Volume 52 Number 3 March 1982

The Radio and Electronic Engineer

Journal of the Institution of Electronic and Radio Engineers

EDITORIAL

IERE NEWS AND COMMENTARY

GENERAL INTEREST ARTICLE

Radio History

The merger of the Appleton Laboratory with the Rutherford Laboratory and its transfer to Chilton marked the end of some 60 years of radio and space research at Slough. The main trends of research followed during that period which saw the development of radio as a major area of science and technology are described subject by subject in this valedictory article.

PAPERS

Filter Design

Partial response filters are used in data transmission systems as they allow an increased bit rate to be used for a certain available band width. The paper presents a design procedure which keeps intersymbol interference errors at a minimum. No separate group delay equalizer is required, thus reducing the number of elements. The computer design time is extremely short: every pole is used as a degree of freedom and few zeros are necessary.

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Volume 52 No. 3 March 1982

The Radio and Electronic Engineer

The Journal of the Institution of Electronic and Radio Engineers

A Strategy for the British Electronics Industry

There is general agreement that the electronics industry of this country has lost the impetus, the inspiration, the initiative which it demonstrated in the decades immediately following the War. It is quite clearly not growing at the rate of its competitors around the world, and its share of the market is notably lagging for those products which are elsewhere demonstrating the fastest opportunities for expansion. To state such facts is easy, what is not easy is to distinguish the contributory factors and, more important, to prescribe corrective measures.

An authoritative and thoughtful attempt to do just these things has recently been made by the Economic Development Committee for Electronics (EEDC) which its chairman, Sir Henry Chilver, has set out in a brief report that has been welcomed by the Minister of State for Industry and Information Technology, Mr Kenneth Baker, as the most important NEDO pronouncement of the year. Subsequently, at a meeting of the National Economic Development Council on February 3rd, a response to the EEDC document by the Minister was received, entitled 'A Programme for Action'. To a certain extent the two documents only reiterate what has been done, or proposed, and not a great deal of 'new thinking' can be discerned in either. As a survey of industry and government interaction however, they are, together, of interest to electronic and radio engineers.

How the EEDC see the Problems and their Solution

The problem has been quantified by the following paragraph in Sir Henry's report: 'In the latter part of the 1970s, the UK market for all electronics products was growing at around 10% a year; this growth was, however, being increasingly met by imports, and a trade surplus of £49M in 1975 was transformed into a trade deficit of nearly £200M in 1980. Total output of the industry in 1980 was £7600M; growth in output in the latter part of the decade has been running at below 9% per year. Over 50% of the UK market is accounted for by the public sector. The industry employs about 470 000 people, and this figure has only marginally declined in the last decade.'

The EEDC sees the objective of its policy as increasing the industry's share of the world market: first by identifying strategically important issues to be tackled systematically and second to consider these issues within the framework of a medium-term policy for the industry.

It is pointed out that UK firms show a broad spread of involvement across technologies and products, but, compared to their international competitors, their operating units are small and fragmented. Moreover, the majority of UK firms are highly concentrated on the domestic market, which is too small to sustain some of the larger systems developments. The UK's overseas competitor firms have increased their shares of the world market—often at our expense—by the application of business strategies which build on their internationally competitive strengths. The building up of the share of the world market will mean that policies in companies and in government must be aimed at building selectively on existing or potential business strengths (and to plug weaknesses which threaten to undermine strengths) in the internationally tradable part of the industry's output. This could mean that, as a deliberate act of policy, certain products and technologies should receive no public support in order to maximize assistance for others.

Those sectors within which the UK could build up competitive businesses include information technology systems; the development of high volume manufacturing capability for information technology terminals and peripherals by building on viewdata and teletext; the application in IT systems of existing strengths such as radio communications; the commercial and civil application of technology from the defence sector; and the underpinning of these equipment sectors through the fostering of core technologies, especially microelectronics, opto-electronics, displays, sensor devices, advanced interconnect techniques and other key components.

The report acknowledges that firms are in favour of this approach but find it difficult to balance commercial objectives and the national interest in making selective choices and channelling the necessary resources. The need is therefore seen for mechanisms to involve the government and industry in perceiving and responding to changing markets and technologies.

The EEDC is now proposing a number of specific approaches to the problem of supporting the potential winners that this collaboration has identified. First of all firms must work together on projects to share both costs and technological know-how: if necessary restrictive practices legislation must be modified. Second comes the long-standing complaint by British industry that it suffers in competition with overseas firms who can count on direct support or guarantees from their government to mitigate risks. The third proposal is that the financial institutions and firms must draw closer to reach a more satisfactory relationship where risk capital is needed. Fourthly, the EEDC believes that procurement policies should be changed from the 'cost-plus' contractual agreements to fixed price agreements where risk is borne by the companies, and in the process lead to better manufacturing disciplines. The fifth point made relates, inevitably, to the use of resources, particularly manpower. Closer exchange of information between firms and between firms and the state to manage market and technological changes more effectively are regarded as essential. But the impact on the electronics industry of university cuts at a time when shortage of graduates is but one aspect of the shortages of skilled manpower are obviously crucial. So, too, is the reduction of industry's investment in training caused by the economic situation.

What the Government has been Doing and Proposes to Do

The first observation made by the Department is that the EEDC paper calls for a number of initiatives on the part of Government. Many of these recommendations have been explicitly accepted by Government as is subsequently pointed out. But few of the EEDC's recommendations call for action on the part of industry. Yet the objective of strengthening the international position of the UK information technology (IT) industry and establishing it as a major competitor in the overall world market will only be attained if all the industrial and financial partners in British industry co-operate in seeking new business growth and international competitiveness. It must therefore be the responsibility of managers and workforce in the supplying industry to pursue positive business objectives; but there is also a responsibility on managers in the investment sector, and in the application areas, to take investment and procurement decisions with the long-term objectives of the UK industry in mind.

The current world deficit of UK IT of about £230M has been estimated as increasing to £1000M by 1990; in major part it seems likely that the poor UK performance relative to Japan and the US is attributable to the scale economies achieved in these two countries compared with the relative fragmentation of the UK industry.

Objectives

In the Government's strategy, the dissemination of new technology through the economy is seen as a critical factor in determining its overall competitiveness. Slow diffusion relative to competitor countries means lost markets generally at home and abroad, missed opportunities for new business and employment creation, and failure to make productivity gains. It also means lost markets for the UK supplying industry, and slower product development compared with our competitors. Acceleration of the diffusion rate of the new technology through the economy is thus a critical factor for the economy generally, and for the supplying industry in particular: it is a key objective of Government.

Secondly, the Government is responsible for setting the Infrastructure and domestic environment in which the IT industries can achieve growth, principally through telecommunications and broadcasting policies, data protection and national standards.

Thirdly, the Government provides direct support to increase the efficiency and technological performance of UK industry, in order to achieve an internationally competitive supply industry. Ultimately this is a task which industry must perform by itself: the role of Government in this area is therefore essentially catalytic, acting principally through the channels of direct R & D support, and public purchasing policy.

Implementation to Date

To meet these three objectives the Government considers that it has already established a positive programme of action aimed at increasing the diffusion of the new technology and strengthening the position of the UK Industry.

Lack of awareness is being remedied by the designation of 1982 as Information Technology Year during which many special events are being arranged. A number of campaigns aimed at specific sectors have been started over recent years such as the Microprocessor Application Project, the Teletext and Viewdata Campaign, and Robotics and CAD/CAM schemes.

In establishing the infrastructure, the 'liberalization' of the telecommunications monopoly is seen as opening significant benefits for suppliers and users by exposing the supply industries more fully to international competition. More urgency in developing national standards and their compatibility with international standards is considered crucial to successful exporting.

In 1981 the Government signed the Council of Europe Convention on Data Protection (bowing in fact to considerable industry pressure!) which will enable British firms to provide computer services to overseas customers who already have such safeguards.

In the role of primary and secondary education the Micros in Schools and associated schemes—having the objective of enabling every 16-year-old leaving school after the end of 1982 to have had the opportunity to have hands-on experience of a microcomputer are regarded as highly significant in fostering IT thinking. (Although there are those in industry and higher education who believe that attention to more conventional numeracy—and of course literacy—is just as important to information technology, as has been pointed out in this Journal on several occasions.)

Training under the various schemes operated by the Manpower Services Commission at levels from the unqualified school leaver to the prospective electronics systems designer is playing an important part. The hope is also expressed that the IT centres for training will lead to many small firms being formed.

Support for the industry by making it more competitive is next examined in the document and it is conceded that the UK IT industry receives less than do those of France and Germany, for instance. (About £500M compared with £750M and £1000M respectively for 1970–1980.) Support already goes to some new IT products, space technology, telecommunications and, more recently, to a selected number of strategic sectors (e.g. fibre optics, microprocessor applications, microelectronics, robotics). It is also pointed out that public sector procurement, such as that by the Ministry of Defence with £5000M spent on equipment each year mostly goes to UK suppliers. The public sector generally is being urged to develop closer links with potential suppliers and to set specifications in ways that take account of world market needs. International collaboration, by joint ventures between companies, is favoured and agreements with the European Space Agency and on Microelectronics with the European Commission are other instances of this approach.

Selectivity in Practice

Having set out its own contributions, the Ministry document then turns to the EEDC's call for selective concentration of public resources. With some asperity, it suggests that the 'shopping list' is very broad and a far higher degree of definition will be necessary if increased selectivity is to have any meaning in these areas.

The Government's own application of selectivity is instanced as follows:

IT Systems: Development of the Office Automation Strategy together with a substantial extension of the Software Products Scheme.

Viewdata/Teletext: A continued Teletext awareness campaign has been mounted.

Radio Communications: Past defence support in this sector has resulted in the UK being well placed in military export markets. To remain internationally competitive, however, the industry must take the fullest advantage of current development in technology and the commercial application of defence communications equipment to civilian markets. It will also depend critically on the regulatory environment and the Government is now examining this.

Commercial Application of Defence Technology: Current arrangements already provide for companies carrying out R & D for the MoD to use the resulting technology for civil purposes; the Ministry also has comprehensive arrangements for consulting industry on the defence research programme. The prime responsibility for the commercial application of this publicly funded knowledge and technology must obviously rest with industry. This will involve industry more closely in the process of determining the design of defence equipment, particularly in cases where industry contributes to the cost of the developments, and should help to improve the commercial attractiveness and export relevance of equipment developed under the auspices of the defence budget.

Core Technologies: Existing support schemes already highlight microelectronic technology and optical fibre/opto-electronics and particular attention is being paid to flat screen technologies. £17M of the £25M Government funds allocated to fibre optics are already committed, yet there remain aspects which require further stimulation.

Beyond these areas the Government has distinguished and is encouraging related developments: cable and satellite systems including the 'broadband into the home' concept; fifth generation computing; databases; and potential linkings such as cable systems and entertainment with local networking, and applying IT to the machine tool industry.

Deciding Priorities

While the Government expresses willingness to consider all these areas with major UK electronics companies, it points out that the present programme of action is already based on perceived priorities from which it would be undesirable to withdraw support. Again, selectivity means exclusion: if a policy of increased selectivity is to be pursued, industry will have to accept that there will be no special effort by Government in non-selected areas.

A final observation in the document is that if the aim of increased international competitiveness is to be achieved it is vital that the role of industry and the finance sector should be fully understood and accepted. The involvement of Government should not be allowed to mask the need for initiatives from the private sector.

It is understood that the NEDC did not dissent from the general content of the Minister's presentation of the document and indeed Sir Henry Chilver expressed pleasure that his Committee's report had met with an overall favourable response. Clearly further discussions at a variety of levels will follow: electronic and radio engineers, most of whom must now regard themselves as being part of the Information Technology industry, will be able to judge the extent to which their own part of that industry is likely to be affected. The broad coverage of the industry must pose the question as to whether there is a clear enough central theme for the whole programme to be realistically supported in the present state of the national economy. But that is a pessimistic approach which the EEDC and the Government must between them counter both by word and deed: good intentions must be backed on all sides by performance if this strategy is to succeed.

F.W.S.

IERE Benevolent Fund

Minutes of the Annual General Meeting of Subscribers

Held at the London School of Hygiene and Tropical Medicine on 29th October 1981

The meeting was opened at 6.30 p.m. by the President, Professor William Gosling, and the Secretary confirmed that, in accordance with Clause 6(i) of the Trust Deed, advice of this Annual General Meeting was contained in a notice which appeared in the July/August 1981 issue of *The Radio and Electronic Engineer* (Volume 51, No. 7/8) and that the detailed agenda had been issued to those present who had subscribed to the IERE Benevolent Fund during the year.

Annual Report of the Management Committee

The President then presented the Annual Report of the Management Committee for the year ended 31st March 1981, and said:

'The Trust Deed which now governs the affairs of our Benevolent Fund assigns the care of the Fund to three primary members of the Management Committee, namely "the President, Treasurer and Secretary of the IERE for the time being in office", and it is in that capacity that I, as Acting President for the period covered, submit to you the 1980/81 Report, which was published in the October 1981 issue of the Journal.

'It is a simple, straightforward report, making the key points appropriate to a Fund of this sort: first, that the assets of the Fund have been properly cared for in accordance with the law and the rules of the Fund, and second, that no deserving applicant has been refused appropriate assistance during the period of the Report. All three members of the Management Committee are present here this evening, and we will gladly respond to any questions that entitled voters, i.e., any that have subscribed to the Fund during the year under report, have to raise.'

No questions were asked and the President formally moved the adoption of the Annual Report of the Institution's Benevolent Fund for the year ended 31st March 1981. This was approved unanimously.

Income and Expenditure Account and Balance Sheet

Turning to Item 2 of the agenda, the President offered for approval the audited accounts of the Benevolent Fund for the year 1980/81, which were published in the October 1981 issue of *The Radio and Electronic Engineer*, and said:

'I would make only one additional comment. Because of the uncertainty of the money markets throughout the year and the fact that for most of that time our investments were in process of transfer to the Official Custodian Trustee, we thought it best to keep the Fund as liquid as possible. This has allowed us to benefit from the higher interest rates offered for most of the period by the short-term money market. We will, however, now be taking expert advice to see whether this is the time to revise that policy and to make a thoroughgoing review of the schedule of investments listed in the Accounts now before you. Once more, we will gladly respond to any questions that entitled voters may wish to raise.'

There being no questions the President then formally moved the adoption of the Income and Expenditure Account and Balance Sheet of the Institution's Benevolent Fund for the year ended 31st March 1981, which was unanimously approved.

Appointment of Solicitors and Accountants

The President then requested approval for the re-appointment of Bax, Gibb and Company as Solicitors for the Fund, and for Gladstone, Jenkins and Company to be re-appointed as the Fund's Accountants. These appointments were unanimously approved.

Any Other Business

Having been advised by the Secretary that no notice of any other business had been received, the President then formally declared the Annual General Meeting of Subscribers to the IERE Benevolent Fund closed.

Alternatives to Gold for Thick-Film Multilayer Circuits

Hitherto, gold has been the standard conductor material for the thick-film multilayer structures which provide the dense interconnections required for present day electronic equipment. But the high price of gold and the pressure generally to reduce costs has led to a search for cheaper alternatives. These substitutes are now to be the subject of an independent technical assessment planned by ERA Technology.

One recent introduction is a group of copper conductor pastes with matching thick-film dielectrics. Copper is an attractive alternative to gold, offering high electrical conductivity at a much lower intrinsic cost. Thick-film copper multilayer circuits are now being used in large-scale computer manufacture in Europe and elsewhere. However, these copper thick-film systems have to be fired in special furnaces with closely-controlled nitrogen atmospheres and processing problems have been encountered.

Some manufacturers are therefore contemplating turning to silver-based systems which are also much cheaper than gold but which can be fired in air in conventional furnaces. In the past, silver-based conductors have not been widely used in multilayer work because of the risk of silver ion migration in the presence of an electric field, but improved compositions are claimed to have overcome this problem.

A further option is to retain gold as the conductor but to use special formulations which fire out to give much thinner deposits than conventional gold pastes, thus saving on materials. There is some reduction in performance but no new reliability hazard is introduced.

Clearly, choosing the best alternative low-cost material for a particular thick-film multilayer application is a complex matter, and since reliable comparative data are not available, individual users have to carry out lengthy and expensive evaluations. To remedy this situation, ERA is to conduct a thorough assessment of alternative thick-film interconnection systems for hybrid circuit manufacturers. This will be funded on a group-sponsored basis and will provide users with an impartial evaluation of commercially-available systems at a fraction of the cost of undertaking the work themselves.

The project will be launched early in 1982 and those interested in participating should contact Keith Browne at ERA Technology Ltd, Cleeve Road, Leatherhead, Surrey KT22 7SA. (Telephone: Leatherhead (03723) 74151, Ext 254.)

Major Prizes for Opto-Electronics Inventions

The Rank Prize Funds have announced the award of three major prizes to scientists who have made significant contributions in the fields of Opto-Electronics and Nutrition. The Rank Prize Funds were established by Lord Rank with donations totalling £2·5M just prior to his death in 1972 to encourage work in these two areas which he believed would be of special benefit to mankind.

There are two prizes for Opto-Electronics which explores the interface between optics and electronics. One, worth £30,000, goes to Professor Calvin F. Quate in recognition of his contribution to medical, biological and physical research through the concept of the scanning acoustic microscope, which uses sound rather than light to form images. Calvin Quate is Professor of Applied Physics and Electrical Engineering at Stanford University, California.

The second Opto-Electronics award is of £10,000 to Dr C.T. Elliott for devising a novel infra-red detector which has given Britain a considerable advance in areas such as medical treatment, where the new technique enables the 'hot spots' associated with disease to be examined in detail. Dr Elliott is now a Senior Principal Scientific Officer leading a research team at the Royal Signals and Radar Establishment at Malvern, Worcestershire.

The Rank Prize for Nutrition is £20,000 and goes to Dr Hamish Munro, a Scot who is now at Tufts University, Boston, and is Director of the US Department of Agriculture's Human Nutrition Research Centre on Aging. It is his long-standing research establishing the balance of nutrients required as the body grows older for which he receives recognition by this latest award.

The Rank Prize Funds make awards on an occasional rather than an annual basis, at the discretion of a board of distinguished Trustees. The Trustees are advised by two specialist committees of senior researchers and academics. The Funds also promote the spread of knowledge in Opto-Electronics and Nutrition by supporting research and organizing international symposia. Disbursements have so far totalled more than £1,340,000.

Scanning Acoustic Microscope

In 1973 Professor Quate conceived an approach of elegant simplicity to produce a microscope that would use sound, rather than light, in order to form images. This achievement, which had been the aim of applied scientists for more than fifty years, led to the extremely rapid development of a microscope which already exceeds the resolution of optical microscopes.

The key idea, which was the recognition of the fact that velocities of acoustic waves in some solids can be as much as seven times greater than the velocity in water, resulted in the production of a lens which could focus a beam of sound, on its axis, without significant aberrations. Whilst such a lens cannot image a complete field, Professor Quate recognized that the axial focus was enough for the realization of a mechanically-scanned microscope in which the image was reproduced point by point. The scanning acoustic microscope has opened up a completely new field of microscopy which permits the direct imaging of biological specimens and the examination of silicon integrated circuits and other solid objects.

The work of Professor Quate will prove of lasting value to medical, biological and physical research and thereby benefit mankind by improving the quality of life.

A New Infra-red Detector

In 1974, from an insight into the mechanism of interaction between incident infra-red radiation and the electron population in a semiconductor, Dr Elliott devised a novel infra-red detector (Patent No. 1488258) which radically simplified the design and construction of thermal imaging systems. Previous detectors exploited solely the increase in electron density consequent upon photon absorption and, to extract adequate information about the viewed scene, it was necessary to couple an array of detectors to complex circuitry; processing the signal by integration and amplification.

Dr Elliott realized that the physical transport of electrons along the detector, if synchronized with the motion of the image, could combine, in a single device, both the function of detection and the operations of signal processing. The concept, which was entirely original, could be proved only after intensive research on carrier transport and recombination in the chosen semiconductor, cadmium-mercury-telluride. At the outset few believed that the idea was feasible but, after two years' research, successful operation was demonstrated and the detector (known as a TED=Tom Elliott Detector) is now universally accepted. Its influence on systems design was dramatic and the novel concept resulted in improved performance and a large reduction in the cost, size and weight of the complete imager.*

The work of Dr Elliott has resulted in the United Kingdom gaining a considerable lead in the field of infra-red imaging which is achieving a growing importance in applications where, for example, seeing in darkness is of significance and in medicine, where the detailed examination of 'hot spots' associated with diseases is leading to better understanding and treatment of those diseases.

Previous prizes in the opto-electronics area have gone twice to RSRE, in 1976 for the development of cadmium-mercury telluride thermal imaging and in 1980 for the invention and development (in collaboration with the University of Hull) of a new range of liquid crystal materials. In 1978 the inventors of optical fibres for communication (Kao and Hockham of STL) received an award and in 1980 experimental techniques developed at Stanford University which led to a greater understanding of the mechanism by which the eye achieves its remarkable sensitivity area the subject of a prize. The Foundation has also supported a number of research projects in British universities by the granting of research fellowships, while notable symposia have been held on such subjects as 'Electronic Imaging' and 'Scanned Image Microscopy'; the proceedings of several of these have been subsequently published in book form.

One of the Trustees is Dr F. E. Jones (formerly Managing Director of Mullard), who is chairman of the Opto-Electronics Advisory Committee, the other members of which are Prof. E. A. Ash (University College London), Prof. H. B. Barlow (Cambridge University), Prof. D. J. Bradley (Trinity College, Dublin), Prof. C. Hilsum (RSRE), Mr D. H. Roberts (General Electric Company), and Dr P. Schagen (formerly Philips Research Laboratories).

*See Webb, D. B., 'Thermal imaging via cooled detectors', *The Radio and Electronic Engineer*, 52, No. 1, pp. 17–30, January 1982.

Members' Appointments

CORPORATE MEMBERS

P. V. Betts, M.Tech., Dip.E.E. (Member 1972, Graduate 1964) has been appointed Head of the Engineering Department at St Albans College. Mr Betts, Chairman of the Kent Section for the past three years, has been on the staff of the Medway and Maidstone College of Technology since 1967.

Cheung Leung Fat (Member 1980, Graduate 1960) who was Local Services Engineer with Cable and Wireless in Hong Kong, has taken up an appointment as Teaching Master at Seneca College of Applied Arts and Technology in Toronto.

T. E. Ivall (Member 1970), Editor of *Wireless World* since 1973 and a member of its staff since 1950, has left the magazine to pursue a freelance writing career.

Group Captain K. G. Lewis, RAF (Member 1968) has now relinquished command of the Central Region Signal Group at HQ AFCENT and has taken up the post of Officer Commanding Royal Air Force Station, Locking.

L. P. Luk, B.Sc.(Tech) (Member 1979, Graduate 1977) has been awarded a UWIST Research Studentship and has returned to Cardiff to pursue studies towards a Ph.D. in the Department of Marine Studies. Since graduating at UWIST in 1977, Mr Luk has been a Lecturer in the Department of Nautical Studies at Hong Kong Polytechnic.

S. R. Taylor, M.Sc. (Member 1971, Graduate 1964) has recently moved to Cairo as representative of the Cheadle Heath Division of Ferranti Computer Systems.

NON-CORPORATE MEMBERS

N. R. Glenn (Associate Member 1962) has taken up an appointment as Manager, Electro-mechanical Development, with Rush Hampton Industries in Longwood, Florida, who specialize on air pollution treatment. He has been working in Canada since 1966, prior to which he was with Plessey Radar in the Isle of Wight.

R. J. Harris (Associate 1971) who joined Cable and Wireless in 1956, is now Branch Engineer for the company in the British Virgin Islands. Prior to this appointment, Mr Harris spent two years as Senior Telecommunications Engineer with the newly formed Botswana Telecommunications Corporation and four months assisting with the design specifications for the offshore telecommunications of the Morecambe Bay Gas Field.

S. C. D. Ikejiaku (Associate Member 1979) who was formerly a Senior Protection Control and Metering Technologist with the National Electric Power Authority, Kainji Dam, Nigeria, is now the Principal Technologist/Assistant Chief Technical Officer as well as Head of the Power and Machines Laboratory section in the Electrical Engineering Department at Ahmadu Bello University, Zaria, Nigeria.

K. E. L. Richardson (Graduate 1966) is now a Principal Engineer with the Plessey Company in West Leigh, Hants. He was previously a Development Engineer with MEL Equipment Co, Crawley.

Flt Lt P. A. Wheeler, B.Sc., RAF (Graduate 1971) is now serving at RAF Germany as an avionics specialist responsible for Phantom and Buccaneer aircraft, his appointment being Air Engineer 1B1.

Lt K. P. White, B.Eng., RN (Graduate 1979) has been recently appointed to HMS *Antrim* as a Weapon Engineering Section Officer, on completion of a course at the Royal Naval Engineering College, Manadon.

Obituary

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

Joseph Avellino (Member 1970, Graduate 1968) died recently in Malta aged 48. Born in Malta where he received his general education, Joseph Avellino studied business administration at the University of San Francisco and physics and electronics at Harris College, Preston. In 1958 Mr Avellino joined Central Rediffusion Services in London as a Senior Technical Author and Lecturer and in 1967 he returned to Malta to take up an appointment with Rediffusion (Malta). He subsequently became Deputy General Manager of Malta Radio and Television and a Director of the Rediffusion Group in the Island. From 1975 up to the time of his death he was General Manager of Malta Radio and Television and from 1977–1980 he was Chairman of TivuMalta.

Mr Avellino was active in industrial, philanthropic and cultural organizations in Malta and in 1980 he was awarded a prize by the Italian Centre for International Culture and Arts.

