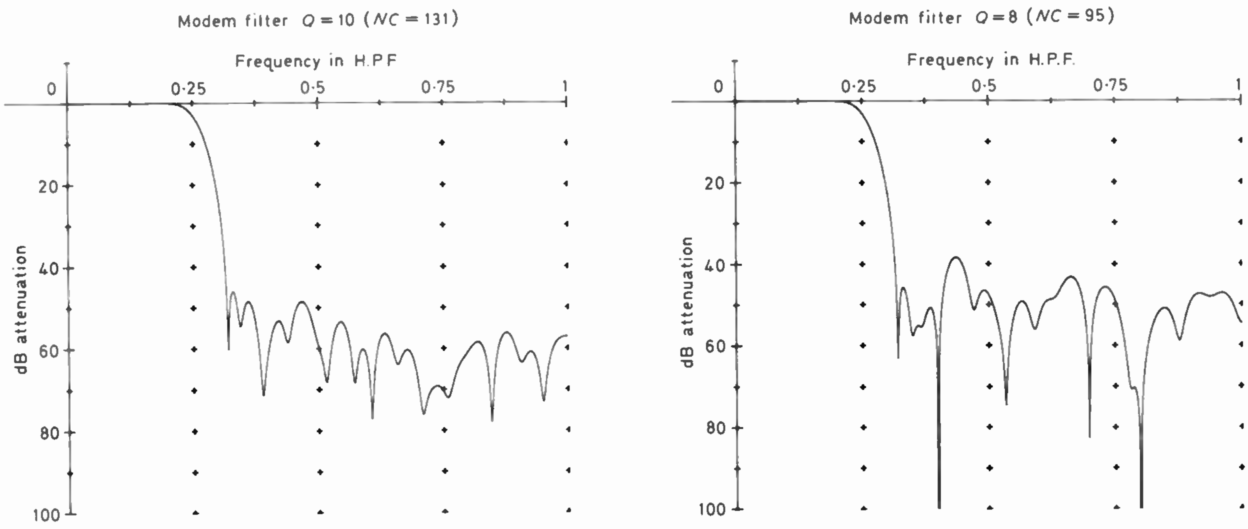

Fig. 6. Examples of responses of modem filters.

repetition frequency (HRF) of the filter. The diagram also show the minimum number (NC) of clock periods required to process the filter. Serial input and/or output will sometimes require extra clock periods.

7.1 Low-pass Filters

Some linear-phase Kaiser window-function filters are shown in Fig. 4. They give some indication of the trade-off between rate of roll-off, minimum stopband attenuation and passband/stopband ratio (or number of samples per lobe of the $(\sin x)/x$ curve). In all cases 10-bit coefficients are used.

7.2 High-pass Filters

Two examples are shown in Fig. 5 computed via different transforms of the same Kaiser window-function filter. Although 24 taps/side were calculated (leading to a 49-tap filter), the outer taps are zero, so only 47 taps need to be implemented.

7.3 Modem Filters

One application for the l.s.i. devices is to provide transmit and receive filtering in a q.a.m. 9600 bit/s data modem. Each filter has to provide a 40 dB stopband attenuation and half of the overall spectral shaping. In addition the receive filter has to be minimum phase, with the phase response of the two filters in tandem being linear. The two filters form a matched pair, with one time response being the time-inverse of the other. Figure 6 shows the responses of the modem filters with coefficients quantized to 10 and 8 bits.

8 Conclusions

A single-chip custom-l.s.i. 48-tap digital processor is being made which can satisfy non-recursive filtering requirements with sampling rates up to 30 000 samples/s. A complete filter can be made using one custom-l.s.i. device, one 256 × 8 r.o.m. and two clock sources. The desired filter coefficients are stored in the r.o.m. in a specially coded form.

Although the processor is intended initially for use in a high-speed modem, it is expected to find other applications; it will for example function as a microprocessor peripheral.

9 References

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UDC 621.372.54:621.374

Indexing Terms: Digital signal processing, Transversal filters, Pulse compression

Economical filters for range sidelobe reduction with combined codes

MARTIN H. ACKROYD, B.Sc., Ph.D., SMIEEE*

SUMMARY

It is shown that approximate inverse filters for combined codes can be implemented using far fewer coefficients than required for direct implementation. This enhances the usefulness of recently reported work on the subject.

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In 1973, Ghani and the present writer pointed out that least-squares approximate inverse filtering techniques, developed by geophysicists, could be used for range sidelobe reduction in pulse compression systems.[1] Morgan, Dassanayake and Liberg[2] have recently presented an interesting paper on the use of these techniques in pulse compression systems using combined Barker codes. A combined code is one formed by combining two codes; each element of one code is replaced by a complete version of the other. The work of Morgan, Dassanayake and Liberg is useful because combined codes generally give undesirably large range sidelobes if used with a matched filter.

One advantage of a combined code, not mentioned in Ref. 2, is that a pulse compression filter for it may be implemented using far fewer coefficients than are needed, in general, for an arbitrary code of equal length. For example, if the codes used to form the combination code have $M+1$ and $N+1$ elements, a matched filter can be implemented as a cascade of two transversal filters with a total of only $M+N+2$ tappings, rather than $(M+1)(N+1)$, the number of elements in the combined code.[3] The purpose of this note is to point out that similar economies are possible with approximate inverse filters. This greatly increases the practical usefulness of the work reported in Ref. 2.

Filters for combined codes can be discussed using z-transforms.[3] If the codes to be combined have z-transforms $A(z)$ and $B(z)$, the combined code has a z-transform given by

$$C(z)=A(z)B(z^M),$$

where the code whose z-transform is $A(z)$ is the one which is used to replace each element of the other code. The inverse filter for the combined code would thus have the transfer function $1/[A(z)B(z^M)]$ but this will be physically non-realizable for any code of practical interest.[1] However, approximations to the ideal inverse filter, with delay to ensure realizability, can be synthesized by cascading two filters: one an approximation to $z^{-K}/A(z)$ and one an approximation to $z^{-LM}/B(z^M)$, where K and L correspond to suitably chosen delays.

The synthesis of these approximate inverse filters is straightforward. Least-squares approximations to $z^{-K}/A(z)$ and to $z^{-L}/B(z)$ are synthesized using the methods set out in Refs. 1 and 2. The computer programs given by Robinson[4] can be used directly for this purpose. The coefficients of the least-squares approximate inverse filters, $(a'_1, a'_2, \ldots, a'_Q)$ and $(b'_1, b'_2, \ldots, b'_R)$ can then be used directly in the filter structure of Fig. 1 to give an approximate inverse filter for the combined code. Because the second stage of the filter uses delay elements with M times the delay of the elements in the first stage, it gives an approximation to $z^{-LM}/B(z^M)$, as required, rather than to $z^{-L}/B(z)$, for which the coefficients were synthesized.

As well as giving a filter using fewer coefficients, this approach also eases the task of computing the filter coefficients. This is because the computation needed for the two short filters is normally much less than that

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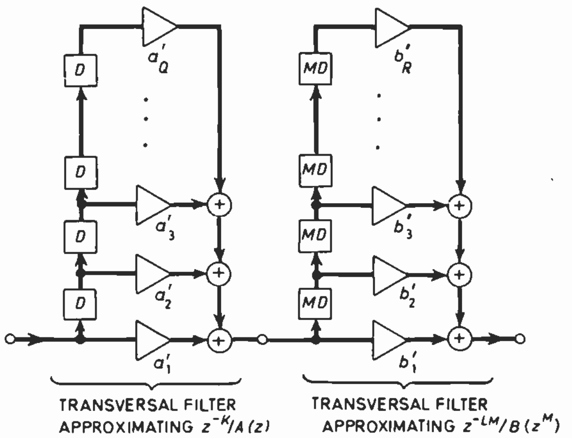

Fig. 1. Block diagram of two-stage approximate inverse filter for a combined code. The second stage has delay elements with M times the delay of the delay elements D of the first stage.

needed for one long one. A filter to compress a combined code formed by combining two 13-element Barker sequences may be considered as an example. Reference 1 gives the coefficients of a least-squares approximate inverse filter, of length 66, which gives a peak sidelobe, in response to a 13-element Barker code, about 60 dB below the main lobe. By using these coefficients in both stages of the filter of Fig. 1, a filter is obtained which will reduce the sidelobes of the 169-element combination code to about 60 dB below the main lobe. This filter has only 132 coefficients, appreciably fewer than some of the filters of Ref. 2, yet it applies to a longer combined code and gives greater sidelobe suppression.