John Victor Bray (Graduate 1970) of Portsmouth died on 13th September 1981, aged 55. After war service as a radio electrical artificer in the Royal Navy, Mr Bray worked first for Rentaset and then for Bell Punch in Portsmouth before joining the Ministry of Defence as a Production Inspector. Subsequently re-graded as a Professional and Technical Officer, he held appointments in Barrow-in-Furness, Leicester and Portsmouth.

Peter Frank Davies (Member 1973, Graduate 1968) of Harrogate died on 5th November 1981 aged 34. After completing a four year apprenticeship with the Marconi Company, Peter Davies was appointed Development Engineer with the Company. In 1969 he went to the Department of Human Biology and Anatomy at the University of Sheffield as Research Assistant and two years later joined Joyce, Loebl and Company of Gateshead as a Senior Electronics Engineer. During this period he submitted a successful thesis on electronic techniques applied in cytology to the University of Salford for which he was awarded the M.Sc. degree. At the end of 1972 he joined the Electronics Group in the Space Division of Hawker Siddeley Dynamics at Stevenage as Senior Engineer working on electronic control sub-systems. In 1974 he was appointed Research Officer in the Scientific Services Department in the North Eastern Region of the Central Electricity Board at Harrogate and four years later went to Fotherby Willis Electronics of Leeds as Senior Engineer. In 1978 he formed his own company of systems consultants, Inca Electronics.

Squadron Leader Kenneth Baker Pearse, RAF (Rtd) (Member 1944) of Sandford, Isle of Wight, died during the summer of 1981, aged 71.

Joseph Albert Purdie (Member 1963) of Hull died in January 1982 aged 50. Following National Service in the RAF as an Air Wireless Mechanic, Mr Purdie studied at Chelsea College of Science and Technology obtaining a B.Sc. special degree in physics in 1954. He then worked successively with the Telegraph Condenser Company, Acton, Rotax in Willesden and Bruce Peebles at Edinburgh where he became engineer in charge of magnetic amplifier design and Deputy Chief Engineer. From 1960 he was Head of the Industrial and Nuclear Systems Laboratory with Ultra Electronics, Acton, where he designed an astatic gyro supply unit for the Fokker *Friendship* aircraft and control equipment for various nuclear power station. In 1962 he contributed a paper on control of nuclear power reactors by means of magnetic amplifiers to the Institution's Journal.

In 1964 Mr Purdie joined the Engineering Department of Watford College of Technology as a Lecturer in Electronics, continuing to act as a consultant to Ultra Electronics and other companies. Since 1973 he had been Head of the Department of Electrical Engineering at Hull College of Technology.

Ronald Arthur Tyler (Member 1961, Student 1950) of Chelmsford died recently aged 53. Mr Tyler was with Marconi Research Laboratories.

Vernon Gearon Welsby (Fellow 1959) died on 19th December 1981 in Queen Elizabeth's Hospital, Birmingham, aged 66, after several months of illness.

Vernon Welsby graduated in mechanical engineering from the University College of South Wales, Cardiff, receiving the B.Sc.(Eng.) external degree of the University of London and a Diploma in Engineering of the University of Wales in 1934. In 1936 he joined the Engineering Department of the General Post Office as a Probationary Inspector and from 1944–1947 served with the Royal Signals, for two years with the rank of Major. In 1946 he was awarded an external Ph.D. degree by the University of London for a thesis based on work on carrier telephone development while at the Dollis Hill Research Station. On his return to the Post Office he was in charge as a Senior Executive Engineer of the team responsible for the mechanical design and overall electrical testing for the British part of the first transatlantic telephone system.

In 1957 Dr Welsby was appointed to a lectureship in the Department of Electrical Engineering at the University of Birmingham and he later became a Senior Lecturer and then Reader in Electrical Engineering. During these years he was an active member of the Department's group working on sonar; his particular specialization was concerned with the design of arrays and he contributed some half a dozen papers to the Institution's Journal in this general area. One of these, on 'Electronic sector scanning' of which he was a co-author, was awarded the Brabazon Premium for 1958. This work led some years later to the award of a D.Sc. by the University of London. He was also the author of a well known textbook on 'The Theory and Design of Inductance Coils'.

Vernon Welsby had taken part in Institution activities both locally and nationally. He was a member and Chairman of the West Midlands Section Committee and he served on the organizing committees for several conferences; he was to have chaired one of the sessions of last year's conference on Electronics for Ocean Technology at Birmingham a few days after he suffered a first heart attack.

James White (Member 1947) of Ashtead, Surrey, died in October 1981 aged 68.

'Jerry' White obtained his early qualifications at Manchester College of Technology and in the years before the War he worked for a number of companies in the Manchester area as a radio engineer. In 1940 he joined the Aeronautical Inspection Directorate and held appointments as Inspector and Chief Examiner at various electronics companies and at Headquarters. Most of his subsequent career in government service was spent in the Ministry of Supply and from 1968–1975 he was First Secretary, Defence Supply and Civil Air Attache at the British Embassy in Paris. He retired in 1975.

While at the Embassy in Paris, Jerry White was an active member of the Institution's Group in France.

* * *

Frank William Appleby of Bexleyheath, died on 23rd November 1981, aged 69. He was Publications Manager of the Institution from 1959 until his retirement in 1977. Frank Appleby's considerable expertise in printing techniques, gained over many years in that industry, was evident in the setting of high standards for the production of IERE journals and other publications.

Standard Frequency and Time Service

Communication from the National Physical Laboratory

Relative Phase Readings in Microseconds NPL—Station

(Readings at 1500 UTC)

NOVEMBER 1981	MSF 60 kHz	GBR 16 kHz	Droitwich 200 kHz	DECEMBER 1981	MSF 60 kHz	GBR 16 kHz	Droitwich 200 kHz
1	−4·9	34·6	77·7	1	−6·0	34·9	75·8
2	−4·9	34·5	77·6	2	−6·0	35·0	75·7
3	−4·8	34·5	77·5	3	−6·1	35·8	75·7
4	−4·9	34·5	77·4	4	−6·0	35·8	75·6
5	−5·0	34·8	77·2	5	−6·0	35·3	75·6
6	−5·2	34·8	77·1	6	−6·2	33·3	75·5
7	−5·2	35·8	77·1	7	−6·1	33·1	75·5
8	−5·3	35·1	77·1	8	−6·2	32·9	75·5
9	−5·4	35·8	77·1	9	−6·2	32·8	75·5
10	−5·4	34·7	77·0	10	−6·2	34·8	75·5
11	−5·5	35·3	76·9	11	−6·3	34·8	75·6
12	−5·6	34·8	76·9	12	−6·2	34·0	75·7
13	−5·6	34·6	76·8	13	−6·5	34·3	75·8
14	−5·7	34·3	76·6	14	−6·5	33·8	76·0
15	−5·6	34·9	76·6	15	−6·5	35·0	76·1
16	−5·6	33·3	76·5	16	−6·7	—	76·1
17	−5·6	34·8	76·4	17	−6·5	37·5	76·1
18	−5·7	35·0	76·5	18	−6·5	37·4	76·1
19	−5·7	34·8	76·5	19	−6·5	37·0	76·1
20	−5·7	35·2	76·3	20	−6·7	36·5	76·2
21	−5·6	35·4	76·2	21	−6·7	36·5	76·3
22	−5·7	35·1	76·2	22	−6·7	35·8	76·2
23	−5·5	35·0	76·1	23	−6·7	37·5	76·1
24	−5·5	34·3	76·0	24	−6·5	37·5	76·0
25	−5·7	31·8	75·9	25	−6·6	37·5	75·9
26	−5·8	35·3	75·9	26	−6·5	37·0	76·0
27	−5·8	32·8	76·0	27	−6·7	36·5	75·9
28	−5·8	35·3	75·8	28	−6·7	37·3	75·9
29	−6·0	35·8	–	29	−6·7	34·5	75·8
30	−6·2	34·8	75·8	30	−6·9	36·0	75·8
				31	−6·9	36·8	75·7

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

Contributors to this issue

Geoffrey Gardiner joined the Radio Research Station at Ditton Park, Slough in 1955 after some years with the Meteorological Office at Kew Observatory. He was on the staff of Stanley Ionospheric Observatory, Falkland Islands, during the International Geophysical Year 1957–1959. He then returned to Ditton Park as a member of the Tropospheric Propagation Group concerned with balloon-borne experiments in the lower atmosphere. Latterly he has been responsible for technical training and information at Ditton Park and has gathered together apparatus and archival material dating back to the founding of the Radio Research Station.

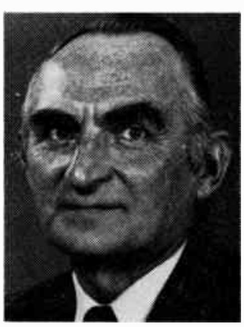
John Lane joined the staff of the Radio Division of the National Physical Laboratory during World War II, working on various problems in tropospheric propagation. Moving subsequently to the Radio Research Station, Slough, he made many contributions to microwave power measurement, radio meteorology, and ràdio communication at very high frequencies. A graduate of London University, he was awarded the D.Sc. degree of that university in 1971. He has been a member of several national and international committees in the general field of radio propagation, such as CCIR and URSI and with NATO. In 1962–63 he spent a period as guest worker in the USA. He is currently a consultant in the Directorate of Radio Technology, Home Office, and is national chairman of the CCIR Study Group 5 (Propagation in Non-Ionized Media).

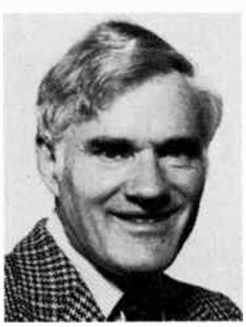
Henry Rishbeth received the degree of B.A. at the University of Cambridge in 1954 and carried out research in radio astronomy and ionospheric physics at Sydney and Cambridge before joining the Radio Research Station at Slough in 1960. He was consultant at the National Bureau of Standards at Boulder, Colorado from 1962 to 1964 and, during a visit to Stanford University, began working in partnership with Owen K. Garriott on the textbook 'Introduction to Ionosphere Physics'. Rejoining the (then) Radio and Space Research Station—later Appleton Laboratory—in 1965, he continued his theoretical studies of the ionospheric F region, and has been involved with the European Incoherent Scatter project, EISCAT, as first Chairman of the Scientific Advisory Committee and as UK Project Scientist. He was Deputy Director of the Appleton Laboratory from 1977 until its merger with the Rutherford Laboratory in 1979. He has been a Senior Visiting Fellow at the University of Southampton since 1981. He received from the University of Cambridge the degrees of Ph.D. in 1960 and Sc.D. in 1972, and in 1981 was awarded the Appleton Prize for Ionospheric Physics by the Royal Society of London and the International Union of Radio Science.

Larry Lind received the B.Sc. degree in electrical engineering from Virginia Polytechnic Institute in 1962, the M.Sc. degree from the University of New Mexico in 1965, and his Ph.D. degree from Leeds University in 1968. He has had industrial experience at Sandia Corporation, Albuquerque, New Mexico (1962) and the Research Laboratories of the Australian Post Office, Melbourne (1974). In 1968 he joined the Department of Electrical Engineering Science at the University of Essex. Since that time he has lectured in a number of areas, including circuit theory, field theory, communication theory, and logic design. Dr Lind is the author of over 25 papers, including one in the Journal in 1978, and is also co-author of the book, 'Analysis and Design of Sequential Digital Systems'. Since 1970 he has acted as a consultant for Trend Communications. His current research interests include circuit theory and communication theory.

Augusto de Albuquerque received the Electrical Engineer degree from the Lisbon Technical University in 1972 and the M.E.E. degree from the Netherlands Universities Foundation for International Co-operation (NUFFIC) at the Hague in 1977. In 1972 he was appointed a lecturer at the Lisbon Technical University; from 1973 to 1975 during his military service he was the adjoint to the Head of the Communications Department in the Portuguese Navy. At present he is working on his Ph.D. thesis at Essex University. His current interests span the areas of circuit theory, communication theory and digital signal processing.


G. Gardiner


J. Lane


H. Rishbeth


L. F. Lind


A. de Albuquerque

Biographies of other authors are given on pages 144 and 154.

General Interest Article

Radio and Space Research at Slough 1920–1981

G. W. GARDINER*, **J. A. LANE**, M.Sc., D.Sc., F Inst P., C.Eng., FIEE*

and

H. RISHBETH, Ph.D., Sc.D.*

1981 saw the end of 61 years of radio science at Ditton Park, Slough—apart from routine ionospheric sounding, which started in 1931 and will continue. The Radio Research Station, formally constituted in 1927, was renamed Radio and Space Research Station in 1965 and Appleton Laboratory in 1973. Following the merger with the Rutherford Laboratory in 1979, the Appleton Laboratory's work has been transferred, mostly to the Rutherford Laboratory site in Oxfordshire. This article describes subject by subject, rather than strictly chronologically, the major trends of the work at Ditton Park since 1920.

1 The beginning

For over 60 years Ditton Park has been the scene of investigations into problems of radio propagation and upper atmosphere research, dating back to the formation in 1920 of the Radio Research Board of the Department of Scientific and Industrial Research, in order to 'direct any research of a fundamental nature that may be required, and any investigation having a civilian as well as a military interest'. Among the people composing the first Board were Lord Rutherford and Admiral of the Fleet Sir Henry Jackson, the Board's first Chairman, who was perhaps mainly responsible for the Station's being sited at Ditton Park. Admiral Jackson was an enthusiastic and successful experimenter in wireless signalling; he was outstripped by Marconi and probably hindered by the need to observe security regulations, but his interest in the subject never flagged. Among those on the four sub-committees dealing with propagation, atmospherics, direction-finding and thermionic valves were E. V. Appleton, R. L. Smith-Rose and R. A. Watson-Watt.

In 1920 the Ditton Park site was manned by one scientist, R. H. Barfield and an assistant working in collaboration with the National Physical Laboratory. A year later a translator with scientific qualifications joined them to produce abstracts for monthly circulation. The work was mainly concerned with studies of field strength measurements, methods of screening apparatus from electromagnetic fields and radio direction-finding.

At the end of the first World War responsibility for the Aldershot Wireless Station of the Meteorological Office had been transferred to the D.S.I.R. Under the supervision of Watson-Watt, this station was equipped for the study of atmospherics as part of the Radio Research Board's programme. When, some years later, the War Department wished to re-occupy the Aldershot site the huts, equipment and Watson-Watt moved to Ditton Park, the move being completed in 1924.

As from 1st December 1927, under the Superintendency of Watson-Watt, the direction-finding, field strength and upper atmosphere research were combined in the new Radio Research Station (subsequently amalgamated in 1933 with the Wireless Department of the National Physical Laboratory). Also on 1st December 1927, the Station suffered a disastrous fire which destroyed the 210 ft lattice tower and some of the surrounding buildings. Nevertheless, wooden buildings in Ditton Park remained as the Station's home until the 'New Building' was occupied in 1956, and part of the 'Old Buildings' remained in use until 1981. By a quirk of history, the administrative obsequies that preceded the handing-over of the site to a commercial firm, early in 1982, were conducted from the 'Old Buildings'.

2 Early Ionospheric Research

The years 1925–27 were of great importance to the science of Geophysics because it was in that period that Appleton and his co-workers proved the existence of a reflecting layer at a height of about 100 km, soon to be followed by the discovery of a further layer at a height of some 250 km. These layers (now known as E and F) of partly ionized gas form a 'mirror', reflecting radio waves (v.l.f. to h.f.) to great distances around the Earth. Scientists at Ditton Park had provided a great deal of the substantiating evidence relating to the existence of the ionosphere which was to occupy the attention of research workers for many years to come. The term 'ionosphere' appears to have originated in 1926,[1,2] being used in letters written by Appleton and Watson-Watt in November of that year. The experimental measurements of reflection heights and penetration frequencies of the ionized layers gradually developed into an observational routine. The 'frequency change' method, much as used in the historic ionospheric experiments of E. V. Appleton and M. A. F. Barnett in 1924, was employed with a transmitter in Windsor Great Park and receiver in Ditton Park. Data on reflection heights exist for 1930, and at noon on 11th January 1931 there began the long series of ionospheric critical frequency measurements.[3] (The critical or penetration frequency of an ionospheric layer is proportional to the square root of the peak electron density within that layer.) Noon values of E layer critical frequency were obtained on 90 days in

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Fig. 1. The 'Old Buildings' of the Radio Research Station at Ditton Park, Slough, with the Director, Dr R. L. Smith-Rose. (*Photograph by Harold White, Hon. FRPS, FIIP.*)

1931 and there were a few 24-hour runs of hourly values that year, all at weekends; the almost unbroken daily sequence of E and F region data started on 20th September 1932. The pulse sounding technique, first used in 1925 by G. Breit and M. A. Tuve in the USA, superseded the frequency change method; an early swept-frequency ionogram obtained by that technique is shown in Fig. 4. During the Second International Polar Year (1932–33), a party from Ditton Park operated equipment from Tromsø and a loan of equipment was also made to a future Director, J. A. Ratcliffe of Cambridge.

Analysis of the large amount of systematic sounding data led E. V. Appleton and various co-workers to important conclusions about the ionosphere. The data suggested that solar ultra-violet light accounts for the production of the normal ionospheric layers and for their daily, seasonal and solar-cycle variations. However, charged particles guided by the earth's magnetic field produce marked effects at high latitude stations, such as Tromsø. The anomalous seasonal variations of the F and D regions at Slough and elsewhere, and the perturbations that accompany magnetic disturbances, remained as puzzles that were solved in general terms only in the 1960s and 1970s, work in which the Ditton Park scientists of the day played their part.

The outbreak of war in 1939 accelerated the evolution of the Ionospheric Forecasting Service to supply predictions of parameters to aid long-distance communications. Wartime work at Ditton Park also produced valuable information on radio direction finding, navigational aids and the locating of thunderstorms. Perhaps the Station's greatest wartime contribution was the Adcock direction-finder, based on previous work at RRS.

3 The Invention of Radar

In January 1935, Watson-Watt was approached by the Air Ministry to investigate the possibility of radiating energy at sufficient flux density to cause damage to an aircraft or its occupants. Although the impracticability of this suggestion was easily demonstrated, Watson-Watt reported to the Air Ministry that 'Meanwhile, attention is being turned to the still difficult, but less unpromising problem of radio detection ... and numerical considerations on the method of detection by reflected radio waves will be submitted when required'. Together with A. F. Wilkins, Watson-Watt calculated the amount of energy capable of being reflected from an aircraft when it was illuminated by realizable transmitter powers. The answers gave them grounds for supposing that a detection scheme was practicable even if the results were an order of magnitude smaller than predicted. Consequently on 12th February 1935 Watson-Watt prepared a draft memorandum entitled 'Detection and Location of Aircraft by Radio Methods'.

This communication has been called one of the most prophetic scientific documents ever produced. It stated the case for detection by reflected radio energy, showed the importance of pulse techniques in determining distance, and proposed the use of rotating beams to provide a system showing range and direction on a cathode-ray oscilloscope display at a single station. The eventual desirability of using shorter wavelengths and possible means of distinguishing between friend and foe were also considered.

An *ad hoc* experiment was arranged for 26th February. Apparatus from Ditton Park was positioned near the BBC's 50 m transmitter at Daventry, which provided the energy to illuminate an aircraft on a pre-arranged course along the axis of the transmitted beam. At 0945 hrs the

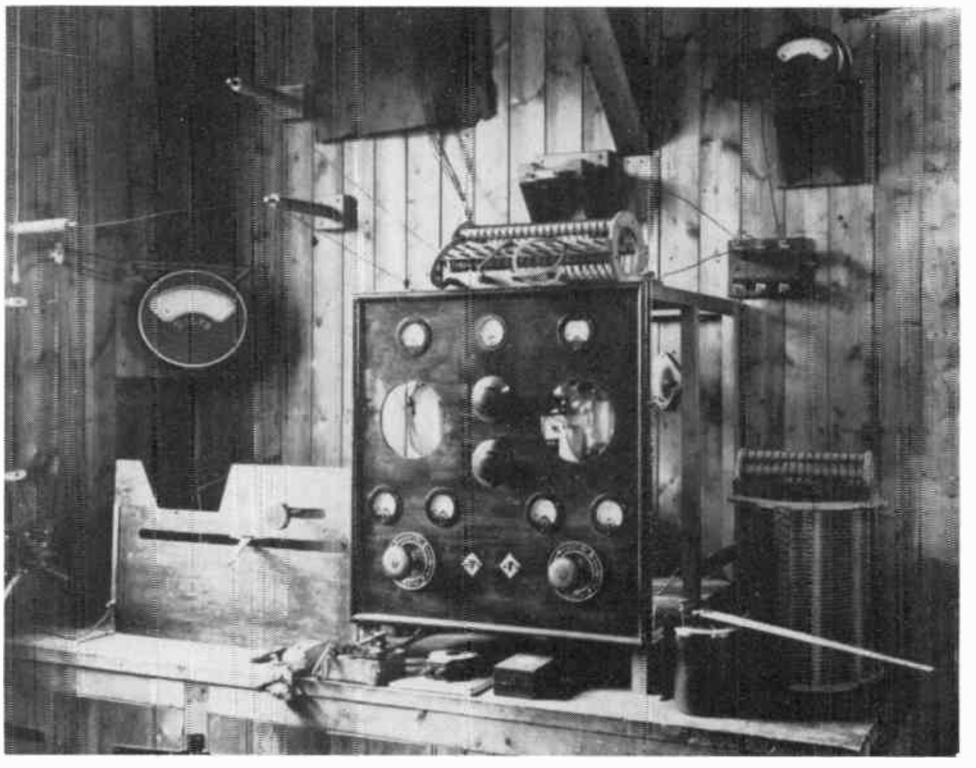

Fig. 2. Frequency-change transmitter, operated in Windsor Great Park for the first routine ionospheric sounding in 1931.

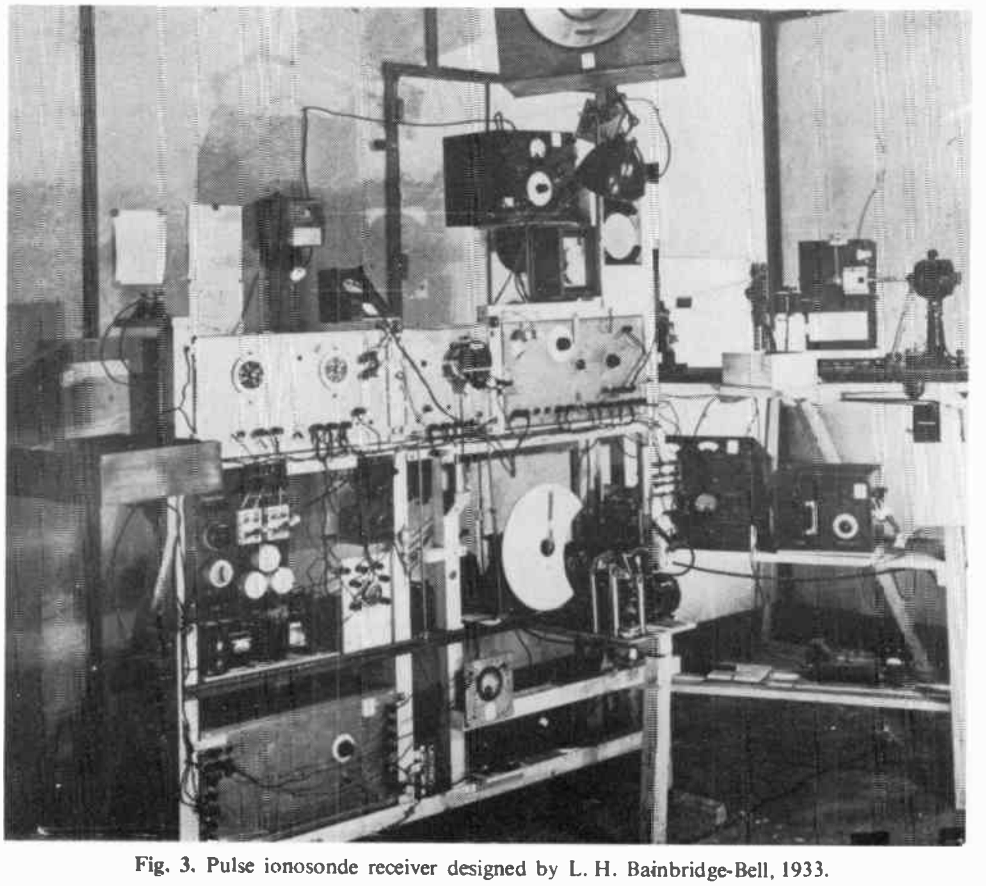

Fig. 3. Pulse ionosonde receiver designed by L. H. Bainbridge-Bell, 1933.

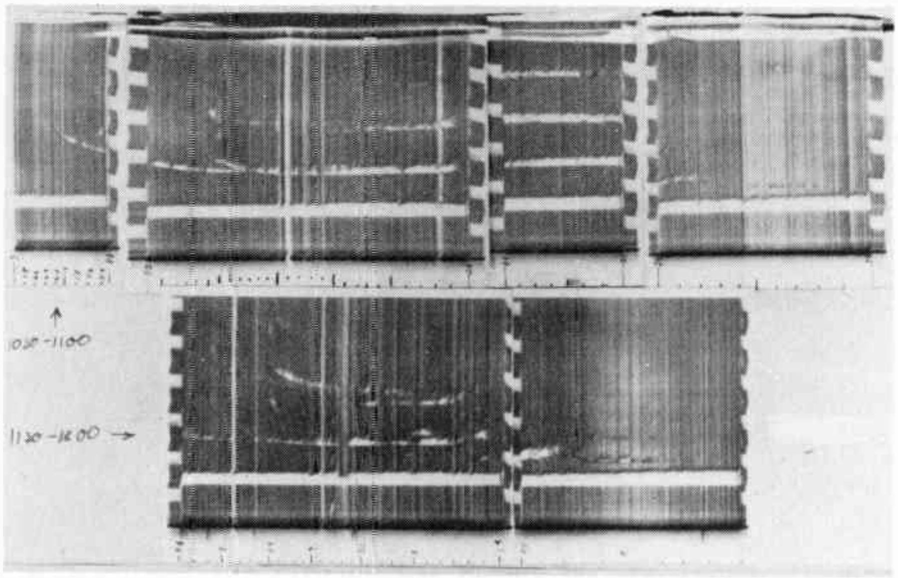

Fig. 4. The first swept-frequency ionograms obtained at Slough, 27th December 1933 at 1030-1100 and 1130-1200 UT.

aircraft duly appeared, not quite on course, but near enough to reflect detectable energy into the receiver and a detection range of eight miles was estimated.[4]

As commented subsequently by Wilkins, 'Considering the crude nature of the apparatus and the lack of preparation, the results obtained were quite creditable.... It was clear to all who watched the tube on that occasion that we were at the beginning of great developments in the art of air defence'. The original apparatus used in that experiment was rescued from a store hut in Ditton Park and presented to the Science Museum in 1958.

An immediate result of the experiment was a strict security blanket and, in May 1935, several RRS staff moved to Orfordness in East Suffolk, and later to Bawdsey where they formed the nucleus from which grew the vast complex of radar. By September 1935 these staff had demonstrated in operation all the necessary elements of a radar system, including measurements of range, azimuth and elevation and estimates of numbers of aircraft; and had been set the task of providing five radar stations for the defence of the Thames estuary, some receivers for which were made at Slough.

4 Other Work prior to 1946

Much work had been carried out (both at Slough and Teddington) prior to 1939 on topics such as metre-wave propagation, field strength measurement and direction-finding. However, the development of radio techniques in World War II and the subsequent exploitation of radio communication generally led to a considerable expansion in the research programme after 1939.

The study of the origin and nature of atmospherics and the consequent variation in the r.f. noise level, as a function of time and space, was extended to meet wartime needs. In addition, direction-finding techniques were improved and applied in meteorology (e.g. to the study of wind velocities) and in several specific operational problems posed by the armed Services. At centimetric wavelengths, the developments in radar led to a range of propagation studies from 1940 onwards. In particular, co-operative projects with the Ministry of Supply were initiated to investigate the effect of the lower atmosphere on transmission in the 3 and 10 cm bands (10 and 3 GHz), especially as regards the characteristics of anomalous propagation associated with unusual variations in the vertical structure of radio refractive index. The effect of obstacles, such as buildings, hills and trees, was also studied in the centimetric band. Supporting work was also carried out on methods of measuring power at centimetric wavelengths, an important requirement in assessing system performance. All of this work provided a basis for the many developments in the post-war programme, especially as regards the exploitation of frequencies above about 30 MHz.

5 Post-war Directions

After the war it was time to consider a revised programme of research more suited to the needs of the future. In 1946 a report was submitted by the Radio Research Board to the Council for Scientific and Industrial Research. It was implicit in this report that the direct connection which had long existed between the Radio Research Station and the National Physical Laboratory should cease, and the Radio Research Organization was formed in 1948 with R. L. Smith-Rose as its first Director. Until 1956, part of the work was carried out at the NPL Teddington site, though the present article gives more emphasis to the Slough-based work.

A number of outstations had already been set up and now more followed. The ionosonde station at Burghead, which operated from 1941 to 1947, was replaced by one near Fraserburgh, subsequently moved in 1948 to Inverness where it operated until 1963. In the Southern

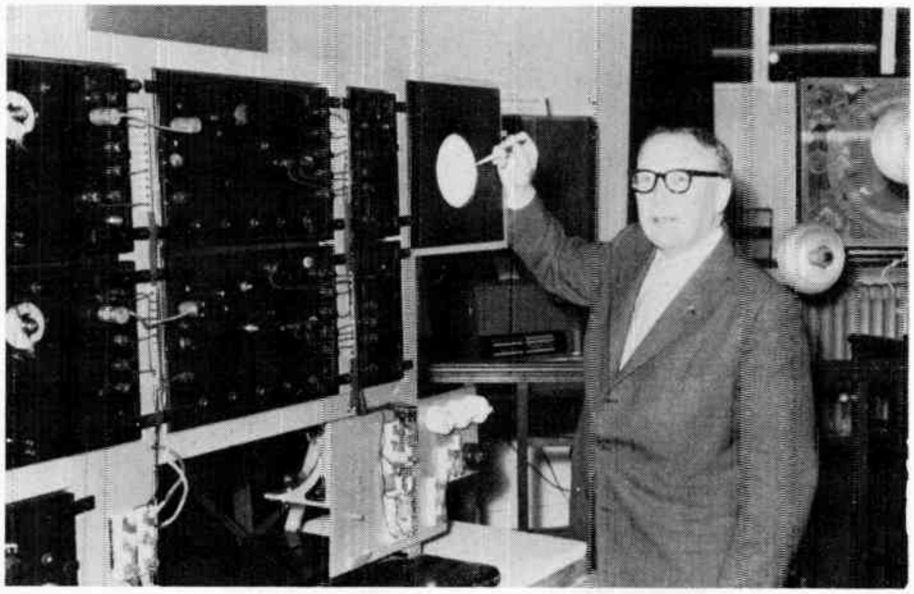

Fig. 5. Sir Robert Watson-Watt, FRS, First Superintendent, R.R.S., with the original British radar apparatus of 1935 made at Ditton Park (*Courtesy of The Science Museum.*)

Fig. 6. On 10th April 1981, A. F. Wilkins O.B.E. (*right*) unveils a plaque in the Old Building at Ditton Park, commemorating the pioneer radar experiment carried out by himself and Sir Robert Watson-Watt in February 1935. Professor J. T. Houghton, Director, looks on.

Hemisphere at (Port) Stanley, Falkland Islands, RRS personnel took over in 1947 an ionosonde station previously operated from 1945 by the Royal Navy. In 1948, a similar installation commenced operation in Singapore for measurements in the equatorial region, and continued until 1971. Also in 1948, RRS apparatus at Port Lockroy in Grahamland provided the first ionospheric observations ever to be made in Antarctica. Other sites, mostly in the vicinity of Slough, were used from time to time.

The post-war developments in radio communication affected the programme in several important ways; in ground-wave propagation, radio noise, tropospheric studies, materials and measuring techniques. Studies of phase changes in l.f. propagation at a land-sea boundary were carried out to assist the development of radio aids for marine and air navigation. Measurements of ground conductivity were also made. In the 2–20 MHz band, a world survey of noise levels was made in the period 1946–57, coupled with further work on the waveform and location of atmospherics. The work was subsequently extended to provide data during the IGY (1957–58), notably by an extensive measurement programme at Slough, Singapore, and Ibadan (Nigeria).

To provide a basis for planning v.h.f. and u.h.f. communications, including broadcasting and television, a long-term study was initiated of the effect on propagation due to both atmosphere and terrain. The early work, collaborative with the BBC and Post Office, was in the band 40–60 MHz and formed an important contribution to national and international planning. The propagation studies were gradually extended upwards in frequency to about 1 GHz during the period 1946–55. It then became evident that further progress, especially in beyond-the-horizon propagation, required more detailed knowledge of refractive-index structure. This requirement brought the Station into the field of tropospheric meteorology, the work being greatly assisted by the development in the period 1955–61 of ground-based radar studies of the layer structure of the troposphere and the use of airborne refractometers for direct measurements of such structure, simultaneously with the radar observations. In tropospheric and ionospheric fields alike, the Station's scientists have contributed greatly to the work of the International Radio Consultative Committee (CCIR), a contribution which continues to this day.

6 The International Geophysical Year

The increasing diversity of the work being handled by the Station accelerated the decision to build a new laboratory capable of combining the facilities previously supplied by the National Physical Laboratory with those existing at Ditton Park. A start was made on the New Building at Ditton Park in 1954. It was occupied in 1956, enabling the remaining Teddington-based work to be finally removed to Ditton Park, and inaugurated with due ceremony on 20th June 1957, the man invited to perform the ceremony being Sir Edward Appleton who had long been associated with the Station and had made many great contributions to ionospheric research (Fig. 7).

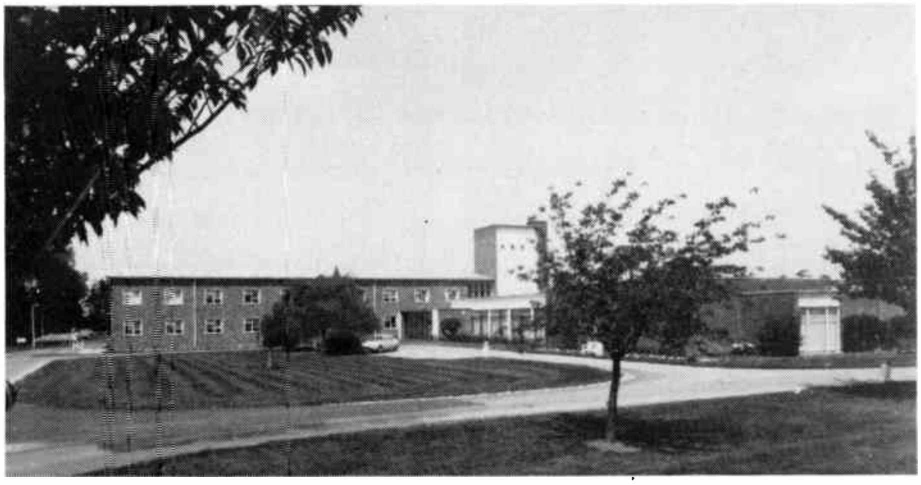

Fig. 7. The Main Building of the Radio Research Station (renamed Radio and Space Research Station in 1965 and Appleton Laboratory in 1973), opened by Sir Edward Appleton in 1957.

A range of measurements on semiconductors and ferrites was carried out in the period 1953–60, partly in collaboration with Imperial College, London. The work included studies of noise in germanium and silicon diodes, the behaviour of transistors in the band 10–1000 MHz, and the measurement of the microwave magnetic properties of a series of ferrites. In the same period, work at Teddington and subsequently at Slough, on comparison of microwave power measurements led to the development of improved techniques, especially in the use of water calorimeters and resistive-film bolometers. These methods were used in the development of improved national and international standards in microwave measurements.

The task of producing the Radio Research Board *Abstracts and References*, which began at Ditton Park in the early twenties, continued throughout this era. An arrangement by which the abstracts were reproduced in a professional Journal began with the publication of Vol. 3 of RRB *Abstracts* in *Experimental Wireless and The Wireless Engineer*, 1926, and the series continued to Vol. 39 which appeared in *Electronic Technology*, 1962. In later years about 4000 articles were abstracted annually (and over 100000 in all), providing to radio scientists and engineers a valuable synopsis of the vast literature of their field.

To this literature the Station has itself contributed by way of articles, reports, conference papers and some books. Apart from routine literature such as ionospheric bulletins, the publications—in later years including some from Culham (Sect. 10)—were processed through a series of 'A-files', which started at A.1 with W. Ross's article on 'Direction of arrival of ionospheric radio waves' in 1950 and ended at A. 1465 with the present article in 1981.

The International Polar Year of 1882–83 had been followed half a century later by a similar venture in 1932–33 in which, as already mentioned, staff and apparatus from RRS played a significant part. These two events had taken place at periods of minimum sunspot activity; now, after a period of twenty-five years, an International Geophysical Year was planned. This was to be a programme of observation and experiment not confined to the polar regions, but extending over the whole globe during a period of sunspot maximum. Within two weeks of being officially opened, the new Station was committed to play a most important part in the enterprise. One of the four World Data Centres for the collection and exchange of ionospheric information was established at Slough and the activities of RRS now ranged from Singapore and Nigeria to South America and the Antarctic.

The IGY began on 1st July 1957; its initial 18-month programme was extended throughout 1959 and the world-wide collaboration in upper atmosphere observations has continued to this day. Through its observational programmes and the World Data Centre, the Station played its part not only in the IGY, but in the International Quiet Sun Year of 1964–65, International Active Sun Year of 1968–69 and other programmes.

Although the ground-based ionospheric work was to continue at Ditton Park for a further quarter century, the IGY marked a new era. On 4th October 1957 the first artificial Earth satellite, *Sputnik I*, began to orbit the Earth, so providing a completely new tool for investigating the Earth's environment. At Ditton Park it was an opportunity for a quickly contrived experiment using a cathode-ray direction finder, a device invented by RRS. From this time on, the Station's future was largely bound up with space science.

7 Ionospheric Research in the Space Era

The Station's contribution to ionospheric research has not been confined to the operation of ionosondes. To the benefit of innumerable users of radio communications via the ionosphere, techniques for predicting propagation conditions have continued to be developed over the years. As regards the basic science the post-war period, no less than the pre-war, saw a sustained programme of analysis and theoretical interpretation of the daily, seasonal, solar cycle and other variations of the ionospheric D, E and F regions. Attention was paid to the opportunities presented by solar eclipses for studying ionospheric parameters, though it became increasingly clear that—interesting as the phenomena may be—it is seldom possible to draw unambiguous conclusions about the processes operating. Under the leadership of J. A. Ratcliffe, a University scientist from the Cavendish Laboratory, Cambridge, who succeeded R. L. Smith-Rose as Director in 1960, the Station made notable contributions to the theory of the ionospheric regions, particularly in such topics as the chemistry of the lower ionosphere, winds and electric fields in the upper ionosphere, small-scale irregularities in the ionosphere and their effect on radio propagation. The last named topic was supported by extensive experimental work using transmissions from satellites. The advent of the Canadian and US topside sounder satellites of the *Alouette*, *Topsi* and *ISIS* series, launched from 1962 to 1971, brought new research opportunities to the Station, which also played a major role in the satellites' ground support and data handling. It became possible to use the ionosphere as a plasma physics laboratory, in which the satellite transmitters could in effect be used for active experiments in the ionospheric plasma. Furthermore, the global coverage afforded by the satellites revealed new aspects of the structure of the ionosphere.

The use of the ionosphere as a laboratory was further developed in the Station's rocket programme of the period around 1963–64. Particular objectives were the study of the r.f. impedance of antennas within the ionosphere, and of the 'resonance rectification' phenomenon in a plasma, originally reported by Japanese scientists, which occurs when the potential applied to a probe is modulated at a (radio) frequency near the 'plasma frequency' of the ambient plasma. Laboratory work was carried out to complement the rocket-borne experiments.

From the mid-1960s the Station's rocket programmes developed in several ways, increasingly in collaboration with university groups. First, propagation experiments using transmissions between rockets and the ground were used to study the distribution of ionization, particularly in the lower ionosphere where conventional ionosondes cannot produce useful results. Second, electric currents in the E layer of the ionosphere were measured in a programme using rocket-borne magnetometers, launched from Sardinia, Woomera (Australia) and Thumba (India); the work was particularly aimed at studying the electrical structure of sporadic E layers, the intense currents in the 'equatorial electrojet', and the current distributions during ionospheric disturbances. Third, the complex chemistry of the lower ionosphere has been investigated, most notably by radiometric techniques using rocket-borne ultra-violet lamps, by which the atomic oxygen distribution in the D and E regions was measured at various times and places from 1974 onwards. The importance of this work lies in the fact that the atomic oxygen distribution plays a key role in the ionospheric chemistry, but was virtually impossible to measure before the advent of the ultraviolet technique. Fourth, measurements of the intensity and energy spectra of the energetic particles that produce the aurora borealis have been made in numerous campaigns from 1967 to 1980. These experiments, carried out by *Skylark*, *Petrel* or *Fulmar* rockets—sometimes launched only after numerous countdowns terminated by adverse scientific, technical, meteorological or other circumstances—into varied types of auroral display, have contributed much to knowledge of a complex and fascinating phenomenon. Still at high latitudes, but at the other end of the earth; the long connection between the Station (or Laboratory) and the Antarctic has been maintained, despite the move of the British Antarctic Survey Ionospheric team to the Survey's new Cambridge HQ in 1976. During 1979/80 the Survey's new computer-controlled ionosonde was commissioned at Ditton Park before being shipped to Halley Bay in late 1980. This instrument—the most advanced of its breed seen at Ditton Park—had the privilege of recording Slough's highest ever F2 critical frequency (17·2 MHz) on the afternoon of 15th December 1979.

After 1981, the ionosonde became the sole remaining ionospheric presence in Ditton Park. It is a modern digital sounder, intended for virtually unstaffed operation and capable of transmitting over a landline to Chilton, from whence a dial-up service of current and past ionospheric information will be provided for users. Elsewhere, the Laboratory's ionosondes at Stanley, Falkland Islands, and at South Uist continue to operate into the 1980s.

8 Tropospheric Research in the Space Era

Although the Station continued to maintain its expertise in wave propagation over the whole radio spectrum, work after about 1960 became increasingly influenced by microwave developments for satellite communications and terrestrial applications. More detailed studies of refractive-index structure were carried out, using refractometers suspended from helicopters, or carried by balloons at Cardington (Beds.). The results were applied in various ways; for example, to estimate field-strength variability in propagation beyond the horizon and in calculating scintillation effects on millimetre-wave line-of-sight systems. Measurements of fading (both for clear air and rainfall) were made on a 55 km line-of-sight path and the data interpreted in terms of performance of terrestrial radio-relay systems. Later work on this topic was carried out jointly with the Post Office, using frequencies in the range 11–37 GHz. A network of microwave links and rapid-response rain gauges was used in Suffolk from 1973 to 1975, to establish a relationship between rainfall structure and fading and a generalized prediction procedure was developed. Further work extended the measurements to frequencies of the order of 100 GHz, where the effect of heavy rain is serious. Microwave scattering from tall buildings was also studied in a series of experiments conducted in 1970, in collaboration with the Post Office, using pulsed signals at 9·4 GHz. The results were used to estimate interference effects on microwave links due to multipath propagation via buildings.

Prior to the availability of centimetre-wave transmission from satellites, an extensive programme of attenuation measurements was carried out from 1968 onwards, using radiometers in either a solar-tracking or a passive sky-emission mode. Starting at frequencies of 11 to 19 GHz, the work was later extended to 37 and 70 GHz. Data were also obtained on 'site diversity', the technique of using alternative terrestrial sites to improve reliability on Earth–space links. Rainfall structure was also studied using a 10 cm radar mounted on the 25 m steerable antenna at Chilbolton (Sect. 9); from 1978 this work used a novel dual-polarization technique which in effect measures the average oblateness of raindrops, which depends on their average size, and which enables rain and ice within clouds to be distinguished.

The physics of atmospheric absorption at frequencies above about 30 GHz was studied by a variety of techniques, including ground-based radiometers, direct measurements on short paths (extended later to infra-red and optical wavelengths), and laboratory measurements using cavities and absorption cells. The work established that there exists a component of absorption in clear air which is not explicable in terms of conventional spectroscopic theory. The detailed explanation remains to be established and a series of experiments to this effect was carried out at Ditton Park and elsewhere. Partly to assist such millimetric work and partly to aid developments in radio astronomy and remote probing, work was initiated and still continues on developing low noise diodes at frequencies up to about 200 GHz. In particular, these devices have been used in a programme of microwave radiometry, in which was studied atmospheric absorption at frequencies up to 220 GHz; the techniques developed have provided a valuable basis for proposed future work on remote probing of the Earth's atmosphere and surface.

Following the availability of satellite transmissions, a range of experiments was carried out in collaboration with universities and Government Departments. The work at 30 GHz included attenuation measurements in conditions, and diversity studies using three spaced receiving sites. At 11 GHz, attenuation and polarization studies using transmissions from OTS were combined with phase dispersion and differential attenuation measurements over a 500 MHz bandwidth, using transmissions from *Sirio*. These programmes were carried out over the period 1975 to 1980.

In 1979 a microwave test range was brought into use at Ditton Park, on which several frequencies in the range 37–220 GHz can be used simultaneously for studies of atmospheric attentuation in clear air, fog, rain, snow etc. Supplementary measurements in the visible and infra-red bands can also be made. The equipment was transferred in 1981 to the Chilbolton site, where it will be operated on paths of up to 20 km, the 25 m radio telescope being available for co-ordinated experiments.

9 New Ground-based Activities

The growing importance of space activities was reflected in the new name 'Radio and Space Research Station' which was adopted when control of the Station passed from D.S.I.R. to the newly-formed Science Research Council on 1st April 1965.

The appointment of J. A. Saxton, an authority on tropospheric radio propagation, who had first come to NPL in 1938 and RRS in 1952 and now became Director in 1966, was appropriately followed by the commissioning of a new instrument capable of serving tropospheric research, as well as ionospheric science and astronomy. This was the fully steerable parabolic radio telescope at Chilbolton, Hants, formally opened on 14th April 1967, by Anthony Crosland, Secretary of State for Education and Science. The telescope is 25 m in diameter and is operable at wavelengths down to 9 mm with a pointing accuracy of ½ arc-minute. The site was chosen because it combined freedom from radio interference with convenience of travel to and from Slough. With this powerful instrument investigations have been undertaken into radio propagation problems in the lower atmosphere; studies of the internal structure of rain clouds; and ionospheric incoherent scatter experiments using a transmitter at the Royal Signals and Radar Establishment, Malvern, and receivers at Chilbolton, Aberystwyth and Jodrell Bank. Used in conjunction with Canadian and US radio telescopes as a transatlantic interferometer operating at 3 cm wavelength, the telescope was used to map the structure of quasars (quasi-stellar radio sources) with a resolution of 10^{-4} arc-second. It has also been used for a variety of university research programmes.

It was known from the 1930s that disturbances on the Sun, in particular solar flares, can affect the ionosphere and thereby disrupt radio communications. The effects include an immediate 'fade-out' caused by ultra-violet and X-rays from the flare, and further disturbances after a time lag corresponding to the travel time of charged particles from the sun to the earth. The detection of solar r.f. emissions by wartime military radars led to the post-war development of solar radio astronomy. Though the solar emissions are rarely strong enough themselves to cause r.f. interference, they serve as important diagnostics of solar disturbances. For this reason some measurements at metre wavelengths were made at Ditton Park in 1948. However, the Station's major contribution came subsequently from observations in the millimetric range. Regular measurements at a number of wavelengths between 4·2 and 107 mm (71–2·8 GHz) were conducted at different times, in an overall programme extending from 1966 to 1979; this work was associated with radiometric studies of the atmosphere, mentioned in Section 8.

Though not first in the field as regards laser measurements of air density, RRS pioneered the use of lasers for upper atmosphere research and introduced many new techniques. Early work, started in 1962, obtained profiles of air density up to heights of about 70 km. Later, when technological advances made it possible to produce efficient lasers operating at the wavelength of the sodium yellow D lines, detailed studies became possible of the distribution of sodium in the night-time upper atmosphere. During three of the High Latitude Rocket Campaigns, in 1973–77, sodium clouds released from rockets were tracked from the ground by the Laboratory's lasers, to measure winds at heights around 150 km. More recently, measurements of the hyperfine structure of the D lines provided a novel means of measuring atmospheric temperature at about 90 km, while dust clouds at around 17 km height were detected over the laser observatory at Winkfield some weeks after the Mount St Helens volcanic eruption in May 1980.

10 Space Science

Though reception of satellite transmissions continued at Ditton Park—in recent years at centimetric wavelengths, to study tropospheric effects on microwaves—the Winkfield outstation, a few miles from Slough, has carried out the bulk of the Station's satellite tracking and data acquisition activity. Constructed under an agreement with the US National Aeronautics and Space Administration and commissioned in 1960, the Winkfield station tracked, commanded or acquired data from about 200 000 passes of over 250 different satellites, until its closure in 1981. Winkfield was a key station in the NASA network and served, in its last two years, as the principal ground station for *Ariel 6*, control of which passed for the satellite's final months to the new control centre at Chilton.

During the IGY, with the launch of the first *Sputniks* in 1957 and numerous US satellites from 1958, the obvious need to generate and distribute information on satellite orbits quickly led to the development of special expertise for orbital predictions and calculations. Based on observational data from numerous professional and amateur sources using optical and radio techniques, in conjunction with well-developed computational methods, the Station gained national and international recognition for its work in this field. In the 23 years up to the eventual transfer of the orbital prediction service to the University of Aston in 1980, around 1·1 million observations of 3600 different satellites were received, of which 14 000 observations were made at Ditton Park itself; and 51 000 weekly predictions were issued for 1900 different satellites.

The decoding of information telemetered to the ground from satellites—or for that matter rockets and balloons—is an essential step in carrying out experiments in space. RRS became involved in reception and processing of telemetered data from *Ariel 2*, launched in 1964, and all further satellites in the *Ariel* series. In the case of *Ariel 5* and *6* the data, processed to yield actual scientific information, could reach a university experimenter within an hour of being acquired, sometimes enabling experimental plans to be modified before the next orbit of the satellite. Furthermore, the Station entered the field of image processing, a computational procedure intended (for example) to optimize and calibrate the output of satellite-borne cameras and to reduce noise and distortion in the digital images. Originally undertaken for *IUE* (International Ultraviolet Explorer), eventually launched in 1978, such work has also been carried out for other projects, notably *IRAS* (Infra-Red Astronomical Satellite) due to be launched in 1982.