Reference 2 mentions the loss in signal-to-noise performance incurred when an inverse filter is used in place of a matched filter. Of course, an inverse filter is ideal only when the dominant form of interference is self-clutter and where receiver or other incoherent noise is negligible. When the dominant form of interference is incoherent noise, a matched filter is ideal. In other situations, a compromise between an inverse filter and a matched filter is best. It is perhaps worth re-making the point that Robinson's formulation enables a filter to be computed which gives the best compromise between an inverse filter and a matched filter.[1,5] In principle, it would be possible to adapt the receiver to suit the prevailing ratio of self-clutter to incoherent noise.

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Book Review

Antenna Theory and Design
ROBERT S. ELLIOTT (University of California, Los Angeles)
Prentice-Hall, Englewood Cliffs, New Jersey, 1981
18 × 24 cm. 594 pages. £26.60

CONTENTS: Source/Field Relations, Single Antenna Elements: The far field integrals, reciprocity, directivity. Radiation patterns of dipoles, loops and helices. Radiation patterns of horns, slots and patch antennas. Array Analysis and Synthesis: Linear arrays, analysis. Linear arrays, synthesis. Planar arrays, analysis and synthesis. Self-impedance and Mutual Impedance, Feeding Structures: Self-impedance and mutual impedance of antenna elements. The design of feeding structures for antenna elements and arrays. Continuous Aperture Antennas: Travelling wave antennas. Reflectors and lenses.

A good balance of material at undergraduate level is aimed at by the author and on the whole I believe that he has achieved this. The book is nicely laid out and has a quality feel about it indicating much care in preparation. In essence the text comprises theory first and antenna types second, but is in fact organized into four parts that are further subdivided into chapters.

The coverage of theory and antenna types is fairly standard in antenna books but it is the logical breakdown into ready assimilated steps that is of particular value to students. In these respects the author has created a worthwhile account, linked together with lucid comments and written very much from the engineer's standpoint. The author admits that he has drawn mainly on his own and local experience and it might be said that some of the content is somewhat dated and of limited scope. I personally see this as an advantage in that the author is in a position to introduce practical realism to the student.

The introductory electromagnetic theory briefly embraces the basic theorems and formulas needed for antenna calculations and, like the rest of the book, is presented from an engineering standpoint. Students with insufficient grasp of the mathematical detail will not necessarily be helped by this approach and the brief suggestions for problem exercises will give little assistance in this respect. There are no worked examples throughout the text although this is a common omission in student books of late. With these few reservations I can recommend this book to those teaching antenna engineering and would be pleased to have it on my bookshelves.

J. R. JAMES

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.

JUNE

SCOTELEX '82 *8th to 10th June EDINBURGH*
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 COPENHAGEN*
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 DALLAS*
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 LONDON*
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 BRIGHTON*
*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 BERGEN*
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 PARIS*
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 *30th June to 1st July MANCHESTER*
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

***Computing** *5th to 7th July GLASGOW*
A conference on Distributed Computing, organized by the Science and Engineering Research Council is to be held at the University of Strathclyde. Information: Mr W. D. Shepherd, University of Strathclyde, Department of Computer Science, Livingstone Tower, 26 Richmond Street, Glasgow G1 1XH (Tel. 041-552 4400 Ext 3299).

●Man/Machine Systems *6th to 9th July MANCHESTER*
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 LONDON*
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 DENVER*
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 HULL*
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 YORK*
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 EDINBURGH*
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 EINDHOVEN*
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

***SIM-HI.FI** *2nd to 6th September MILAN*
International Exhibition of Music and High Fidelity, organized by ANIE, will be held at the Milan Fair Centre. Information: General Secretariat SIM-HI.FI-IVES, Via Domenichino 11,20149 Milano, Italy.

Microwaves *6th to 10th September HELSINKI*
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 LIVERPOOL*
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 LONDON*
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 LONDON*
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 BRISTOL*
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 MANCHESTER*
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 ANAHEIM*
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 BRIGHTON*
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 YORK*
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 GUILDFORD*
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 PARIS*
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 CANNES*
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 LONDON*
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 BADEN-BADEN*
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 LUTON*
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).

OCTOBER

Electronic Displays *5th to 7th October LONDON*
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 ATHENS*
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 BRIGHTON*
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 LONDON*
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 LONDON*
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).