With the growth of space science activities, the support of the British space programme became the major function of 'the Station', or 'the Laboratory' as it later became, the ceremony of renaming as Appleton Laboratory being performed on 7th November 1973 by the Secretary of State for Education and Science, Mrs Margaret Thatcher. Basically—in ways such as outlined above—the work entails the provision, through the Laboratory's own resources and through industrial contracts, of facilities for university and Laboratory scientists to carry out experiments using satellites, rockets, and high-altitude balloons. The Laboratory's own research programme has been increasingly integrated with those of universities. Besides the technological aspects of the work, responsibilities for project management were greatly enhanced by the transfer from London in 1972 of the SRC's Space Research Management Unit. The Astrophysics Research Unit of the Culham Laboratory became part of the Station in 1973, though remaining at Culham till its eventual move to Chilton in 1981. The expertise of the Culham team brought a new dimension, notably in solar physics, to the Station's space activities.

With its new responsibilities the Laboratory became involved in managing the UK scientific balloon programme, although Laboratory scientists have never yet themselves conducted experiments using these facilities. However, the Laboratory's expertise in h.f. radio communications was available for the tracking and telemetry of the transatlantic balloon flights of 1976. Unfortunately these flights ended disastrously; one balloon disappeared only a few hundred miles from the North American coast (the platform being washed up on the Cornish coast some months later) and the other payload was destroyed by fire after a successful landing and recovery in Kentucky. In recent years the Laboratory has been involved in the development of a stabilized balloon platform with arc-minute pointing accuracy, which underwent successful trials in Texas in 1979. Laboratory scientists and engineers have contributed to technological developments, notably as regards telemetry, both for balloon and rocket programmes.

Admiral of the Fleet Sir Henry Jackson, FRS, Chairman Radio Research Board, 1920.

R. L. Smith-Rose, C.B.E., Superintendent 1936–1948, Director 1948–1960.

J. A. Ratcliffe, C.B., C.B.E., FRS, Director 1960–1966.

Sir Edward Appleton, FRS, Secretary of Department of Scientific and Industrial Research 1939–1949, Nobel Laureate 1953.

J. A. Saxton, C.B.E., Director 1966–1977.

F. Horner, Director 1977–1979.

J. T. Houghton, FRS, Director 1979–

Research using rockets is dealt with elsewhere in this article but the Laboratory's management role in the UK programme, acquired in 1972, deserves mention here. The final UK *Skylarks* launched in 1978 well represent the science involved: one—launched in Australia—carried instruments of the Laboratory's Culham team for observing the Sun's corona; the other—launched in Norway after 43 countdowns—carried experiments to investigate auroral phenomena. These firings brought to an end the UK programme of 257 *Skylark* rockets which had started in 1957. As for other vehicles, six *Fulmar* rockets were launched in 1976–77, while the *Petrel* rocket programme—with 220 launches since 1967—continues, nowadays mainly from the South Uist range in the Hebrides with some launches from Kiruna.

11 Satellite Instrumentation

Construction of hardware for satellite experiments began with the provision of instruments for measuring terrestrial h.f. noise, carried by *Ariel 3* launched in 1967, and energetic particle detectors carried by *ESRO 1A* and *1B* launched in 1968–69. In later years some members of staff worked at University College London on the ultra-violet television camera systems which were the SRC contribution to the *IUE* satellite launched in 1978.

An exciting international satellite project was *Firewheel*, initiated by the Max-Planck-Institut at Munich with UK, US and Canadian participation. It was proposed to release, in the tail of the Earth's magnetosphere at a distance of several Earth radii, a barium cloud to create a magnetic cavity and subsequently to release a lithium cloud to create an enhancement of plasma density. The Laboratory had the task of designing and integrating one of the four *Firewheel* subsatellites. Successfully completed, the work came to nothing upon the failure of the *Ariane* launch vehicle in 1980, though a new project, AMPTE, was started in 1981. It should be pointed out, however, that the major satellite instrumental projects of the Appleton Laboratory have been carried out by the Astrophysics Research Division at Culham, rather than at Slough. The Culham group built up a worldwide reputation for experimental and theoretical astrophysical spectroscopy, particularly in solar work. The programme culminated in the construction and flight of the X-ray polychromator for the Solar Maximum Mission (launched 1980) and the Coronal Helium Abundance Spacelab Experiment developed for launch on a future Spacelab mission. Both at Culham and at Slough the satellite hardware projects have largely been collaborative with universities, in particular University College London.

12 The Merger and The Future

Following three years of discussion, the Science Research Council decided in 1978 to merge the Appleton Laboratory (with a total staff now exceeding 300) with the much larger Rutherford Laboratory. The rationale of the merger was that the transfer of Appleton Laboratory programmes to the Rutherford Laboratory site at Chilton in Oxfordshire would enable the radio and space research to draw on the engineering, technological and computing resources available at Chilton. The merger took effect on 1st September 1979 upon the retirement of F. Horner who, in the office of Director after J. A. Saxton's retirement in 1977, had had to steer the Laboratory through the difficult pre-merger period. An authority on the management of the r.f. spectrum and on atmospheric r.f. noise, he had joined NPL in 1941 and came to RRS in 1952.

For the second time, a University scientist took the post of Director, J. T. Houghton from Oxford becoming Director Appleton within the merged Laboratory, under the Director General, G. H. Stafford. With the new Director came plans to enter the field of remote sensing from satellites, along with further space projects which complemented other major Appleton Laboratory programmes now coming to maturity. These include control of the Infra-Red Astronomical Satellite, due for launch in 1982; the application of h.f. radar techniques to monitoring ocean waves and currents over a large part of the North Atlantic, using a site in Wiltshire; and participation in the EISCAT (European Incoherent Scatter) ionospheric radar project in northern Scandinavia. The pursuit of radio research of benefit to UK Government Departments and other national organizations has been consolidated into a programme overseen by the Departments. Co-operation with Universities plays an important part in the Rutherford Appleton Laboratory's programmes of the 1980s, which may confidently be expected to live up to the standards set by the Radio Research Station in the 1920s.

13 Acknowledgment

This article is dedicated to all the staff who, by their competent and devoted service in a wide range of crafts and occupations, have supported the scientific work at Slough and so made possible the achievements we have described. Grateful acknowledgment is made to all who, in discussion and correspondence, have provided information and background for this article. We apologize for any errors that we may have committed, and for the omissions that are inevitable in what must be a selective account.

14 References

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Indexing Terms: Electric filters

The design of optimal partial response data transmission filters

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and

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SUMMARY

The time response of a Class I partial-response filter to an impulse should be large at two adjacent sample points, and near-zero at all other sample points, in order to minimize intersymbol interference.

This paper gives an analytic design procedure for a realizable filter that meets this requirement in an optimal manner, and also optimizes other frequency and time domain criteria. Numerical examples are given, which illustrate the performance that can be attained with this class of filter.

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1 Introduction

Class I partial-response systems[1] are attractive for data transmission, in that they allow an increased bit rate to be used for a certain available bandwidth. They have various advantages and disadvantages, which are well documented.[2] It is interesting to note that the partial-response (p.r.) method is being used commercially, at present.[3,4] Examination of the literature reveals that the filters used in a p.r. system are normally conventional filters, such as Chebyshev filters.[2] The ripple level and degree of these filters are carefully chosen to give satisfactory selectivity. Although these filters are good, one has the intuitive feeling that they are probably not optimal for the waveshaping to be done. In particular, Chebyshev and many other conventional filters have been designed to give a specified *attenuation* (and not phase) response in the frequency domain, rather than a specified impulse response. Hence group delay equalizers are necessary. These equalizers create additional problems. They contain many elements, which implies cost and tuning problems. They give additional loss to the system, due to non-ideal components. And they are sensitive to small changes in element values, which can arise through drift and temperature variations.

The object of this paper is to give a design procedure for filters that are intended specifically for partial response systems. Every available degree of freedom is used to optimize their performance for partial response. This means that the intersymbol interference error is at a global minimum, with the implication that the frequency selectivity will be as sharp as possible. Both attenuation and group-delay performance are included in the one design, and so no separate group-delay equalizer is required. This feature leads to a substantial reduction in the number of elements, with the benefits of lower loss, lower cost, less tuning and lower sensitivity.

Previous work on the direct approximation of these filters has involved c.a.d. optimization procedures, and has only been moderately successful.[5] The reason is simple: for moderate-degree filters, hundreds of local minima exist, and are clustered about the global minimum. The situation becomes worse and worse as the degree of the filter increases. Here, sufficient analytic work is done to give a good *analytic* approximation to the final solution. Then a computer program is used to home in on the global optimum. At no time are extensive search procedures used. It turns out that the analytic approximations are sufficiently accurate to allow linearization of the equation set. The Jacobian matrix is found numerically, and inverted. Relatively few iterations are necessary to converge to the solution point. For example, the run time for the design and analysis of a 9th-degree filter takes only a few seconds.

2 Impulse Response

The time-domain impulse response for an ideal causal p.r. filter is shown in Fig. 1, where the arrows show the ideal sample instants. An important requirement is for the main lobe (third and fourth) samples to have the same value. The response at the infinity of other sample points should be zero, for zero intersymbol interference

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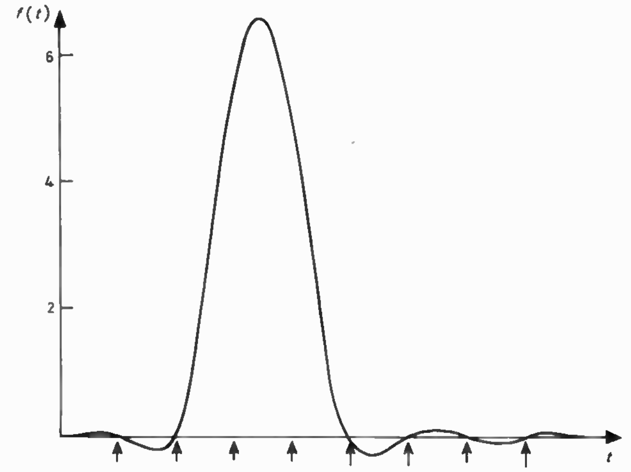

Fig. 1. The impulse of an ideal causal partial response filter. The ideal sample points are indicated by arrows.

(i.s.i.). Since a realizable filter has a finite number of elements (and hence degrees of freedom), it can never achieve this zero i.s.i. ideal behaviour. It can however *approximate* to the ideal in some optimal manner, over this infinite point set.

Recently a method has been found for the design of Nyquist I filters.[6] This method consists of two stages. The first stage is involved with the determination of all-pole responses which have the shapes of Fig. 2(a) and (b). The method allows the total postcursor squared error to be fixed to a preset value.

Even though each of these responses has the same postcursor i.s.i., the main lobe portions are different in shape, due to the specification of t_m, the main sample point. t_m is a 'floating variable' in the design, and can be positioned wherever the designer chooses to put it. For the Nyquist I filters it was found that for best results t_m should be adjusted so that the peak of the main lobe occurs at t_m, as in Fig. 2(a). The second stage of the design is concerned with adding transmission zeros to eliminate precursor i.s.i., as shown by the dotted line in Fig. 2(a).

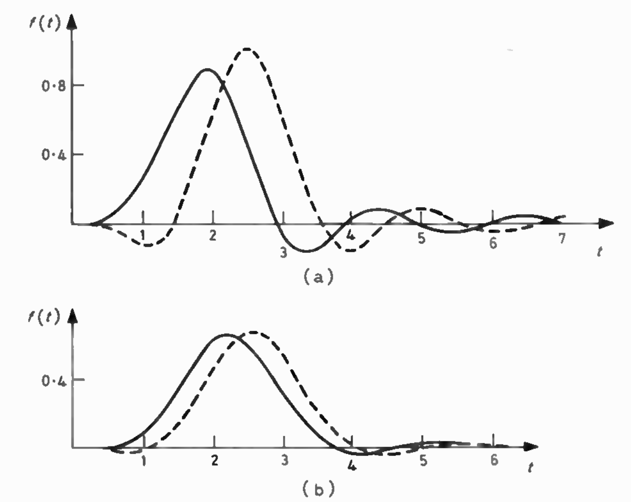

Fig. 2. All-pole (solid line) and precursor equalized (dotted line) impulse responses for 6th-degree filters. In (a), t_m has been adjusted for optimal Nyquist I performance. t_m has been shifted to the right in (b), to give optimal Nyquist II performance.

Referring to Figs. 1 and 2, it is plausible to believe that t_m could be shifted to the right as in Fig. 2(b), until the two main sample points t_m and $t_m - 1$ have equal-valued responses. Then transmission zeros could be introduced to eliminate the precursor i.s.i., just as before. The result would be the dotted line performance of Fig. 2(b). Some computer experiments showed that this was exactly the case.[10] This simple extension of the Nyquist I design provides the partial response filters of this paper. Even though this extension is theoretically simple, it leads to some profound differences in frequency domain performance, as will be seen.

3 Outline of the Design Method[6,7]

The frequency domain transfer function of a realizable filter can be written in partial fraction form as:

$$F(s) = \sum_{n=1}^{N} \frac{r_n}{s - p_n}$$

and has a corresponding impulse response

$$f(t) = \sum_{n=1}^{N} r_n \exp(p_n t).$$

The total squared *postcursor* i.s.i. is given by

$$g^2(t_m) = \sum_{k=1}^{\infty} f^2(t_m + k).$$

It is desirable to minimize the ratio of this error divided by the squared response at the main sample point, with respect to the residues. Let

$$E \triangleq \frac{g^2(t_m)}{f^2(t_m)}.$$

Then, to form a global optimum, we demand

$$\frac{\partial E}{\partial r_n} = 0, \quad n = 1, 2, \ldots, N.$$

Following the development in Ref. 6, we obtain the closed-form result

$$r_n = \frac{f(t_m)\exp(-p_n t_m)}{\sinh\left(\sum_{i=1}^{N} p_i\right)} [\sinh p_n] \prod_{\substack{i=1 \\ i \neq n}}^{N} \frac{\sinh(p_n + p_i)/2}{\sinh(p_n - p_i)/2}, \quad n = 1, 2, \ldots, N. \quad (1)$$

We also know that an all-pole filter has the residue-pole relation

$$r_n = \frac{\prod_{i=1}^{N} (-p_i)}{\prod_{\substack{i=1 \\ i \neq n}}^{N} (p_n - p_i)}, \quad n = 1, 2, \ldots, N. \quad (2)$$

Hence we can equate equations (1) and (2) to give

$$\exp(-p_n t_m)[\sinh p_n] \prod_{\substack{i=1 \\ i \neq n}}^{N} (p_n - p_i) \left[\frac{\sinh(p_n + p_i)/2}{\sinh(p_n - p_i)/2}\right]$$

$$= \frac{\sinh\left(\sum_{i=1}^{N} p_i\right)\prod_{i=1}^{N}(-p_i)}{f(t_m)} \qquad n = 1, 2, \ldots, N. \quad (3)$$

The equation set contains none of the r_n, but is an implicit set of relations for the poles. When pole values are found which satisfy this equation set, we have an optimal filter.

Equation set (3) is of prime importance, and is used in two distinctly different ways. Firstly, this set can be simplified to give an approximate solution in closed form, that is, good starting point locations for the poles.[1] Without these estimates, it is doubtful that the global optimum would be found for there are a large number of local minima in the error space. Secondly, equations (3) are used exactly in a computer program to iterate to the global optimum. Typically it has been found that a starting point location is within a few percent of its final value.

The remaining portions of the partial response computer program are concerned with the iterative adjustment of t_m and the transmission zeros. These adjustments have a small effect on the pole locations, and so, after each adjustment, the poles are relocated so as to satisfy equations (1). In this manner, a solution is arrived at which has the specified impulse response. The solution output consists of pole and zero locations, and t_m and $t_m - 1$, the main sample points. The input to the program is N, the degree of the filter, and E, the total squared post-cursor i.s.i.

The technique is capable of extension to non-impulsive inputs. Consider for example the rectangular input

$$u(t) = 1/D, \quad 0 < t < D$$
$$= 0, \quad \text{otherwise.}$$

The filter output $g(t)$ to $u(t)$ is found by convolving $u(t)$ with $f(t)$, which gives

$$g(t) = \sum_{n=1}^{N} \frac{r_n[1-\exp(-p_n D)]}{D p_n} \exp(p_n t), \quad D < t$$
$$= \sum_{n=1}^{N} a_n \exp(p_n t), \text{ otherwise.}$$

Therefore, since the a_n and p_n can be found by using the above method, the r_n can be found from coefficient comparison; that is,

$$r_n = \frac{D p_n a_n}{1-\exp(-p_n D)}, \quad n = 1, 2, \ldots, N.$$

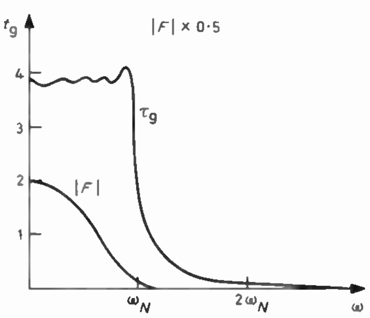

Fig. 3. Frequency response of the 9th-degree filter example.

4 Results

Using a PDP10 computer, 65 filters have been designed. For values of $N > 5$ it was necessary to use zeros for precursor i.s.i. cancellation. Both even and mixed numerator transfer functions were considered, allowing for passive and active implementations.

As numerical examples we consider two cases where $E = 10^{-4}$ and the input is an impulse:

(i) a 9th-degree filter with an even numerator,
(ii) an 8th-degree filter with a mixed numerator.

For the first filter the pole locations are at:

$-0{\cdot}351402590 \pm j\,2{\cdot}80786510$
$-0{\cdot}493529470 \pm j\,2{\cdot}14865070$
$-0{\cdot}562653940 \pm j\,1{\cdot}44007170$
$-0{\cdot}593432040 \pm j\,0{\cdot}721841800$
$-0{\cdot}603184080$

and the zero locations are at:

$\pm 2{\cdot}339814300 \pm j\,1{\cdot}008434400$
$13{\cdot}2246980.$

The amplitude response of this filter follows very closely the theoretical cosine shape having at $f_{N/2}$ an attenuation of 3·048 dB and at f_N an attenuation of 23·912 dB.

The group delay has a small bump around $0{\cdot}9 f_N$. These responses are shown in Fig. 3.

For the second filter the pole locations are at:

$-0{\cdot}383751240 \pm j\,2{\cdot}79169040$
$-0{\cdot}558077040 \pm j\,2{\cdot}07417290$
$-0{\cdot}657791340 \pm j\,1{\cdot}27036220$
$-0{\cdot}702990460 \pm j\,0{\cdot}427352820$

and its zeros are at

$2{\cdot}424728400 \pm j\,0{\cdot}9955973200$
$10{\cdot}42712200.$

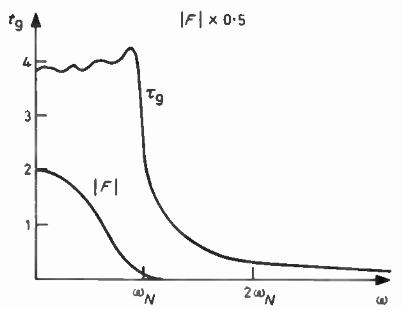

Fig. 4. Frequency response of the 8th-degree filter example.

The amplitude response has an attenuation of 3·028 dB at $f_{N/2}$ and 23·345 dB at f_N. The amplitude and group delay responses are shown in Fig. 4. It is seen that they closely resemble those of Fig. 3.

Although these filters have been calculated subject to being optimal relatively to squared intersymbol interference, they are also extremely good in the frequency domain. A general procedure has been developed to split the filter into two portions: a transmitting filter and a receiving filter. When applied to the $N = 9$, mixed numerator and $E = 10^{-4}$ case, the poles and zeros were

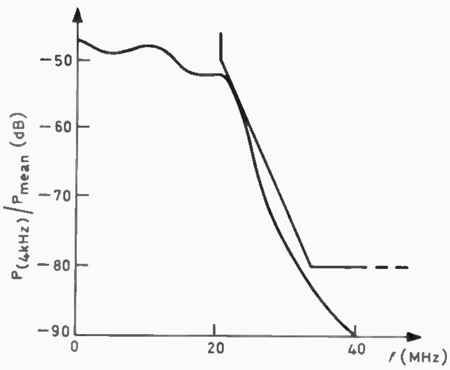

Fig. 5. Transmitter filter spectrum compared to FCC out-of-band emission limitation.

divided between transmitter and receiver to optimize performance. When used with a q.p.r.s.[3] system this division gives a bandwidth efficiency of 2·27 bits/Hz, as can be seen from Fig. 5, for a 91 Mbit/s rate. This efficiency is obtained with a deterioration in s.n.r. relative to the ideal matched filter of less than 0·3 dB.

This example has been chosen to satisfy the frequency domain specifications for the Canadian DRS–8 q.p.r.s. system. The above 9-pole filter possesses negligible i.s.i., and so does not require an additional decision-directed feedback network. Therefore, there are considerable savings over the conventional solution, which requires a composite 10-pole filter and a decision-directed feedback network.[3]

It is interesting to see how close the realizable filters are to ideal performance in the frequency domain. In the time domain, an ideal impulse response f_{id} should satisfy the constraint

$$f_{id}(t)\,.\,\mathrm{rep}_T\delta\left(t-\frac{T}{2}\right)=\delta\left(t-\frac{T}{2}\right)+\delta\left(t+\frac{T}{2}\right)$$

which leads to the frequency requirement:

$$\sum_{n=-\infty}^{\infty}(-1)^n F_{id}(\omega-2n\omega_N)=k\cos\frac{\omega T}{2}+\mathrm{j}0 \qquad (4)$$

where k is a simple scaling constant. The actual impulse response has pure time delay (due to its causal nature) of $t_0=t_m-\frac{1}{2}$. Hence the alternate folded summation of a realizable filter should be computed from

$$H(\mathrm{j}\omega)=\sum_{n=-\infty}^{\infty}(-1)^n\exp[\mathrm{j}(\omega-2n\omega_N)t_0]F[\mathrm{j}(\omega-2n\omega_N)]$$

where $F(\mathrm{j}\omega)$ is the rational function of the realizable filter.

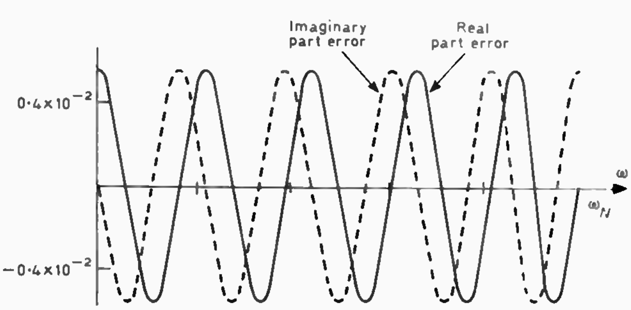

Fig. 6. Error performance of alternate folded summation for the 9th-degree example.

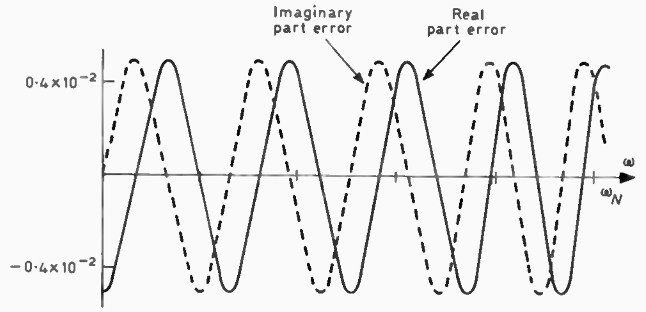

Fig. 7. Error performance of alternate folded summation for the 8th-degree example.

In Figs. 6 and 7 we show the deviation of the real and imaginary parts of $H(\mathrm{j}\omega)$ from $k\cos\omega T/2+\mathrm{j}0$, where k has been determined by a least squares fit. In practice it is found that k is very close to unity. It can be seen that these errors are equi-ripple and interlaced, in exactly the same manner as for the Nyquist I design.[6] These errors are closely approximated by†:

$$(-1)^{N+1}\exp[-\mathrm{j}(N+\tfrac{1}{2})\omega]$$

which has an inverse Fourier transform of

$$(-1)^{N+1}\delta[t-(N+\tfrac{1}{2})].$$

Therefore the main i.s.i. error should occur at the t_m+N sample point, which has been verified computationally. Other i.s.i. errors occur because the error function is not exactly periodic. It is interesting to consider that if these designs had been carried out in the frequency domain, then the error curves of Figs. 6 and 7 would have been required as *inputs*. That these non-periodic error curves are the 'best' criteria is by no means obvious from general considerations.

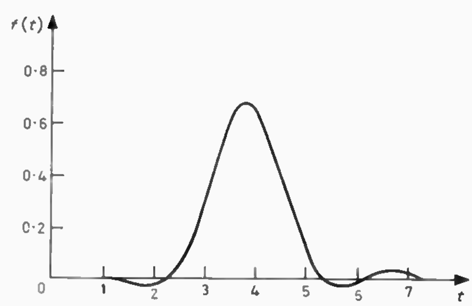

Fig. 8. Impulse response of the 9th-degree filter example ($T=1$ s).

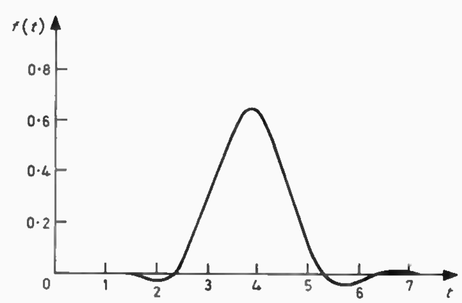

Fig. 9. Impulse response of the 8th-degree filter example ($T=1$ s).

† S. E. Nader, Private communication.

Turning now to the time domain, the impulse response of the ideal bandlimited partial response channel can be found from equation (4) to be

$$f_{id}(t) = \frac{\pi}{4}\left[\frac{\sin \pi\left(\frac{t}{T}+\frac{1}{2}\right)}{\pi\left(\frac{t}{T}+\frac{1}{2}\right)} + \frac{\sin\left(\frac{t}{T}-\frac{1}{2}\right)}{\pi\left(\frac{t}{T}-\frac{1}{2}\right)}\right] \quad (5)$$

The actual impulse responses of the filters in the above examples have been computed and are displayed in Figs. 8 and 9.

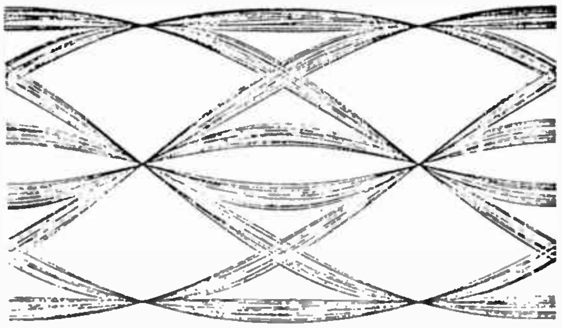

Fig. 10. Eye diagram for the 9th-degree example.

It was found that for the $N = 9$ example the main lobe ratio $f(t_m)/f(t_m - 0{\cdot}5)$ is within 1% of the ideal value of $\pi/4$. This ratio decreases as E and N decrease. The overshoot for the first example was 6·32% and for the second example was 6·22%, as compared to the equation (5) overshoot of 7·089%. The eye diagram for the 9th-degree filter referred to in the example is shown in Fig. 10. From it we can see that there is practically no i.s.i. at the zero crossings, resulting in a very good waveshape for timing recovery. This is one of the reasons for modelling the channel in such a way that its timing extraction may be done by the threshold method.

5 Conclusions

An analytic theory has been developed for the design of an optimal partial response filter in which every pole is used as a degree of freedom.

The resulting performance is at a global minimum, which implies that this performance is not sensitive to small changes in filter parameters. Component spreads are therefore accommodated in the design.

Although the filter has been designed to optimize certain conditions in the time domain, it also has optimal attenuation and group-delay properties in the frequency domain, and satisfies the conditions set by Lender.[9] Hence this filter requires no separate group-delay equalization.

Typically, a complete filter design takes less than one second of computer time. Practical examples of this design process are given. Since only a few zeros are necessary in this design for precursor i.s.i. elimination, the remaining ones can be used for stopband shaping or other purposes.

The design procedure is capable of extension to non-impulsive inputs and non-ideal channels.

6 Acknowledgment

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8 Appendix

Here we are concerned with finding the effects of frequency domain distortions on the alternate folded summation criterion of equation (4). As suggested by Nyquist,[8] we can add the impulse response $s(t)\cos \pi t/T$ to a Class I partial response filter without perturbing the response at the instants $t = 2(k+1)T/2$.

This function has a Fourier transform:

$$S\left(\omega - \frac{\pi}{T}\right) + S\left(\omega + \frac{\pi}{T}\right).$$

As $s(t)$ is a real function of time the real part of its FT is symmetric and the imaginary part skew-symmetric, that is

$$\mathrm{Re}\, S(\omega) = \mathrm{Re}\, S(-\omega)$$
$$\mathrm{Im}\, S(\omega) = -\mathrm{Im}\, S(-\omega).$$

So we can add to the ideal cosine transfer function another function whose real part is symmetric around π/T and symmetric around the origin and the imaginary part skew-symmetric around π/T and skew-symmetric around the origin, without affecting the ideal i.s.i. performance.

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Indexing Terms: Television receivers, Satellites, broadcasting

Direct broadcast satellite receivers

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

SUMMARY

Broadcasting of television and radio programmes from satellites to individual receivers is now technically possible. For Europe, allocations of up to five channels in the 12 GHz band to each country have been authorized and several countries plan to implement services in the next few years. The required characteristics of receivers for individual reception such as sensitivity, selectivity, signal processing, etc. are reviewed and it is shown that these can be satisfied by available technology.

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1 Introduction

The feasibility of broadcasting direct to the home from a geostationary satellite was recognized at least two decades ago and by the mid-1960s the receiver industry was already considering the implications. In 1971 the World Administrative Radio Conference (WARC) allocated a number of frequency bands for satellite broadcasting, either exclusively or on a preferentially-shared basis with other services. Of particular interest for Europe (Region 1) was the allocation from 11·7 to 12·5 GHz, which offered multi-channel capability free from interference to or from existing terrestrial u.h.f. services. The WARC in 1977 made detailed allocations of channels and other parameters for broadcasting in this band—not merely to permit orderly implementation of satellite broadcasting but also for the benefit of other services permitted to share the band. Table 1 shows these allocations as modified by a subsequent WARC in 1979.

Table 1

Satellite broadcast allocations

Band (GHz)	Frequency range (GHz)	Region*	Notes
0·7	0·62– 0·79	All	Television only, subject to interference constraints
2·5	2·5 – 2·69	All	Community only
12	11·7 –12·1	1, 3	
	12·1 –12·2	All	
	12·2 –12·5	1, 2	
	12·5 –12·7	2, 3	Community only in Region 3
	12·7 –12·75	3	Community only
23	22·5 –23	All	
42	40·5 –42·5	All	
85	84–86	All	

* Region 1 Europe and Africa, Region 2 North and South America, Region 3 Asia and Australasia.

Although there was (and perhaps still is) a widespread misconception that satellite broadcasting would enable the individual to receive an almost unlimited selection of programmes from space, technical (and political) considerations strongly favoured a system of national satellites providing total coverage of individual countries with a limited number of programmes. A prime factor was the decision that programmes should be capable of individual reception anywhere in a country—although in urban areas community reception via cable distribution systems (often already existing) was likely as a convenient means of cost-sharing. Another consideration was frequency re-use without interference, so that each of the many countries involved in the European plan could be allocated enough channels to make a new service and new receiving equipment worthwhile.

2 Allocations and Transmission Parameters

In planning for satellite television broadcasting at 12 GHz a number of assumptions were made—in

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particular the use of existing 625-line colour television signal standards (with video signal base-bandwidth of about 5 MHz). Considerations of likely satellite power, receiver sensitivity, aerial directivity and permissible co-channel interference showed that for a practicable system of essentially national services conventional amplitude modulation was precluded. A plan was therefore evolved assuming wideband frequency modulation, but other systems, such as digital modulation, are not excluded. Eventually, after considerable computer modelling, a scheme providing five 27 MHz wide channels for each country was arrived at (although some groups of adjacent countries have opted for some channel sharing). Although allocated primarily for television, one or more channels may, if desired, be used by each country for other purposes, such as multiple programme sound broadcasting. However, parameters for such use have yet to be agreed.

Table 2

Probable 12 GHz television standards

baseband video	standard 625-line
baseband audio	f.m. subcarrier
12 GHz carrier modulation	f.m.
carrier deviation	13·3 MHz p-p
pre-emphasis	CCIR Rec. 405–1
signal bandwidth	~ 27 MHz
energy dispersal	600 kHz p-p

In arriving at the parameters for television broadcasting the following was assumed (Table 2):

(1) Wideband f.m. would be used with 13·5 MHz p-p deviation for a 1 V composite video signal at ~ 1·5 MHz (pre-emphasis according to CCIR Recommendation 405–1 being assumed). With a single subcarrier sound signal, having a peak-to-peak deviation of 2 MHz, this would give a signal bandwidth of about 27 MHz. The possibility of two (or more) television sound subcarriers for stereo or a second language commentary is, however, not precluded.

(2) Co-channel (and hence image) and adjacent channel protection ratios would be 31 dB and 15 dB respectively.

(3) The receiving system would have a nominal G/T of 6 dB/K—equivalent to a 0·9 m diameter receiving aerial of agreed polar response with a front-end noise figure of 6 dB and a 2 dB allowance for pointing and other losses.

(4) The received flux should yield good pictures within the service area even after allowance for satellite and receiver aerial pointing errors, some transmitter power loss with life and propagation losses due to rain, etc. (which were assumed to be < 3 dB for 99% of the worst month of the year).

(5) Satellite station-keeping and aerial pointing accuracy would each be $< \pm 0{\cdot}1°$. Satellite locations would be chosen to minimize co-channel interference and to ensure that loss of solar power during the nocturnal satellite eclipses at the spring and autumn equinoxes would not occur until well after midnight when transmissions would have ceased.

(6) The service area 'footprint' of each satellite would be limited as far as practicable to the shape of the receiving country.

Table 3

12 GHz allocations—Region 1

channel spacings	19·18 MHz
(signal bandwidth	27 MHz)
number of channels	40
channels/country	5
national 'spread'	360 MHz
polarization	circular (L or R)
nominal power flux at Earth's surface	-103 dBW/m^2
satellite locations	$+5°, -1°, -19°, -25°, -31°, -37°$

Based on these assumptions the following plan was arrived at (Table 3):

(1) The design flux (99% worst month, etc.) at a receiving installation at the edge of the service area is -103 dBW/m^2. For a geostationary satellite in a 42000 km radius equatorial orbit this is equivalent to an e.i.r.p. of about 62 dBW. For a satellite with a 1° beam (~ 1·5 m dish), suitable for the UK, this corresponds to a satellite r.f. power of around 250 W, which is now practicable.

(2) A limited number of satellite locations have been chosen so that in many cases signals can be received (albeit at lower quality) from satellites serving adjacent countries without aerial re-pointing. For most large West European countries satellites are located at 19°W with the exception of those of the UK, Eire, Portugal and Spain which are at 31°W.

(3) Although slightly less convenient for the receiver, circular polarization (left- or right-hand) will be used since it eases satellite aerial design (and, incidentally, receiving aerial alignment).

(4) Channel spacing is set at 19·18 MHz which yields 40 channels in the 800 MHz band. Note that with a *signal* bandwidth of 27 MHz some overlap appears to occur, but adjacent channels are not allocated to the same or adjacent service areas and the necessary protection is allowed for.

(5) As stated, most countries have been allocated 5 channels, but there are some exceptions, e.g. the Scandinavian countries by arrangement have 3 national channels each and 2 common channels.

(6) For most countries the spread of allocations is limited to 360 MHz in the top or bottom half of the band, with guard bands 3 channels wide. This simplifies receiver design.

(7) To reduce potential interference to terrestrial fixed services energy dispersal will be used consisting of an additional 600 kHz p-p triangular deviation of the carrier at a sub-multiple of field frequency, e.g. 25 Hz.

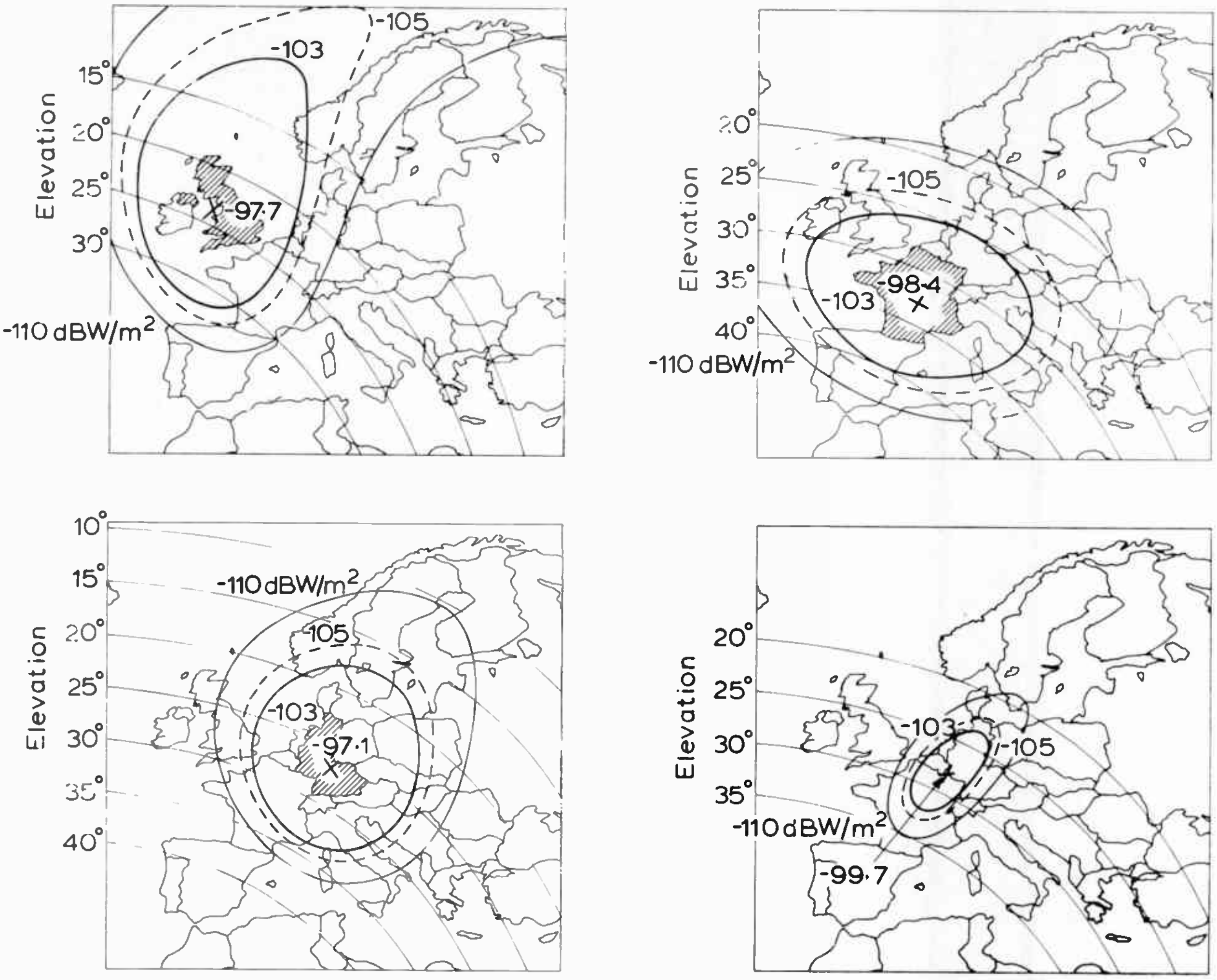

Fig. 1. Parameters and footprints for satellite coverage of the United Kingdom, France, West Germany and Luxembourg

Parameters							
	Beam	Polarization	Channels	Peak e.i.r.p.	Orbital position	Orientation	Antenna peak gain
United Kingdom	1·8° × 0·7°	1	4, 8, 12, 16, 20	65·2 dBW	31°W	142°	43·0 dB
France	2·5° × 1·0°	1	1, 5, 9, 13, 17	64·0 dBW	19°W	160°	40·4 dB
West Germany	1·6° × 0·7°	2	2, 6, 10, 14, 18	65·7 dBW	19°W	147°	43·6 dB
Luxembourg	0·6° × 0·6°	1	3, 7, 11, 15, 19	63·1 dBW	19°W	0°	48·7 dB

Figure 1 summarizes the characteristics of the UK, French, West German and Luxembourg allocations.

3 Broadcast Satellite Plans

Within the next few years a number of European countries are expected to implement a 12 GHz satellite television service. West Germany and France, in particular, are already committed to launching full service satellites in 1983/4. These will probably have two accompanying sound channels and signals from the latter will be capable of reception with good quality in at least the southern part of the UK. Luxembourg, Switzerland and the Nordic countries are also known to be seriously considering introducing services. In the UK, with use of our fourth u.h.f. channel only recently authorized and the future use of the now obsolescent 405-line v.h.f. channels not yet resolved, there seems no urgency to do anything—but it is nevertheless argued that we should not fall behind. However, in Europe as a whole, it is clear that in only 2 or 3 years' time a rapidly growing receiver market is likely and the industry needs to be ready.

new satellite signals. However, since most households already have at least one receiver designed only for the existing service, the initial requirement will be for one or more add-on units capable of converting the s.h.f. f.m. signal to a u.h.f. a.m. signal suitable for the standard receiver (see Fig. 3). This is an unwelcome complication, but already proposals exist for fitting future standard receivers with a video and audio signal interface socket and this would at least remove the need for a u.h.f. amplitude re-modulator.

Another factor affecting receiver design is the need to keep the length of the s.h.f. feed from the aerial to the receiver to an absolute minimum on grounds of cost vs. loss. This leads to the double-superhet. concept, in which a fixed-tuned broad-band frequency down-converter (the outdoor or 'head' unit), located as near as possible to the aerial, translates all the available programme channels to

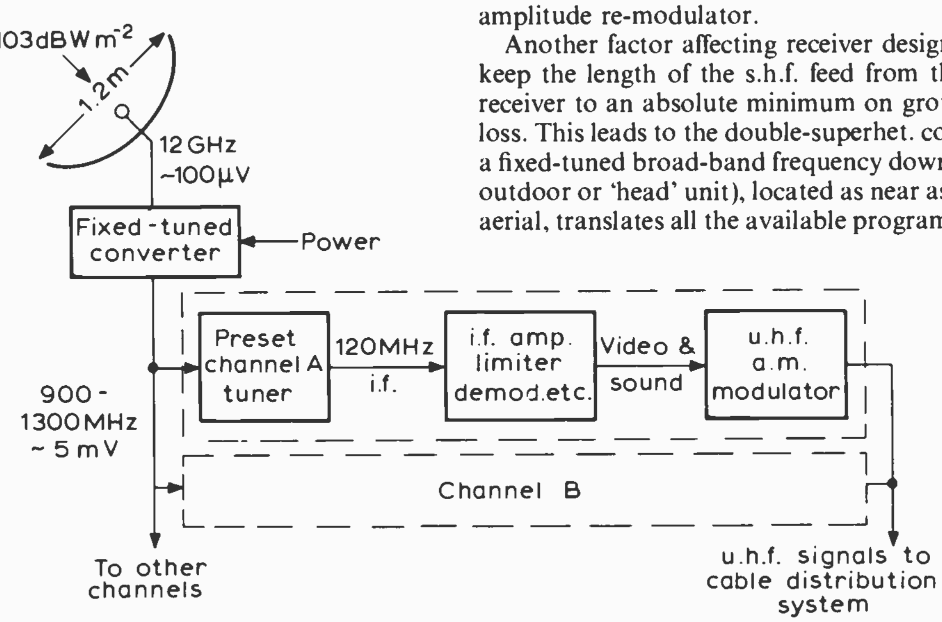

Fig. 2. Outline of typical community television receiving system.

a lower intermediate frequency, typically in the region of 1 GHz. Low-cost, low-loss cables for the downlead are then readily available. A second (indoor) unit associated with each standard television receiver then provides channel selection and signal demodulation to yield the required composite video colour signal and one or more subcarrier sound signals, which may then be suitably processed for feeding to the standard receiver.

An important advantage of this double superhet. approach, using a common outdoor unit and separate indoor units, is that, in the increasingly common situation where two or more receivers are used in the same household, the facility for viewing different national programmes simultaneously is retained. Indeed, in many European countries it will even be possible, without equipment modification, simultaneously to receive (albeit at lower quality) programmes of adjacent countries whose signals have the same polarization and satellite position and fall within the same half of the 12 GHz band. Only when one or more of these criteria is not met will it be necessary to have more complex receiving equipment.

4 Television Receiving Systems

The form of receiving equipment required for satellite television reception will obviously depend on whether environmental and other circumstances compel or favour individual or small or large community systems. Of these the first is of greatest interest and poses the most difficult task for the manufacturer, since large quantities will be required having adequate performance at an economic price. Attention will therefore be concentrated on such receivers, since for large communities, particularly where cable distribution systems already exist, the problems are much less severe. For these, although multiple (and more sensitive) receivers will be required for simultaneous multi-channel signal reception and reprocessing (to v.h.f. or u.h.f. a.m.) for distribution, the greater cost can be shared between a large number of viewers. A typical scheme is outlined in Fig. 2. More modest versions could be used for small communities such as a large block of flats or a small housing estate.

Let us therefore examine in more detail the requirements of an individual receiving system for a single household, assuming frequency modulation of the 12 GHz carrier by a conventional composite video colour signal (e.g. PAL) accompanied by one or more f.m. sound subcarrier signals.

In the longer term we can expect there to be composite receivers on the market capable of receiving both conventional u.h.f. a.m. terrestrial broadcasts and the

5 Individual Satellite Television Receiver Components

Having discussed the general requirements for a satellite television receiver the various sub-units will be considered in rather more detail. (See Fig. 4.)

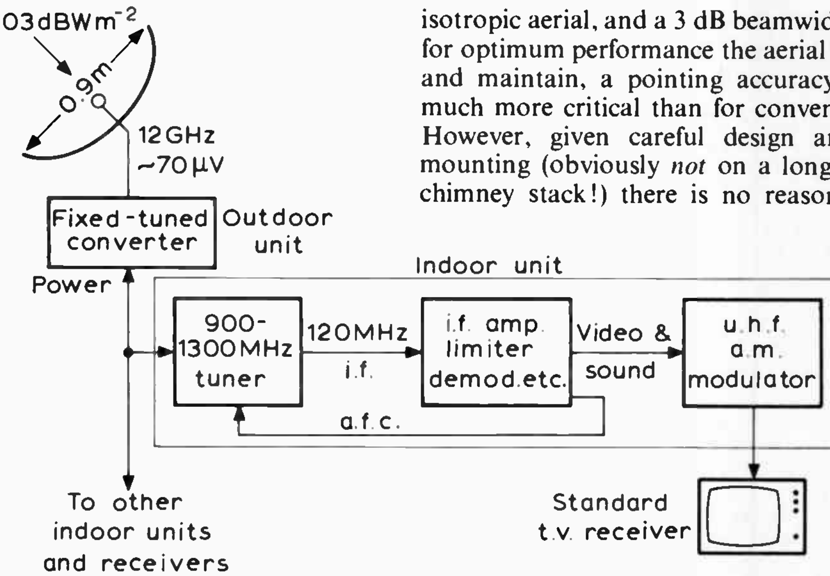

Fig. 3. Outline of typical individual television receiving system.

isotropic aerial, and a 3 dB beamwidth of about 2°. Thus for optimum performance the aerial must be set up with, and maintain, a pointing accuracy of < 0·5°. This is much more critical than for conventional u.h.f. aerials. However, given careful design and location of the mounting (obviously *not* on a long pole strapped to a chimney stack!) there is no reason why this stability

5.1 The Aerial

In the system planning an individual receiver aerial equivalent to a parabolic dish of 0·9 m diameter was assumed as being a reasonable compromise in terms of gain, directivity (including side-lobe response), required pointing accuracy and stability and, of course, cost. Such an aerial has a gain of about 37 dB relative to an

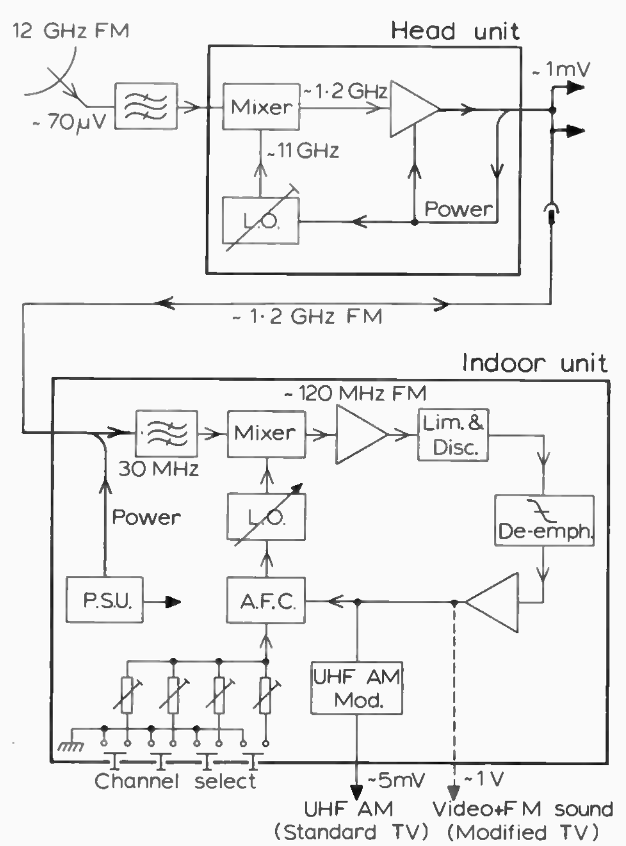

Fig. 4. Double superheterodyne 12 GHz f.m. converter for individual reception.

should not be maintained under all reasonable environmental conditions. As regards initial alignment, experience has shown that given a simple compass and spirit-level (which might be an integral part of the aerial assembly) it is possible to provide initial mechanical alignment sufficiently accurately for a signal to be obtained from the satellite and then to use the signal itself for final precise alignment. It should be noted that the required aerial elevation and bearing will depend on the latitude and longitude of the receiving location relative to the satellite position. For the UK the elevation will range from about 28° in the Scilly Isles to about 17° in the Shetlands and the bearing from about 27°W of S in the Outer Hebrides to about 39°W of S in Kent.

It is generally assumed that the most likely form of aerial will be a prime feed or Cassegrain parabolic reflector with a waveguide feed from the focus. Depending on quantities and relative economics the reflector could either be pressed from metal, or moulded in, for example, glass reinforced plastic (fibre-glass), with an embedded conducting mesh, or a suitably-protected metallic surface coating. In either case, to achieve the required gain and polar response (Ref. 1 and Fig. 5) the profile must be accurate to within ±1 mm.