● Military Microwaves '82 *19th to 22nd October LONDON*
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

·Electronics *20th to 22nd October HONG KONG*
Second Hong Kong Electronics Fair, sponsored by the Hong Kong Exporters' Association and the Hong Kong Electronics Association, will be held at the Miramar Hotel, Kowloon. Information: Secretary, Hong Kong Electronics Fair Committee, c/o Hong Kong Exporters Association, 1625 Star House, 3 Salisbury Road, Kowloon, Hong Kong.

Multivariable Systems *26th to 28th October PLYMOUTH*
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 LONDON*
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 LUTON*
(See item for 30th September)

Pattern Recognition *19th to 22nd October MUNICH*
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 SAARBRUCKEN*
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 GAITHERSBURG*
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 LONDON*
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 SEVENOAKS*
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 LUTON*
(See item for 30th September)

DECEMBER

·ONLINE *7th to 9th December LONDON*
The Sixth International Online Information Meeting, organized by *Online Review*, will be held at the Cunard Hotel, London. Information: Organizing Secretary, Online Information Meetings, *Online Review*, Learned Information, Besselsleigh Road, Abingdon, Oxford OX13 6LG. (Tel. 0865-730275)

·INDEX-EL '82 *10th to 15th December ATHENS*
Second International Electrical & Electronics Engineering Exhibition, will be held at The Zappio Palace, Athens. British exhibits sponsored by EEA. Information: EEA, 8 Leicester Street, London WC2H 7BN. Tel. 01-437 0678.

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FEBRUARY

MECOM '83 *7th to 10th February BAHRAIN*
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*
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 BRIGHTON*
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 BIRMINGHAM*
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 PARIS*
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.

·● ICAP '83 *12th to 15th April NORWICH*
Third International Conference on Antennas and Propagation organized by the IEE in association with the URSI, IEEE, IMA, IoP and the IERE, will be held at the University of East Anglia, Norwich. Information: IEE Conference Department, Savoy Place, London WC2R 0BL. (Tel. 01-240 1871, ext. 222)

MAY

·Noise *17th to 20th May MONTPELLIER, FRANCE*
Seventh International Conference on Noise in Physical Systems/3rd International Conference on 1/f Noise, will be held in Montpellier, France. Information: Dr B. Jones, Department of Physics, University of Lancaster, (Tel. Lancaster 65201); or Professor H. Sutcliffe, Department of Electronic & Electrical Engineering, University of Salford.

·Electron Tubes *18th to 20th May GARMISCH-PARTENKIRCHEN, WEST GERMANY*
Conference on Electron Tubes organized by VDE (NTG) in association with the German Section of the IEEE, will be held in Garmisch-Partenkirchen, Bavaria. Information: Conference Chairman, Dr H. Heynisch, Siemens AG, Werk für Röhren und Sondergebiete, St Martinstrasse 76, D-8000 Munchen 80.

JUNE

·IOOC '83 *27th to 30th June TOKYO*
The Fourth International Conference on Integrated Optics and Optical Fibre Communication, sponsored jointly by the Institute of Electronics and Communication Engineers of Japan and the Institute of Electrical Engineers of Japan, will be held at Keio Plaza Hotel, Tokyo. Information: Y. Suematsu, Department Elec. Phys. Tokyo Institute of Technology, 2-12-1, O-okayama, Meguro-ku, Tokyo, 152 Japan.

JULY

·● Electronics *6th to 8th July LEEDS*
The Leeds Electronics Show, organized by the University of Leeds and Evan Steadman Services, to be held at the University of Leeds. Information: Theresa Austin, The Leeds Electronics Show, 34–36 High Street, Saffron Walden, Essex CB10 1EP (Tel. 0799-22612)

SEPTEMBER

·Simulators *26th to 30th September BRIGHTON*
International Conference on Simulators, organized by the IEE, will be held at the University of Sussex. Information: IEE Conference Department, Savoy Place, London WC2R 0BL (Tel. 01-240 1871) *(Synopses by 4th October.)*

Weightech '83 *13th to 15th September LONDON*
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 GENEVA*
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).

·Security Technology *4th to 6th October ZURICH*
17th Carnahan Conference on Security Technology, organized by the Institute for Communication Technology at the Eidg. Technische Hochschule Zurich, in association with the College of Engineering, University of Kentucky, will be held in Zurich. Information: P. de Bruyne, ETH Zentrum-KT, CH-8092 Zurich, Switzerland. (Tel. 411-2562792)

Organizers of appropriate events are invited to submit details to the Editor for inclusion in this calendar.