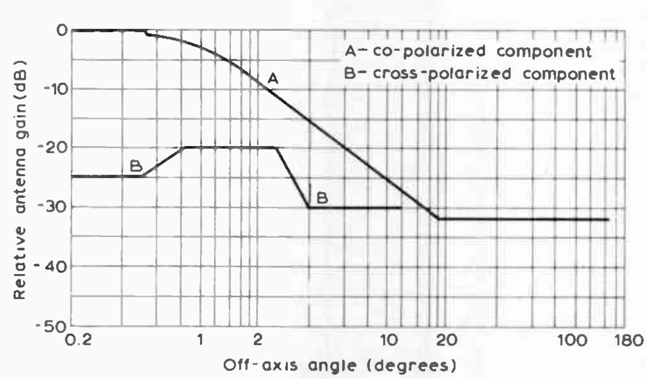

Fig. 5. Individual receiver aerial response.

However, provided they can meet the specification other solutions are not excluded. For example, some years ago the BBC suggested a Fresnel reflector.[2] This is essentially a sliced/compressed paraboloid which offers the interesting possibility of the choice of an offset angle of maximum response. There is also continued speculation about the prospect of arrays of small printed dipoles on a large dielectric panel. However these are still a very long way from yielding the required gain and directivity.

5.2 The S.H.F. Feed

The s.h.f. feed to the head unit has the additional functions of providing rejection of image frequency signals, suppression of local oscillator power radiation into the sky and selection of the appropriate signal polarization—if necessary with conversion to linear polarization for interfacing to the input circuit of the outdoor unit. Possible arrangements are a short circular waveguide feed, with the outdoor head unit located near the focus of the main reflector, or a Cassegrain (dual-reflector) arrangement with the head unit located behind the reflector.

In principle the problems of image signal rejection and oscillator radiation suppression can most simply be solved by choosing an 'oscillator-low' configuration. In this case the necessary filtering is readily achieved by virtue of the fact that a suitably chosen feed diameter will have a high-pass characteristic. This would be satisfactory in most locations but in some situations high-power radars in the 10 GHz band could be a potential source of interference. The 'oscillator-high' configuration may therefore be preferable in this respect, but involves extra complication, such as the insertion of iris diaphragms into the waveguide to yield the required bandpass response. Selection of the required polarization, and its conversion to linear polarization suitable for the head unit if required, may be achieved without too much difficulty, for example by insertion of appropriately dimensioned and spaced posts into the waveguide. However, where it is desired to be able to select either of the two circular polarizations at will by means of a switch, the feed will become much more complex and expensive.

5.3 The Head Unit

This unit converts and amplifies the band of received signals from 12 GHz (at a level $\simeq 70\,\mu$V) down to a first i.f. low enough (e.g. $\simeq$ 1 GHz) for feeding to the indoor unit via conventional coaxial cable (at a level of $\sim$ 5 mV). For reasons associated with the design of the indoor unit tuner it is desirable to restrict the bandwidth to around 400 MHz. The 5 channels assigned to most countries have therefore been arranged to fall in either the upper or lower half of the 800 MHz total band from 11·7 to 12·5 GHz. To minimize the risk of interference from other services, especially u.h.f. television, the first i.f. band is likely to be chosen to be 900–1300 MHz.

Until relatively recently, in the absence of low-cost s.h.f. pre-amplifiers, prototype down-converters have consisted of a direct mixer input, using Schottky barrier diodes, followed by i.f. amplification. One possible solution based on this approach uses conventional microstrip technology, in which distributed-line circuits are evaporated on alumina or a similar dielectric substrate, and incorporates a balanced mixer.[3] It is now known that with this arrangement it is difficult to achieve a noise figure much below 6 dB. A second solution developed by Konishi[4] consists of a single-ended mixer on a planar-in-waveguide configuration and has yielded noise figures approaching 4 dB.

To drive the mixer requires 10–20 mW of local oscillator power at s.h.f. Assuming automatic frequency control can be applied to the second local oscillator in the indoor unit, this s.h.f. oscillator must be stable to 1 or 2 MHz over an outdoor unit temperature range which in some parts of Europe can be 60°C or more. An obvious solution is to use a v.h.f. crystal oscillator followed by varactor or step-recovery diode multiplication, but the arrangement usually preferred has been a Gunn transit-time device in a low-cost waveguide cavity with dielectric temperature compensation (e.g. by means of a TiO_2 post) to obtain the required stability.

However, the last few years have seen rapid developments in low-noise GaAs field-effect transistors so that all-transistor head-units, with discrete-transistor s.h.f. preamplifiers, mixers, i.f. amplifiers and dielectric-resonator-stabilized s.h.f. oscillators, having a noise figure of less than 4 dB are now technically possible.[5,6] Given sufficient market impetus these should come down in price to an acceptable level. In the meantime, work is in progress aimed at monolithic integration of such units. In all cases it is a simple matter to provide power for the head-unit via the i.f. downlead.

It should be noted that whilst the 2–3 dB improvement in head-unit noise figure offered by these developments, as compared with the figure assumed originally for the planning, is not essential, it does offer a greater margin above threshold against propagation-fading and alignment errors, as well as the prospect of better reception of signals primarily intended for neighbouring countries.

5.4 The Indoor Unit

The remaining signal processing necessary to yield the desired satellite television programme will be provided by an indoor unit, which in the early stages of a service will be a separate unit associated with each standard receiver. This will incorporate a number of functions.

First, a 900–1300 MHz tuner of essentially conventional design, with automatic frequency control of the local oscillator, but having an output bandwidth of $\sim$ 27 MHz, will be used to select the required channel and convert it to the final i.f., which seems likely to be chosen to be around 120 MHz. Since the incoming f.m. signal will have better immunity against interfering signals than conventional a.m. (cf. Sect. 2) the requirements on image and adjacent channel rejection will be less stringent. The main i.f. selectivity will, however, need to have a carefully designed flat response over a 27 MHz bandwidth with a steep roll-off beyond, together with a good phase response, to keep intermodulation distortion products down to an acceptably low level. With conventional lumped LC elements this would require at least a four-pole filter configuration, but an attractive alternative is the acoustic surface wave filter, which can yield the desired response with only a single component.

The following functions of i.f. amplification and frequency demodulation can already be realized using available integrated circuits and the design of suitable mass-production custom i.c.s should present no real problems. After demodulation the necessary h.f. de-emphasis can be applied together with removal of the energy dispersal waveform. It is expected that in a domestic installation this latter function can be adequately performed by a simple d.c. restorer. (For community systems, however, more sophisticated techniques such as keyed-clamping, may be desirable.) After filtering the composite video colour signal and accompanying intercarrier sound signal(s) are available for further processing as necessary. From this point also can be derived the automatic tuning control for the indoor unit.

As indicated above, beyond this point the form of signal processing required will largely depend on the input facilities available in the domestic television receiver. For existing receivers designed only for u.h.f. input of an a.m. video signal with only a single sound carrier signal it will be necessary to convert the sound subcarrier deviation, etc. to the standard form (e.g. by demodulation and remodulation) and to remodulate this and the video signal to provide the required u.h.f. signal. If the satellite channel programme also has more than one accompanying sound signal it will also be necessary to select the one desired.

None of these processing functions poses any fundamental difficulties but they will obviously increase cost and could degrade the overall performance. Clearly, the planned availability in the near future of standard receivers with a composite video and audio input socket will offer useful advantages in both these respects.

In the longer term, of course, when the new satellite services become more generally available we can expect to see the additional satellite-reception indoor functions combined in an optimized cost-effective way in a dual purpose (satellite/terrestrial) television receiver.

6 Sound Broadcast Reception

The 12 GHz broadcast plan allows the use by each country of one or more of its channels for multiple-programme sound broadcasting. So far no standards for this purpose have been laid down but Ref. 7 lists some possibilities. An important factor affecting the choice is the resulting receiver complexity. For example, in a domestic installation it seems desirable—if not essential—to use at least the same aerial and outdoor unit as for television. This probably precludes the use of wideband ($\sim$600 kHz) simple f.m. systems, which although efficient from the point of view of satellite power and the number of programmes possible in a 27 MHz channel, require very high receiver local oscillator stability. At some cost in power efficiency and number of programmes, it therefore seems preferable to multiplex the different programmes in some way prior to frequency modulation of the 12 GHz carrier. If this is achieved within the 5 MHz video baseband the existing satellite television indoor unit could then also be used to recover this multiplexed signal at the video output point. It would then be a relatively simple matter to retrieve the required programme by de-multiplexing and demodulation, either in a separate unit or as an extra part of the existing television indoor unit. Possible methods include analogue f.m. in frequency division multiplex (f.m./f.m.) or digital signals in time division multiplex (e.g. p.c.m./f.m.). In view of the general move towards digital techniques and the likely inter-modulation problems with f.m./f.m., systems of the latter type would appear more attractive, but studies are not yet complete.

7 Conclusions

Satellite television broadcasting services will be introduced by a number of European countries in the next two or three years. The requirement for mass-produced equipment for individual reception of such signals has been shown to be realizable and development of low-cost solutions is now in progress. Although not yet resolved, solutions also exist for multiple sound broadcasting which are largely compatible with those for television, thus offering maximum economy for dual-purpose equipment in the home.

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Indexing Terms: Frequency changers, Digital techniques

A real-time digital frequency transposer for the 100 Hz to 1·25 MHz band

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SUMMARY

A real-time digital signal processor is described which, because of its wide band operation, close tolerance bandwidths and linear phase/frequency response, has a wide range of applications. These include general data handling and conditioning, frequency shifting and transposition, pre-processing for spectrum analysers and bandwidth compression.

Apart from analogue-digital convertors and pre-amplifiers, the instrument is wholly digital and makes use of a combination of microprocessor techniques and hard-wired logic to achieve performance goals.

The transposer can either be operated manually or directly from a host digital computer via a parallel interface. Spare card space is also available to include either an IEEE488 or RS232 interface, as required.

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1 Introduction

In a wide range of signal processing applications, it is necessary to be able to manipulate bands of signals occupying prescribed parts of the frequency domain. By way of examples, we may need to extract a limited band from a wide spectrum so that its content can be studied in more detail. Alternatively, we may wish to shuffle signalling channels in a block. Particularly in the latter case, it is essential that no phase distortion is introduced in this process so that wave shape is maintained.

The digital frequency transposer described in this paper is able to accept a band of signals (spectrum) lying in the range of 100 Hz to 1·25 MHz, to trim the bandwidth to any one of the 64 widths in the range 50 Hz to 50 kHz, and to resubmit the spectrum, shifted if necessary, to any other part of the same band.

Some of the important attributes of the way in which this instrument performs these functions are as follows:

(i) All the signal processing functions are carried out using algorithms having a linear phase/frequency (group delay) characteristic. This was considered particularly important as an obvious application would be in the handling of keying signals and digital data for which waveform integrity is vital.

(ii) The reconstituted signal should have a minimum of spurious signal components in other parts of the operational spectrum created as a result of digital signal processing.

(iii) A large number of bandwidths should be available in a geometric progression. (It is not easy to maintain a group delay characteristic in a system required to achieve a 50 Hz bandwidth on signals centred at 1·25 MHz.)

The instrument is a hybrid consisting of a microcomputer plus hard-wired logic. Broadly, the logic is needed to program signal processing at a 5 MHz sampling rate and the microcomputer performs the management role. Tasks for the latter include preparation of computer or keypad originating data, performance and ranging checks, bandwidth and spectral inversion control and self check.

2 Design Approach

In order to achieve the required design, in particular the variable bandwidths, it was decided that the incoming signal should first be converted to baseband using a quadrature down-convertor to create both sine and cosine components. It would then be possible to substitute low-pass filters having cut-off frequencies over the appropriate range rather than band-pass filters with a performance up to 1·25 MHz.

As baseband conversion implies the creation of up as well as down products, the sampling frequency for the a.d.c. (Fig. 1) has to be at least 4 times the highest incoming frequency. By so doing, the up-conversion products would not be folded back (in frequency) to the intended pass-band. A sampling frequency of 5 MHz was therefore chosen.

The linear phase/frequency characteristic of the system, as a whole, tended to imply that each processing component would individually have to obey such a law. The possibility of post-correcting phase errors created in

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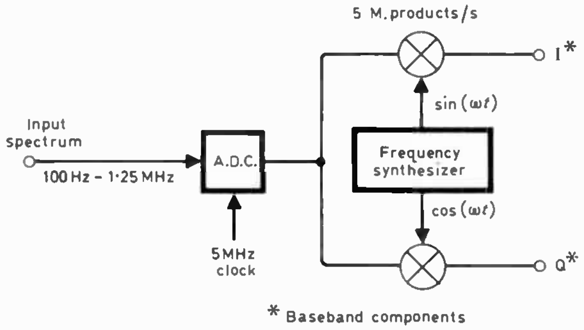

Fig. 1. A.d.c./quadrature down-convertor for transposer.

say, infinite impluse response filters, was discounted both on grounds of quality of correction and difficulty of processing. The main bandwidth-limiting filters are provided by transversal elements. The number of delay line sections and coefficients for such filters is very high to achieve a reasonable filter shape factor (100 in this case) and in order to avoid having to store 64 sets of such coefficients, a single set was chosen and a sampling rate shifting method adopted to change bandwidth. (Fig. 2)

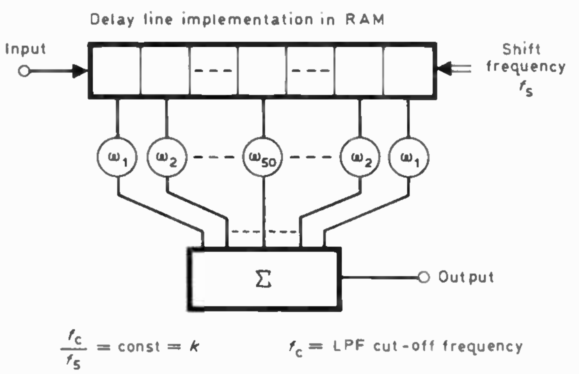

Fig. 2. Transversal filter to give linear phase/frequency characteristic.

The implication of the above is that the 5 MHz sampling rate must be changed to a lower rate which depends on the bandwidth chosen. A decimation type filter was chosen for this purpose. Again, it had to have a linear phase/frequency characteristic, but even more important, be capable of changing sampling rate without abrogating the Shannon sampling criterion for the outgoing sample stream. (The a.d.c., working at 5 MHz, is capable of handling signal components up to 2·5 MHz. For this design, if the selected bandwidth is 50 Hz, corresponding to 25 Hz in each of the quadrature channels, the sampling rate through the transversal filters is only 250 Hz.) The decimation filter therefore has two jobs to do:

(i) Reduce the sampling rate to that of the transversal filter
(ii) Suppress signal components above the Shannon limit for the outgoing sampling rate to a level acceptable for the instrument specification.

The required characteristics were obtained by a series of averaging circuits as described in Section 4.

The signal flow diagram for the front end of the transposer appears in Fig. 3.

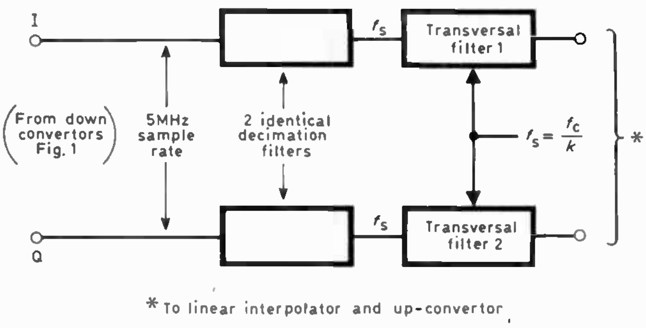

Fig. 3. Decimation and transversal (LPF) filters for the digital transposer.

In order to reconstitute the signal (which is, at present, in quadrature form) and to centre it on some new 'carrier', an up-conversion process is required. In its simplest form, it can just be the converse of the down (baseband) convertor. However, it is clear that the range of outgoing frequencies demands the same sampling rate, 5 MHz, as the incoming one. In order to carry out up-conversion, say by the use of digital multipliers, it is evident that the samples entering the multiplier should both be at a 5 MHz rate. This will automatically be the case for the local oscillator signal if it is designed on the same lines as that for the down convertor. However, the output of the transversal filters will be at some lower rate, possibly down to 250 Hz. If the same signal sample is applied many times to an up convertor to synthesize a zero-order hold (Fig. 4) because of lack of sample values, a series of spurious products is generated in the frequency domain. These products will be spaced at intervals of this (lesser) sampling frequency from the carrier frequency. Although they will not dominate the outgoing spectrum, they will be very significant and unacceptable. Thus in the transposer, a sampling rate convertor (linear interpolator) was interposed between the transversal filters and the up convertor. (Fig. 5) Of course, simply increasing the sampling rate to 5 MHz does not solve the problem as this is no different from allowing the same transversal filter samples to reside at the input to the up-convertor multiplier for many 5 MHz clock periods. The analogue signal implied by the transversal filter output must be 'adjusted' so that, in effect, it 'justifies' the new sampling rate with which it associated. This is achieved by interpolating between the low sample rate samples and creating genuine 5 MHz

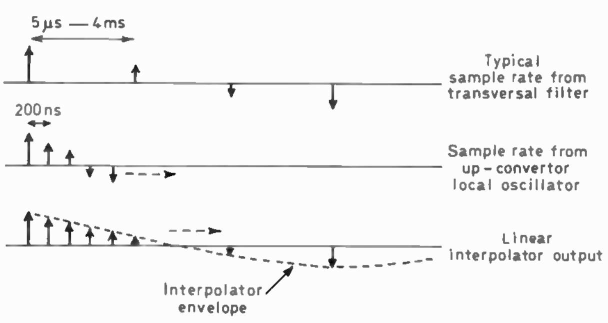

Fig. 4. Linear interpolator and up-convertor sampling frequency relationships.

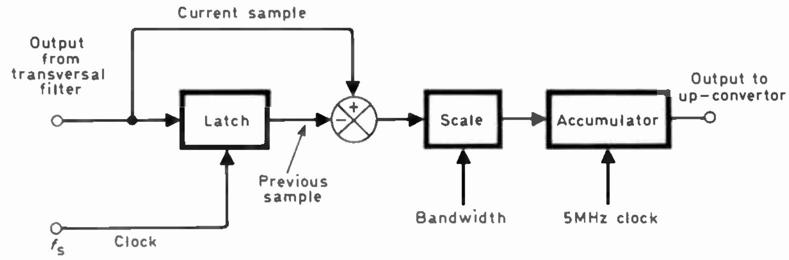

Fig. 5. Linear interpolator implementation.

samples from it. It was found that, although interpolation can be made complex without limit in this application, particularly in view of the possible tremendous disparity between sample rates (250 Hz vs. 5 MHz), a simple linear (first-order) interpolator could suppress unwanted signal components to a suitable level. A possible improvement on such a scheme and one which would not be unduly costly could use a two-panel Simpson type of second-order formula.

The final up-converted signal, now implanted on its new carrier, is ready for conversion to analogue form and low-pass filtering to restrict the band to 1·25 MHz. (Digital-analogue conversion creates steps in the outgoing analogue waveform due to the zero-order hold nature of its output which, theoretically, creates frequency components without limit.)

3 Signal Processing Algorithms

The final analogue signal flow diagram is shown in Fig. 6. It will be seen that a number of hardware processing algorithms are involved of which some demand very prodigious processing rates. Because of this, and for maintenance reasons, each section shown in Fig. 4 has its own physically co-existing hardware counterpart. Taking each processing section in turn:

(i) The down convertor consists of a pair of multipliers, fed with sine and cosine of the output of a frequency synthesizer whose output can be programmed to lie in the range 100 to 1·25 MHz and whose sampling rate is 5 MHz. Separate l.s.i. multipliers are used for the real and quadrature channels each capable of forming a product in around 100 ns. (A 5 MHz sampling rate gives 200 ns for each conversion.) The frequency synthesizers are based on an accumulator which is incremented by a literal, itself directly proportional to the frequency to be synthesized. The accumulator advances and overflows modulo–2^N, i.e. cycles. Its output can be considered as an angle for application to an angle/sine convertor. The latter is provided in the form of a bipolar p.r.o.m. The p.r.o.m. is only programmed for one quadrant: a combination of complementary addressing and programmable output inversion makes possible the creation of sines of the other three quadrants and all the corresponding cosines.

(ii) The decimation filters accept samples at a fixed rate of 5 MHz, and output at between 250 Hz (for a 50 Hz bandwidth) to 200 kHz for a 50 kHz bandwidth. Two types of averager are used (Fig. 7): the first provides averaging without a change of sampling rate and is useful for band tailoring; the second provides sampling rate reduction as well. A total of three of these was used to obtain the required out-of-band rejection.

(iii) The transversal filters had to be able to work on incoming samples of up to 200 kHz. In the time available (5 μs), all 100 coefficients for each filter had to be processed. Fortunately, these filters, in order to achieve the group delay characteristic, have filter coefficients in a cosine symmetric set (Fig. 2) so that only 50 different values are involved. This reduces the problem to 50 multiplication in each filter every 5 μs. Use of tandem l.s.i. multipliers in each filter operating in a 2-phase format makes possible this throughput.

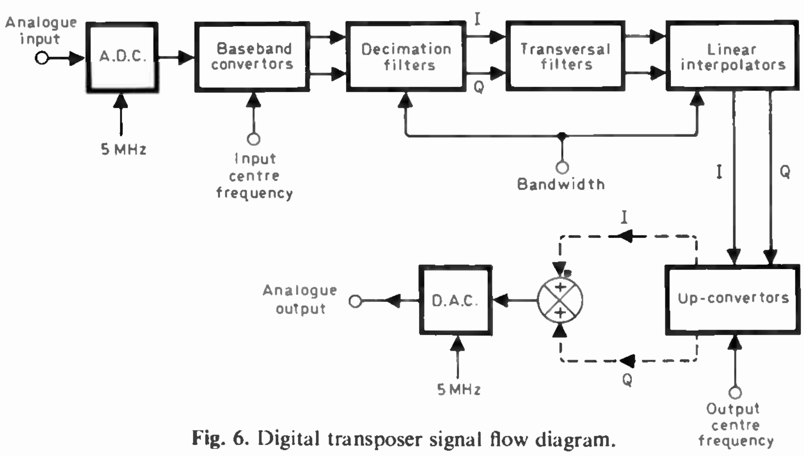

Fig. 6. Digital transposer signal flow diagram.

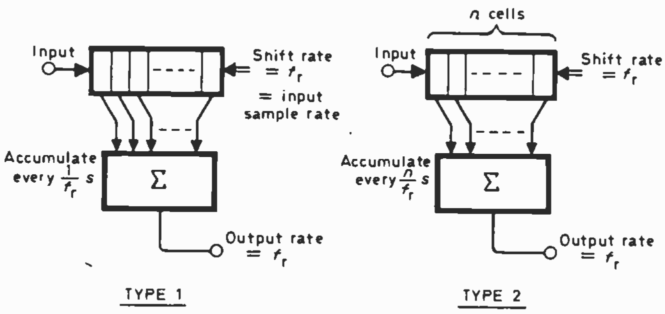

Fig. 7. Decimation filters (averagers) types 1 & 2.

4 Hardware Design

The design of some of the hardware has already been described. Amongst the more important items not mentioned in detail are the following.

4.1 The Decimation Filters

Type 1. The filters which provide no sample rate frequency reduction must produce output at up to 5 MHz. The incoming samples are written to contiguous locations in a bipolar r.a.m. and also read from contiguous locations. The relationship between write and read locations simply reflects the length of block to be summed at any time. The addressing is modulo the longest block length. If the filter is assumed initially empty, an accumulator can be cleared to indicate this. During run time, each incoming value to the filter can also be added to the accumulator's contents, and each outgoing one subtracted. By so doing, the total contents of the block need not be totalled at each iteration. This would take far too long in practice.

Type 2. This filter only has to perform the block summation each time the block is filled. A simple accumulator can perform this function. The number of samples in a block can be varied according to the filter characteristic required. Generally speaking the lower the cut-off frequency required, the longer the block length.

4.2 The Transversal Filters

There is one such filter per channel. The delay line for each is synthesized by a bipolar r.a.m. which can accept samples and issue them at about 10 MHz. In one iteration of the filter, which may be as little as 5 µs, it is necessary to store a new sample (from the decimation filter) and to read all previous entries to the 'delay line'. In this instrument there are 100. The r.a.m. and its addressing mechanism is split so that separate sub-sections work concurrently to achieve the desired throughput. As with the type 1 decimation filter, a sliding r.a.m. address mechanism is used except that all r.a.m. contents are read each iteration rather than just one. Because there are only 50 different transversal filter coefficients (one per pair of delayed signal samples), pairs of samples are read from the r.a.m. and summed before being weighted (multiplied) by the appropriate coefficient. This saves about 40% of processing time because multiplication is relatively slow. Although most of the transversal filter works to 12 bits precision, the accumulator which sums pairs of delayed, weighted samples works to 18 bits. This prevents bias errors and allows the use of coefficients, not of just 12-bit precision but 12 plus a 2-bit quaternary scale factor. (Many of the coefficients required away from the centre of the 'delay line' are very small indeed and without a 'quasi-floating-point' representation would be recorded as zeros).

4.3 The Linear Interpolator

This unit takes pairs of contiguous samples from the transversal filter and uses them to determine the slope of the waveform between them. This is then used to form samples at 5 MHz which will fit between the two samples first mentioned. (Fig. 4). The gain of the linear interpolator (and for that of the decimation filters) depends on the bandwidth chosen. It is not a simple power of two. A multiplier appears in the signal path in the system purely to equalize gain between bandwidth settings. Its value is simply determined by the current bandwidth setting.

5 Microcomputer Functions

This transposer makes use of a combination of hard-wired logic techniques to perform the signal processing functions and a connected microcomputer to perform management tasks. The latter, of course, put very little pressure on the microcomputer in terms of speed and capacity.

The functions of the micro are basically as follows:

(i) Provide an operator interface in order to simplify setting up of the signal processor parameters

(ii) Provide alarms processing and any possible self-check facilities

(iii) Provide an interface between the signal processor and other equipment, either via one of the standard buses (such as IEEE 488) now in common use, or by some special, in-house, system.

The provisions (ii) and (iii) are fairly self-evident. However, there are several ways in which this instrument can operate—in particular how incoming and outgoing centre frequencies can be set up. In some applications, it would have been in order to use ordinary thumbwheels or a decimal keypad to set up the 7 digit frequencies. The latter is provided in this instrument. However, under a number of circumstances, it is preferable to be able to increment or decrement frequency by possibly small amounts, rather than to make changes on a radical basis. To allow this mode as well, the instrument has a 'joystick' control on the front panel. According to its position (it is sprung to centre zero), a particular digit out of the 7 will increment or decrement according to whether the stick is pulled or pushed. The significance of the digit adjusted is greater for greater stick movements. The rate of change can be altered but is found to be best set between 1 and 2 changes per second. Overflow and underflow of any digit automatically sends carries/borrows to the next significance. The microcomputer interrogates the stick position by comparing the voltage from it with an internally generated ramp and adjusts the set frequency accordingly. It has been found useful to incorporate a non-linear law into the stick interrogation so that, for instance, any slack in the spring zero mechanism can be taken up.

An INTEL 8080 microprocessor is used in a largely conventional μ/r.o.m./r.a.m./I–O configuration. However there are several points of note. The frequency gain and other operational parameters for the signal processor require a total of around 100 bits of information. The majority of this can be linearly mapped into r.o.m. and accessed according to parameter value. In addition, about a dozen gas discharge displays are driven by the microcomputer. It was found most economical to group both parameters and display outputs into a single, long bit serial field. This saves a considerable number of I/O ports. By having display information at the far end of the serial chain, a ready indication is available to the operator of the correct insertion of any parameter.

A serial shifting error is bound to cause an easily recognized display anomally.

This system was developed and debugged using a microcomputer development system, assembly language programming and an in-circuit emulator. During development, it was found convenient and quicker to be able to separate the signal processor and microcomputer. They were constructed in a card frame and as a bolt-on adjunct to the front panel respectively. Whilst separated, a waveform/constants generator was connected to the signal processor to emulate a source of parameters. A surrogate shift register was inserted in the middle portion of the microcomputer's serial bit chain to make up the correct shift register length. This both allowed the display to operate normally and made the computer parameters available for ready inspection.

6 Performance

The instrument is undergoing early trials at present and those sections which have been tested accord well with software simulations and predictions. Out-of-band rejection figures for the decimation filters and linear interpolators are well within the specified targets.

This instrument is very flexible and has a performance which probably exceeds many specifications. However, with the experience gained in its design and realization, many, probably less expensive, variations could be produced. These might use less bandwidths or have a narrower frequency range or slacker spurious signal rejection characteristic.

7 Conclusions

A general-purpose frequency transposer/spectrum pre-processor has been described which has a wide range of applications. Most of these fall within the analogue and digital communications field but also in general data handling and logging.

8 Acknowledgment

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Indexing Terms: Electric filters, bandpass, Surface acoustic waves, Radio systems, mobile

Surface acoustic wave filters for use in mobile radio

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

SUMMARY

Surface acoustic wave filters have not, for a number of reasons, become widely used in mobile radio equipment to date. A brief resumé of the historical development of receiver architecture is presented and the design of two receiver systems is described in which the properties of surface acoustic wave filters are used to best advantage: a double conversion superheterodyne portable radio telephone receiver and a synchrodyne paging receiver. The use of surface acoustic wave filters in these receiver systems is seen to offer advantages with regard to size and power consumption.

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1 Introduction

The design of receiver systems is influenced, to a very great extent, by the performance of the filters that are available. Surface acoustic wave (s.a.w.) filters,[1] operating in the frequency range of 10 MHz to 1·5 GHz, can offer frequency response characteristics that are difficult or impossible to obtain with conventional filters. This has made them popular for a number of applications in both the consumer and professional markets, ranging from intermediate-frequency filters for television,[2] now in large-scale production, to filters of various types for advanced radar systems. S.a.w. filters have not, however, become widely used in mobile radio equipment, as many of the characteristics that have made them popular for other systems are of little merit here and they do not provide attractive direct replacements for the conventional filters used in such equipment at present. In consequence, the design of such receivers must be modified if best advantage is to be taken of their properties.

In this paper, after a brief resumé of the historical development of receiver architecture, the advantageous use of s.a.w. filters is considered in two receiver systems in which there are severe constraints on cost, size and power consumption: a receiver in a portable radio telephone and a receiver for the British Telecom National Paging system. No description of the theory and design of s.a.w. filters is offered, as this has already been described elsewhere, though their performance capabilities are presented in an Appendix.

2 Historical Development of Receiver Architecture

The earliest broadcast radio receivers rarely consisted of more than a first-order filter and a detector.[3] The adjacent channel rejection of this type of receiver was clearly very poor, but was adequate when only a small number of transmitters were in operation. However, as broadcasting gained in popularity and the number of transmitters increased, the spectrum became more congested and the problem of how to obtain sufficient selectivity became severe. The controlled use of positive feedback[4] was found to give improved channel selectivity, but only at the expense of attenuation of the higher frequency components of the modulation of the wanted channel: the need for higher-order filtering was thus identified.

The tuned radio frequency, or t.r.f. receiver[5] provided an increase in selectivity by the use of a cascade of radio-frequency amplification stages and filters before the detector, but problems associated with the simultaneous tuning of the filters proved to be very serious in practice. Nevertheless, for fixed frequency applications, this type of receiver did have advantages and was popular in early broadcast television receivers until these were required to be readily tunable by the user.

In the superheterodyne receiver,[6] frequency conversion of the wanted channel, by means of a mixer and local oscillator, to a fixed intermediate frequency before detection enabled essentially all of the selectivity to be obtained from a high-order fixed filter which did not then suffer from the problems experienced in the t.r.f. receiver. This type of receiver did, however, suffer from the problem of image reception since, in general, two

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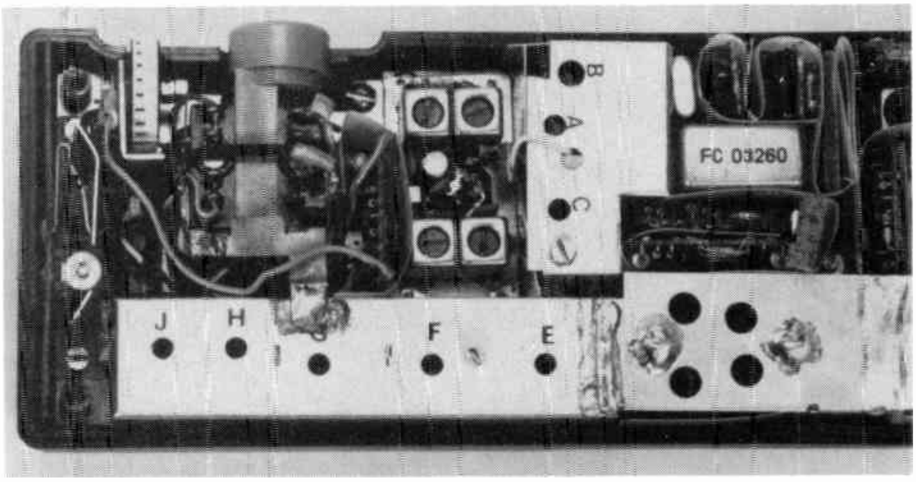

Fig. 1. Typical portable radio telephone. (Approx. 4/5ths actual size).

incoming signals would be converted to the i.f. for each setting of the local oscillator frequency. This called for front-end filtering, before the mixer, to ensure that reception of only the wanted channel took place. The frequency at which the i.f. filtering was performed was, therefore, constrained primarily by the image rejection that could be obtained from the front end filters which were available. If sufficient selectivity could then not be obtained from the filters which were available at this i.f., a second frequency conversion would be performed to a second i.f. The first i.f. filter was then called upon to provide image rejection for the second intermediate frequency. In principle, this process of frequency conversion and filtering would be continued until an i.f. was reached at which sufficient selectivity could be achieved by the filters which were available, though it was rare for anything more than triple conversion to be required.

The superheterodyne receiver rapidly became established as the most popular type of receiver since, as the frequency of broadcast and other transmissions increased, it became the only form of receiver in which the desired degree of selectivity could be achieved at all. The superheterodyne receiver thus came to be used in applications calling for an untuned receiver, despite the fact that it was originally developed to overcome the problem experienced in obtaining sufficient selectivity in tuned receivers.

The synchrodyne receiver,[7] a direct-conversion receiver in which essentially all of the selectivity was achieved at audio frequencies, never proved to be a serious threat to the superheterodyne receiver, being conceived decades before it could be effectively realized. The direct-conversion type of receiver is now seen as offering a real challenge to the superheterodyne receiver for some applications.

3 Portable Radio Telephone Receiver

Figure 1 shows the internal arrangement, with the exception of the rechargeable battery, of a typical portable radio telephone operating in the 440 MHz to 470 MHz frequency band.

The receiver is designed to operate on a single channel and, despite the apparent simplicity, it would not be attractive to use a t.r.f. receiver circuit with fixed filters offering a few tens of kilohertz of bandwidth at this frequency if they were available. This is because it would require the manufacture of a small number of filters at a large number of different frequencies, which would be uneconomic. A double-conversion superheterodyne receiver circuit is used instead, with intermediate frequencies of 23·455 MHz and 455 kHz, the block diagram of which is shown in Fig. 2.

Helical resonators offer the narrowest bandwidth of any conventional type of tunable filter at 440 MHz to 470 MHz and four of them are used, in two pairs, as

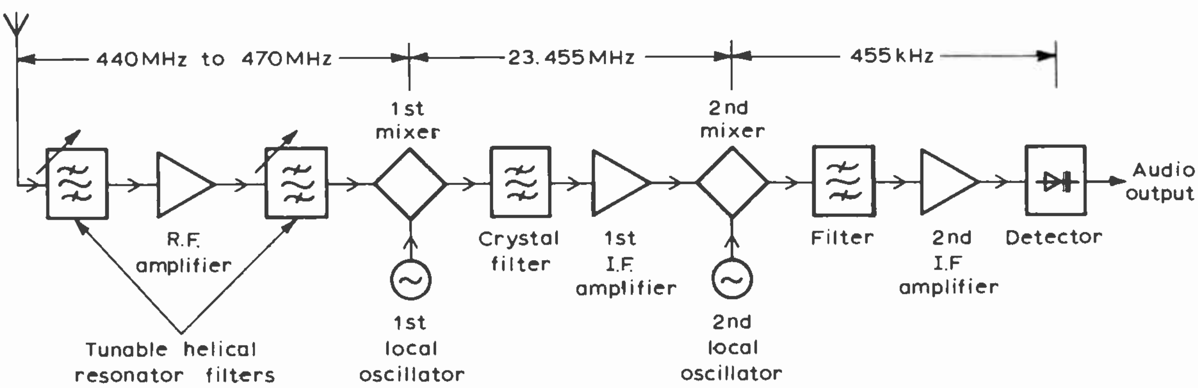

Fig. 2. Block diagram of portable radio telephone.

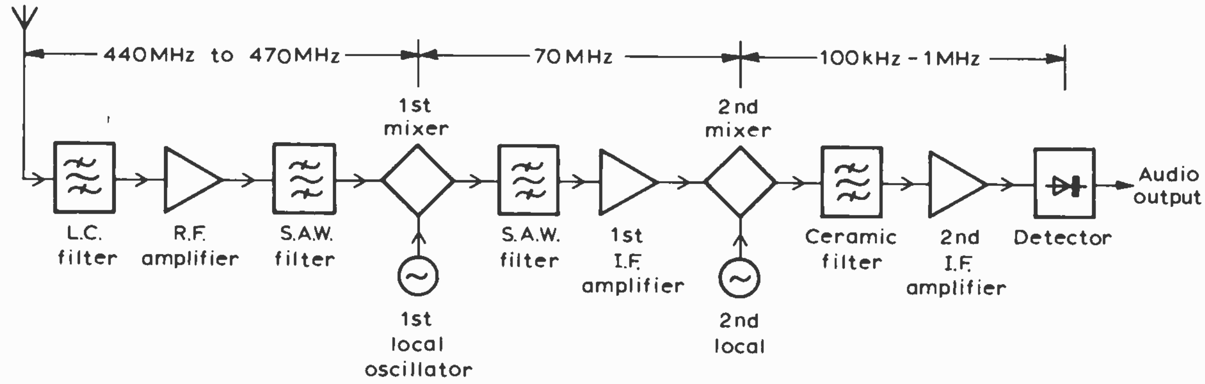

Fig. 3. Block diagram of modified portable radio telephone using s.a.w. filters.

front-end filters. They would, typically, provide a 4 MHz bandwidth and 70 dB image rejection for a first i.f. which, as they are not particularly selective, has to be at the rather high frequency of 23·455 MHz. The first i.f. filter would, typically, be an eighth-order crystal filter composed of four monolithic duals and, with a 25 kHz channel spacing, provides essentially all of the adjacent channel rejection of, typically, 70 dB. In this receiver, the second stage of frequency conversion is performed primarily to reduce power consumption and ease the task of detection, but requires the crystal filter to provide 70 dB image rejection for a second i.f. of 455 kHz. This can be achieved, though with some care as crystal filters tend to suffer from spurious responses.

The following points should be noted about this receiver:

(i) The helical resonators must be aligned to the particular section of the 440 MHz to 470 MHz frequency band required.

(ii) The helical resonators and the crystal filter are expensive to produce and are relatively bulky components. They are, in particular, rather tall, thus necessitating complex printed circuit board arrangements to achieve a compact layout.

The use of s.a.w. filters can bring about some improvement to these points and Fig. 3 shows the block diagram of a suitably modified receiver in which this has been achieved.

The helical resonators have been replaced by a wide-band, low-loss transversal s.a.w. filter offering frequency response characteristics which could not be obtained with conventional filters. This filter, whose frequency response is shown in Fig. 4, has a sufficiently wide bandwidth to cover the complete 440 MHz to 470 MHz frequency band and a sufficiently narrow transition bandwidth to permit the use of a 70 MHz first i.f. The insertion loss is 8 dB and could probably be reduced somewhat, whilst 65 dB of image rejection is obtained which, being limited at present by electromagnetic breakthrough, could also be improved. The input and output impedances are 50 Ω, no matching or phase adjusting inductors are required and the filter is packaged in a TO-5 can. Figure 5 shows the filter, with the top of its package removed, adjacent to the helical resonators it replaces. The dramatic reduction in volume is clearly apparent.

The 70 MHz first i.f. filter, whose frequency response is shown in Fig. 6, is a second-order narrowband s.a.w. resonator filter which, with a 25 kHz channel spacing,

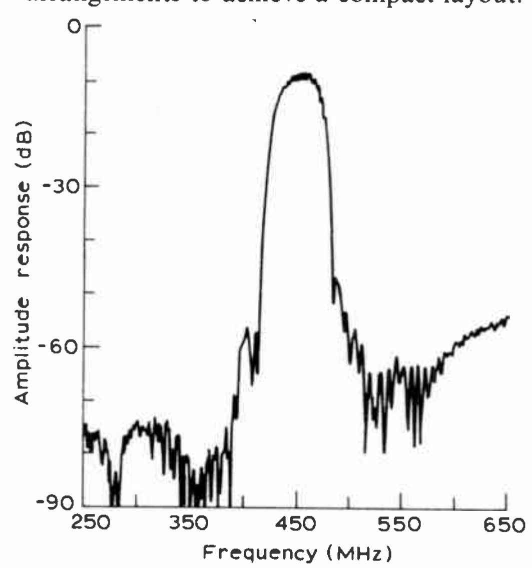

Fig. 4. Frequency response of wide-band, low-loss transversal s.a.w. filter.

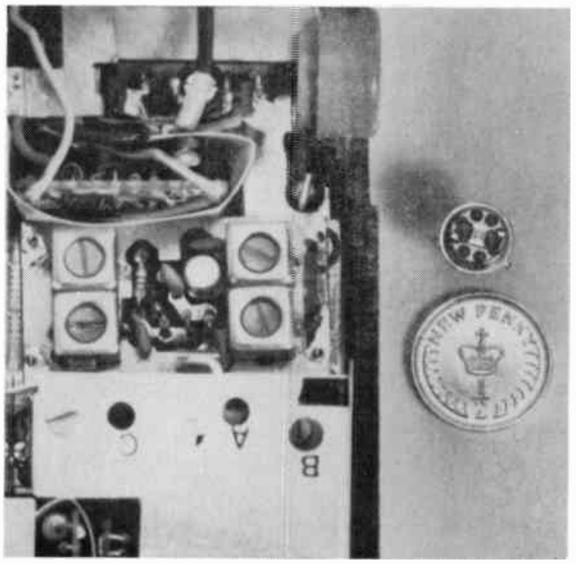

Fig. 5. Comparison of helical resonators with wide-band, low loss transversal s.a.w. filter. (Approximately actual size).

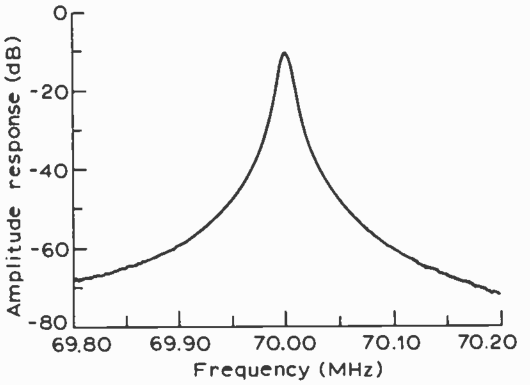

Fig. 6. Frequency response of second-order, narrowband s.a.w. resonator filter.

provides 25 dB of adjacent channel rejection and at least 60 dB image rejection for a second i.f. of between 100 kHz and 1 MHz. The insertion loss is 10 dB, having been deliberately increased from the minimum possible in the interests of size, which could be reduced still further. A remaining 45 dB of adjacent channel rejection would be provided by a ceramic filter at the second i.f., which could be achieved in a very compact and economical manner at 455 kHz. The input and output impedances of the 70 MHz filter are 50 Ω, again no matching or tuning inductors are required, and it is packaged in a dual in-line can.

Figure 7 shows the filter with the top of its package removed. The difference in volume occupied by this filter with its attendant 455 kHz ceramic filter and the crystal filter they replace is small, but the much lower height of the s.a.w. filter makes it rather attractive.

The advantages of this approach are, therefore:

(i) No alignment.
(ii) Small size and, especially, low height.

4 Paging Receiver

Narrowband s.a.w. resonator filters, of the type used as an i.f. filter in the portable radio telephone receiver which has just been described, can also be economically used as front-end filters in receivers for any system in which there are a large number of receivers and few channels. An excellent example of this type of system is

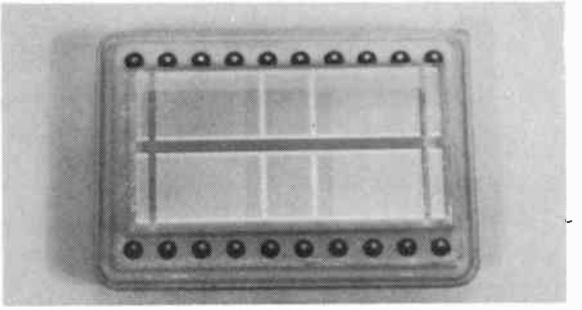

Fig. 7. Second-order, narrowband s.a.w. resonator filter. (Approx. twice actual size).

the British Telecom National Paging system operating at approximately 153 MHz. A conventional paging receiver would use a double conversion superheterodyne circuit but, as low cost, small size and low power consumption are of paramount importance, alternative receiver architectures of the direct-conversion type are under consideration.[8] Figure 8 shows how the properties of s.a.w. resonators have been exploited as both a front-end filter and as an oscillator control element to make an experimental synchrodyne paging receiver.

No receiver system, except for one in which all of the adjacent-channel rejection is achieved before the incoming signal encounters any active devices, is free from the problem of very strong adjacent-channel signals causing interference by virtue of the inherently limited signal-handling capabilities of active devices. This problem is usually relieved by arranging for all active devices which are used before sufficient adjacent-channel rejection has been achieved to operate in as linear a manner as is possible. This inevitably involves the consumption of much power, which is particularly unattractive in a paging receiver. In the experimental synchrodyne paging receiver, a low-loss, third-order narrowband s.a.w. resonator filter is used to provide some adjacent-channel rejection before the phase comparator, enabling its power consumption to be considerably reduced, whilst maintaining a satisfactory overall performance. Figure 9 shows the frequency response of this filter. The insertion loss is 5 dB and, with a channel spacing of 25 kHz, there is 22 dB of adjacent-channel rejection. The response shape is close to

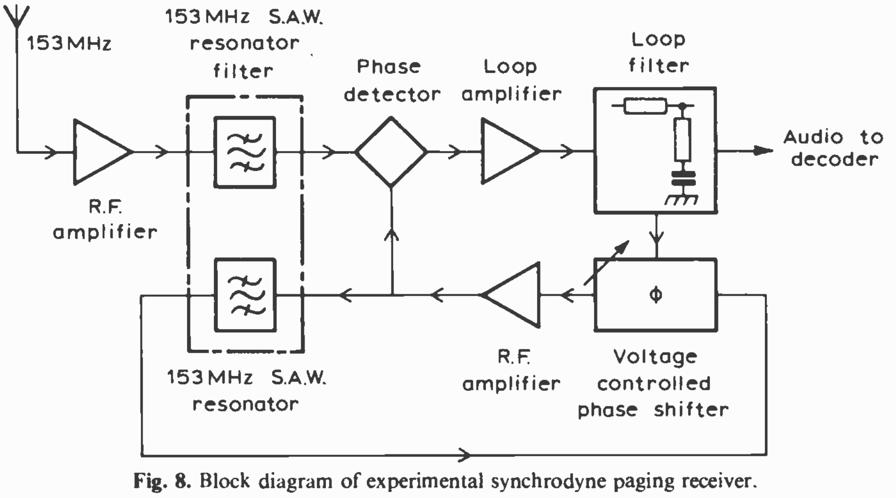

Fig. 8. Block diagram of experimental synchrodyne paging receiver.

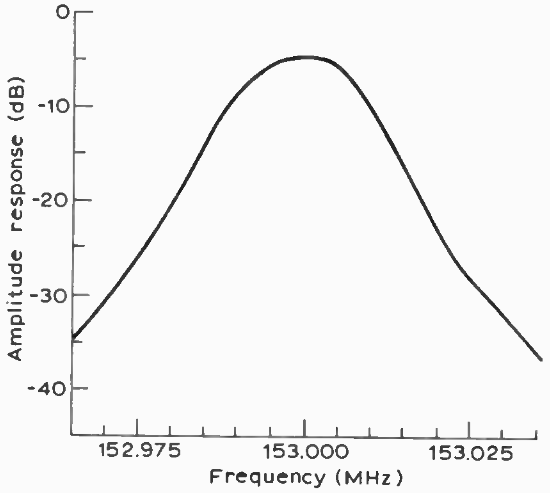

Fig. 9. Frequency response of low-loss, third-order narrowband s.a.w resonator filter.

Gaussian, thus ensuring good transient performance in a system which uses digital modulation. The input and output impedances are 50 Ω, once again no matching or tuning inductors are required, and the filter is packaged in a dual in-line can. Figure 10 shows the filter with the top of its package removed.

The local oscillator which, in a synchrodyne receiver, operates at essentially the same frequency as the incoming signal, is very conveniently realized using a fourth s.a.w. resonator, an r.f. amplifier and a voltage-controlled phase shifter which, in the prototype, was realized as a low-pass-filter Π network with variable capacitance diodes as the shunt arms. A 10 kHz oscillator pulling range was achieved whilst operating from a 3 V supply. The s.a.w. resonator could be incorporated, at no extra cost, on the same substrate as the front-end filter. Electromagnetic coupling between them, measured at −90 dB, would not be the major origin of re-radiation of the local oscillator signal, a potential problem with this type of receiver. A re-radiated level of, typically, −50 dB relative to 1 mW of local oscillator power could be achieved, the phase comparator, being of the double-balanced-type

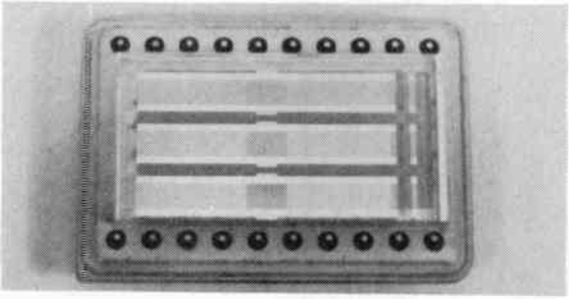

Fig. 10. Low-loss, third-order narrowband s.a.w. resonator filter. (Approx. twice actual size).

contributing 30 dB of suppression, reverse transmission through the r.f. amplifier 15 dB and the front-end filter another 5 dB. This oscillator would provide a standard of frequency stability comparable with that offered by a crystal oscillator and multiplier, but at a fraction of the cost.

In prototype, a recovered audio output of 700 mV rms was obtained, sufficient to drive directly an integrated circuit for decoding of the POCSAG† code.[9] Although this simple circuit would require some refinement in order to be made into a practicable paging receiver, the possibility of making a very small module in this way can be readily appreciated.

5 Conclusion

It has been shown that the use of s.a.w. filters in mobile radio, rather than leading to a dramatic improvement in technical performance, can play a valuable part in the reduction of size and power consumption in certain types of receiver. In a portable radio telephone, replacement of helical resonators and a crystal filter by two s.a.w. filters and a ceramic filter permits a useful reduction in volume and, in particular, height to be achieved. In a paging receiver for the British Telecom National Paging system the properties of s.a.w. resonators can be exploited to provide the basis for an extremely compact receiver with low power consumption.

† Post Office Code Standardization Advisory Group.

6 Appendix—S.A.W. Filter Capabilities

Table 1. Transversal filters

Centre frequency range	10 MHz–1·5 GHz
Minimum bandwidth (−3 dB)	100 kHz or 0·2% of centre frequency, whichever is greater
Maximum bandwidth (−3 dB)	100% of centre frequency
Minimum transition bandwidth (−50 dB to −3 dB)	100 kHz or 0·2% of centre frequency, whichever is greater
Group delay	1 µs–5 µs
Group delay ripple	<2% of group delay, peak-to-peak
Insertion loss	15 dB–25 dB; can be reduced by the use of low-loss-filter structures
Passband amplitude ripple	<0·5 dB, peak-to-peak
Stopband level	50 dB close to passband; 70 dB further away from passband
Package	TO-8

Table 2. Resonator filters

Centre frequency range	50 MHz–1·5 GHz
Minimum bandwidth (−3 dB)	0·01% of centre frequency at 50 MHz; 0·02% of centre frequency at 500 MHz
Maximum bandwidth (−3 dB)	0·05% of centre frequency with inherently good temperature stability; 1·0% with external temperature stabilization
Insertion loss	<6 dB
Stopband level	60 dB
Size	25 mm by 15 mm at 50 MHz; 7·5 mm by 7·5 mm at 500 MHz.

The limits quoted in these tables are typical and should not be regarded as rigid constraints. Furthermore, it is not the intention to suggest that all of the extrema could be achieved simultaneously. For example, a transversal filter at 10 MHz with 100 kHz of bandwidth could not be expected to fit into a TO-8 can and transversal filters with bandwidths exceeding 50% of the centre frequency are only realizable with high insertion loss, whilst a resonator filter at 500 MHz with a very narrow bandwidth would have an insertion loss greater than 6 dB.

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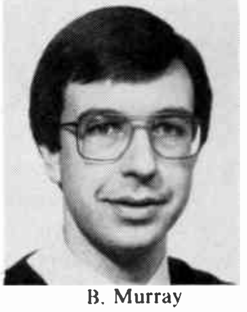
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Indexing Terms: Data transmission, Equalizers, attenuation, Noise performance, Computer simulation

Non-linear equalizers having an improved tolerance to noise

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SUMMARY

The paper describes a simple development of a conventional non-linear equalizer which involves only changes to the equalization process itself and can, under certain conditions, give a useful improvement in tolerance to additive white Gaussian noise. The technique is particularly effective when binary data-symbols are transmitted and there is severe amplitude distortion in the received signal. The optimum design of the modified equalizer together with an interesting sub-optimum arrangement, for the application of greatest practical interest, are derived first and then approximate estimates of their tolerances to additive white Gaussian noise. Finally, results are presented of computer-simulation tests, with binary signals transmitted over various channels in the presence of additive white Gaussian noise, and the results are compared with the corresponding theoretical predictions.

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1 Introduction

In the transmission of digital data at relatively high rates over a linear channel that introduces significant intersymbol interference into its output signal, it is common practice to remove this interference by means of a non-linear (decision-feedback) equalizer in the receiver.[1–8] Such an equalizer often achieves a useful advantage in tolerance to additive noise over the corresponding linear equalizer,[8] the tolerance to noise being defined as the signal/noise ratio for some given (generally quite small) error rate in the detected data symbols. In many practical applications the equalizer is adjusted adaptively to minimize the mean-square error in the equalized signal at its output, this signal being fed to a simple threshold-level detector.[1,2,4,6,8] At very high signal/noise ratios, such an equalizer removes nearly all the intersymbol interference that is introduced by the channel and so it acts as the approximate inverse of the channel.[8] Another arrangement of the non-linear equalizer that has been extensively studied minimizes the mean-square error in the equalized signal, subject to the perfect equalization of the channel (removal of all intersymbol interference), so that the equalizer is now the exact inverse of the channel, regardless of the signal/noise ratio.[3,5,7,8] A common feature of all these equalizers is the attempt to reduce the level of the intersymbol interference to a low or zero value. It has not however been shown that this adjustment of the equalizer in every case leads to the lowest error rate in the detected data symbols, obtainable with the corresponding equalizer.

In the applications to be studied here a real-valued baseband digital data-signal is transmitted over a time-invariant linear baseband channel. The latter could involve a coaxial cable or 600 Ω pair or else it could include a bandpass transmission path such as a telephone circuit, together with a linear modulator at the transmitter and a linear demodulator at the receiver. With the transmission of a baseband data-signal having 4 or more levels, and where the intersymbol interference is not too severe, a conventional non-linear equalizer gives a good tolerance to additive noise which is often no more than some two or three decibels below that obtainable with the corresponding maximum-likelihood detector.[8–12] Clearly the design or adjustment of the equalizer must here be close to that giving the best available performance of the equalizer. However, with the transmission of a binary signal and when there is severe amplitude distortion of the transmitted signal, the tolerance to noise may be very considerably below that obtainable with a maximum-likelihood detector.[8,13,14] A recent theoretical investigation into the reason for this large degradation in performance has revealed that, under the particular conditions, the design or adjustment of the conventional non-linear equalizer is no longer optimum. In other words, in the presence of additive white Gaussian noise, at high signal/noise ratios, the non-linear equalizer that achieves the accurate equalization of the channel no longer even approximately minimizes (for an equalizer) the error rate in the detected data signal. In order to take account of and hence to correct for this effect, in the particular case

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where a linear channel has introduced severe amplitude distortion into a binary baseband data-signal, the design of the non-linear equalizer for the given channel is now modified, essentially by relaxing the requirement for the exact or even approximate equalization of the channel and hence by using a fundamentally different criterion in the optimization process involved in the design.

The paper is concerned with the basic theory of the modified (optimized) equalizer and describes both the conditions under which it differs from the conventional equalizer and the simple modification by means of which it is derived from the latter. A sub-optimum version of the modified equalizer is also studied. The conventional equalizer is here taken to be the ideal non-linear equalizer that minimizes the mean-square error in the equalized signal subject to the accurate equalization of the channel. The received signal is assumed to be a binary real-valued baseband signal that has been subjected to severe amplitude distortion. The baseband signal may itself have been fed over the transmission path or else it may have been derived by the linear demodulation of a received suppressed-carrier a.m. signal. The optimum design of the modified equalizer is derived first and then an approximate expression for its tolerance to additive white Gaussian noise. The latter is used to predict the performance of the modified equalizer over a wide range of channels, and hence to compare its performance with that of a conventional non-linear equalizer. The whole procedure is next repeated for an interesting sub-optimum design of the modified equalizer. Finally, results are presented of computer-simulation tests, with binary signals transmitted over various channels in the presence of additive white Gaussian noise, and the results are compared with the corresponding theoretical predictions.

2 Conventional Equalizer

The model of the data-transmission system when using a conventional non-linear (decision-feedback) equalizer is shown in Fig. 1. This is a synchronous serial system in which the information transmitted is carried by the binary data-symbols $\{s_i\}$, where

$$s_i = \pm 1, \tag{1}$$

the $\{s_i\}$ being statistically independent and equally likely to have either binary value. The impulses at the input to the channel are regularly spaced at intervals of T seconds and form a sequence of binary polar signal-elements. It is assumed that $s_i = 0$ for $i \leqslant 0$, so that $s_i\delta(t-iT)$ is the ith transmitted signal-element.

The linear baseband channel includes the transmitter output filter, a linear baseband transmission path and the receiver input filter. The transmission path could, of course, itself include a linear modulator at the transmitter, a bandpass transmission path such as a telephone circuit, and a linear demodulator at the receiver. The resultant channel has an impulse response $u(t)$ and adds a Gaussian noise waveform $v(t)$ to the output data signal to give the output waveform

$$q(t) = \sum_i s_i u(t-iT) + v(t) \tag{2}$$

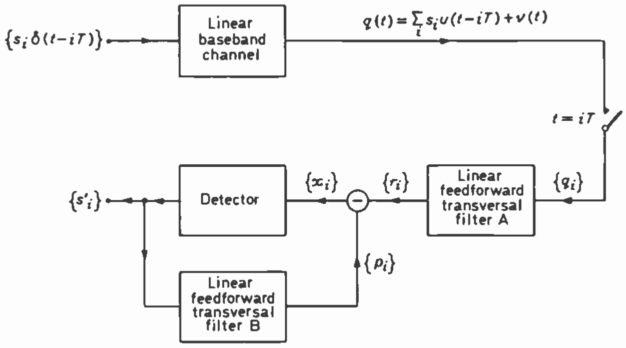

Fig. 1. Model of data-transmission system with a conventional non-linear equalizer.

where $q(t)$, $u(t)$ and $v(t)$ have only real values. $u(t-iT)$ is taken to be time invariant over any one transmission, in the sense that its shape is not a function of i, and to have a finite duration, at least for practical purposes. $v(t)$ is a stationary Gaussian random process with zero mean, variance σ^2 and such that any two samples, at a time interval that is a multiple of the sampling interval T, are uncorrelated and therefore statistically independent Gaussian random variables. It is assumed that $v(t)$ originates from white Gaussian noise that is added to the data signal at the output of the bandpass or baseband transmission path and therefore at the input to the receiver filter within the linear baseband channel in Fig. 1. The received waveform $q(t)$ is sampled at the time instants $\{iT\}$ to give the samples $\{q_i\}$ which are fed to the linear feedforward transversal filter A, whose corresponding output samples are the $\{r_i\}$.

It is assumed that the non-linear equalizer in Fig. 1 is adjusted to maximize the signal/noise ratio in the equalized signal, subject to the accurate equalization of the channel. The linear feedforward transversal filter A uses a process of pure phase equalization (which, of course, does not change the amplitude distortion introduced by the channel) and introduces no gain or attenuation into the received signal. It has been shown that under these conditions the resultant response of the channel and linear filter is minimum phase, with the energy of each received signal-element concentrated towards the start of that element.[8] Furthermore, the noise components in the output samples $\{r_i\}$ from the filter A have exactly the same statistical properties as the noise components in the input samples $\{q_i\}$.

Suppose now that the resultant sampled impulse-response of the linear baseband channel and filter A (Fig. 1) is given by the time-invariant vector

$$Y = [y_0 \ y_1 \ \ldots \ y_g] \tag{3}$$

where the delay in transmission, other than that involved in the time dispersion of the signal, is neglected, so that $y_0 \neq 0$ and $y_i = 0$ for $i < 0$ and $i > g$. The filter A here introduces the linear orthogonal transformation which maximizes the magnitude of y_0.[8] The sample r_i at the output of the filter A, at time $t = iT$, is

$$r_i = \sum_{h=0}^{g} s_{i-h} y_h + w_i \tag{4}$$

where the $\{w_i\}$ are statistically independent Gaussian

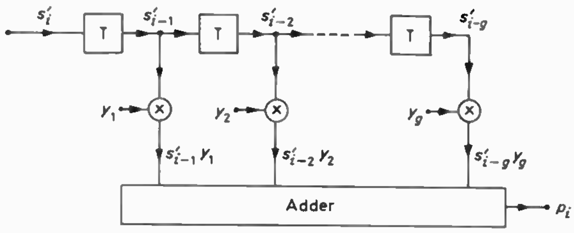

Fig. 2. Linear feedforward transversal filter B.

random variables with zero mean and fixed variance σ^2, and the probability density function of w_i, at a sample value w, is

$$\frac{1}{\sqrt{2\pi\sigma^2}}\exp\left(-\frac{w^2}{2\sigma^2}\right).$$

In the conventional arrangement of the non-linear equalizer, the linear feedforward transversal filter B is as shown in Fig. 2. It has g taps with gains equal to $y_1, y_2, \ldots, y_g$, and operates on the detected data-symbols $\{s'_{i-h}\}$ at the detector output, such that, at time $t = iT$ and just prior to the detection of s_i, the output signal from the filter B is

$$p_i = \sum_{h=1}^{g} s'_{i-h}y_h. \tag{5}$$

This is subtracted from r_i to give the equalized signal

$$x_i = r_i - p_i \tag{6}$$

at the detector input. With the correct detection of the data-symbols $s_{i-1}, s_{i-2}, \ldots, s_{i-g}$,

$$x_i = s_i y_0 + w_i \tag{7}$$

and, regardless of the correct detection of these symbols, s_i is now detected as its possible value s'_i such that $s'_i y_0$ is closest to x_i.

Assume for the moment that equation (7) holds and let

$$Q(e) = \int_e^{\infty} \frac{1}{\sqrt{2\pi}} \exp\left(-\tfrac{1}{2}f^2\right) \mathrm{d}f \tag{8}$$

where e and f are scalar quantities. Then it can be seen from equations (7) and (8) that, with the correct detection of $s_{i-1}, s_{i-2}, \ldots, s_{i-g}$, the probability of correct detection of s_i from x_i is

$$P_3 = Q(|y_0|/\sigma) \tag{9}$$

where $|y_0|$ is the absolute value (modulus) of y_0. P_3 is a lower bound to the actual error rate and, at low error rates of below 1 in 10^5, it can be taken as an approximate estimate of this rate.[8]

3 Modified Equalizer

In the absence of any amplitude distortion or attenuation in the received data-signal, $y_0 = 1$ and $y_i = 0$ for $i = 1, 2, \ldots, g$, any phase distortion in the received signal being removed by the filter A and therefore not affecting the values of $y_0, y_1, \ldots, y_g$.[8] The greater the level of the amplitude distortion in the received signal, the greater the values of $|y_1|, |y_2|, \ldots, |y_g|$ relative to $|y_0|$. When the distortion is severe and involves considerable bandlimiting of the data signal, the latter being transmitted at a rate close to the Nyquist rate, it is likely that

$$|y_1| > |y_0|. \tag{10}$$

This condition is often satisfied in applications such as the very high speed transmission of digital signals over long lengths of a coaxial cable, or where a digital signal is being transmitted at the highest practical rate over any bandlimited channel. The condition is also often observed in data-transmission systems operating at 4800 or 9600 bit/s over the poorer telephone circuits. (The baseband digital signals here may, of course, be complex valued.) It will now be shown that, when equation (10) holds (together with the other conditions described in Section 2) and when binary data symbols are transmitted (equation (1)), it is possible to achieve an improved tolerance to noise through the appropriate modification of the non-linear equalizer.

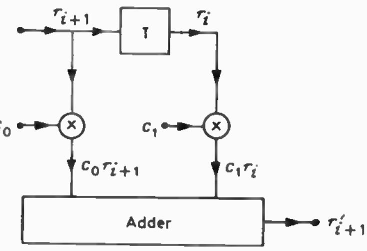

Fig. 3. Linear feedforward transversal filter C.

The equalizer is modified by inserting the linear feedforward transversal filter C, as shown in Fig. 3, between the filter A and the subtractor circuit in Fig. 1. In practice, of course, the filter A could be modified into the equivalent resultant filter, but it is simpler here to consider the two separate filters. The output sample from the filter C, at time $t = (i+1)T$, is

$$r'_{i+1} = c_0 r_{i+1} + c_1 r_i \tag{11}$$

and the resultant sampled impulse-response of the channel and filters A and C is given by the $(g+2)$-component row vector Y' whose $(i+1)$th component is

$$y'_i = y_i c_0 + y_{i-1} c_1 \tag{12}$$

where, of course, $y_i = 0$ for $i < 0$ and $i > g$. Thus the sample at the output of the filter C, at time $t = (i+1)T$, is

$$r'_{i+1} = s_{i+1}y'_0 + s_i y'_1 + \ldots + s_{i-g}y'_{g+1} + w'_{i+1} \tag{13}$$

where

$$w'_{i+1} = w_{i+1}c_0 + w_i c_1. \tag{14}$$

Clearly, w'_{i+1} is a Gaussian random variable with zero mean and variance

$$\eta^2 = \sigma^2(c_0^2 + c_1^2) \tag{15}$$

the $\{w'_i\}$ being correlated.

The linear feedforward transversal filter B has g taps, as before (Fig. 2), but now with tap gains equal to $y'_2, y'_3, \ldots, y'_{g+1}$ in place of $y_1, y_2, \ldots, y_g$, respectively. At time $t = (i+1)T$ the detector has just determined s'_{i-1} and the output sample from the filter B is

$$p'_{i+1} = s'_{i-1}y'_2 + s'_{i-2}y'_3 + \ldots + s'_{i-g}y'_{g+1}, \tag{16}$$

which is subtracted from r'_{i+1} to give the equalized signal

$$x'_{i+1} = r'_{i+1} - p'_{i+1}, \tag{17}$$

which is fed to the detector. With the correct detection of $s_{i-1}, s_{i-2}, \ldots, s_{i-g}$,

$$\begin{aligned} x'_{i+1} &= s_{i+1}y'_0 + s_i y'_1 + w'_{i+1} \\ &= s_{i+1}y_0c_0 + s_i(y_1c_0 + y_0c_1) + w'_{i+1}. \end{aligned} \tag{18}$$

The detected data-symbol s'_i is taken to be the possible value of s_i such that $s_i(y_1c_0 + y_0c_1)$ is closest to x'_{i+1}. $s_{i+1}y_0c_0$ is an intersymbol-interference component in x'_{i+1} and is equally likely to have either value $\pm|y_0c_0|$. It is evident that, for the best tolerance to noise, y_1c_0 and y_0c_1 must both have the same sign which for convenience is taken to be positive. Thus

$$y_1c_0 \geqslant 0, \qquad y_0c_1 \geqslant 0 \tag{19}$$

and the detected data-symbol s'_i is taken as 1 or -1 depending upon whether $x'_{i+1} > 0$ or < 0, respectively. It can be seen from equation (18) that a necessary condition for the correct operation of the equalizer is that

$$y_1c_0 + y_0c_1 > |y_0c_0| \tag{20}$$

which will therefore be assumed.

At high signal/noise ratios, practically all errors in the $\{s'_i\}$ occur when $s_{i+1}y_0c_0$ has the opposite sign to $s_i(y_1c_0 + y_0c_1)$, since under these conditions

$$|s_{i+1}y_0c_0 + s_i(y_1c_0 + y_0c_1)|$$

is minimum, and the Gaussian noise component w'_{i+1} needed to produce an error in s'_i now has the smallest magnitude. Thus, the condition under which nearly all errors in the $\{s'_i\}$ occur is that

$$s_{i+1}y_0c_0 = -s_i|y_0c_0| \tag{21}$$

or

$$x'_{i+1} = s_i(y_1c_0 + y_0c_1 - |y_0c_0|) + w'_{i+1}. \tag{22}$$

Let

$$dc_0 = y_1c_0 - |y_0c_0|, \tag{23}$$

bearing in mind that $y_1c_0 \geqslant 0$ (eqn. (19)). Thus, nearly all errors occur when

$$x'_{i+1} = s_i(dc_0 + y_0c_1) + w'_{i+1}. \tag{24}$$

Since $s_i = \pm 1$ and $dc_0 + y_0c_1 > 0$ (eqns. (1), (20) and (23)), the condition for an error in the detection of s_i from x'_{i+1} is that w'_{i+1} has a magnitude greater than $dc_0 + y_0c_1$ and the opposite sign to s_i. Furthermore, w'_{i+1} has zero mean and variance η^2 (eqn. (15)), and equation (24) is satisfied with a probability of $\frac{1}{2}$. Thus the probability of error in the detection of s_i, at high signal/noise ratios and given the correct detection of $s_{i-1}, s_{i-2}, \ldots, s_{i-g}$, can be taken to be $\frac{1}{2}Q(a)$, where

$$a = \frac{dc_0 + y_0c_1}{\eta} = \frac{dc_0 + y_0c_1}{\sigma\sqrt{c_0^2 + c_1^2}} \tag{25}$$

from equation (15). An approximate estimate of the actual error rate in the $\{s'_i\}$, at high signal/noise ratios (but, of course, not now assuming the necessity of the correct detection of $s_{i-1}, s_{i-2}, \ldots, s_{i-g}$), can be taken to be $Q(a)$.

$Q(a)$ is minimum when a (eqn. (25)) is maximum, which (as may readily be shown from the theory of matched filters[8]) occurs when

$$c_0 = bd, \qquad c_1 = by_0 \tag{26}$$

and b is any positive constant. For equation (26) to hold with $c_0 \neq 0$ it is necessary that in equation (25) $dc_0 > 0$, which means that $y_1c_0 > |y_0c_0|$, from equation (23), so that $|y_1| > |y_0|$ as in equation (10). Under these conditions, and from equations (23) and (26),

$$\frac{c_0^2}{b} = y_1c_0 - |y_0c_0| \tag{27}$$

so that

$$|c_0| = b(|y_1| - |y_0|) \tag{28}$$

and c_0 has the same sign as y_1. c_1 is, of course, given by equation (26). Furthermore,

$$|d| = |y_1| - |y_0| \tag{29}$$

and d has the same sign as y_1. Now, from equations (25)–(29) (where $dc_0 > 0$), the theoretical estimate of the error rate in the $\{s'_i\}$ is

$$P_1 = Q(a) = Q\left(\frac{dc_0 + y_0c_1}{\sigma\sqrt{c_0^2 + c_1^2}}\right), \tag{30}$$

which simplifies to

$$P_1 = Q\left(\frac{\sqrt{(|y_1| - |y_0|)^2 + y_0^2}}{\sigma}\right). \tag{31}$$

If $|y_1| \leqslant |y_0|$, then, from equations (19) and (23), $dc_0 \leqslant 0$, so that a in equation (25) is maximum when $c_0 = 0$ and $c_1 = by_0$. The arrangement now degenerates into a conventional non-linear equalizer and the theoretical estimate of the error rate in the $\{s'_i\}$ becomes P_3 in equation (9). Thus, for any advantage to be gained from the given system it is necessary that equation (10) holds, which will now be assumed.

It follows from equations (9) and (31) that, at a given low error probability, the value of σ for the modified equalizer is approximately

$$\sqrt{(|y_1| - |y_0|)^2 + y_0^2}/|y_0|$$

times that for the conventional equalizer, which means that, at low error rates, the modified equalizer gains an advantage of about

$$20\log_{10}\left(\frac{\sqrt{(|y_1| - |y_0|)^2 + y_0^2}}{|y_0|}\right) = 10\log_{10}\left(\frac{(|y_1| - |y_0|)^2}{y_0^2} + 1\right)\ \text{dB} \tag{32}$$

in tolerance to the Gaussian noise over the conventional equalizer. Figure 4 shows how the theoretical advantage in tolerance to Gaussian noise of the modified equalizer (the optimum system) over the conventional equalizer varies with $|y_1|/|y_0|$.

A sub-optimum but very interesting arrangement of

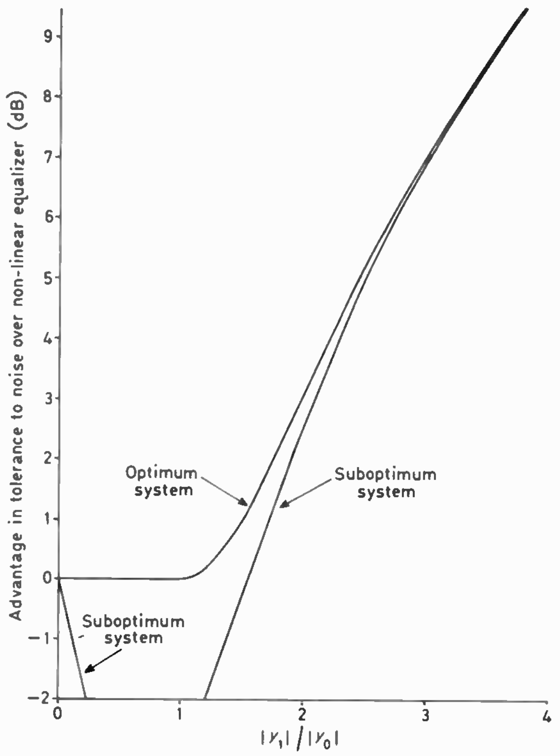

Fig. 4. Theoretically predicted performance of the two modified equalizers at high signal/noise ratios.

the modified equalizer is to set

$$c_0 = by_1, \qquad c_1 = by_0 \tag{33}$$

where b is any positive constant, as before. This ensures that equations (19) and (20) are both satisfied. From equation (30), the theoretical estimate of the error rate in the $\{s_i'\}$ here becomes

$$P_2 = Q\left(\frac{dc_0 + y_0 c_1}{\sigma\sqrt{c_0^2 + c_1^2}}\right) \tag{34}$$

which simplifies to

$$P_2 = Q\left(\frac{y_1^2 + y_0^2 - |y_0 y_1|}{\sigma\sqrt{y_1^2 + y_0^2}}\right). \tag{35}$$

It can be seen from equation (9) that, at low error rates, the advantage in tolerance to noise (measured in dB) gained by this arrangement of the modified equalizer over the conventional equalizer, is about

$$20 \log_{10}\left(\frac{y_1^2 + y_0^2 - |y_0 y_1|}{|y_0|\sqrt{y_1^2 + y_0^2}}\right)$$

which, with some rearrangement of the terms, becomes

$$10 \log_{10}\left(\frac{(|y_1| - |y_0|)^2}{y_0^2} + 1 - \frac{y_0^2}{y_1^2 + y_0^2}\right).$$

The variation of this advantage with $|y_1|/|y_0|$, for the sub-optimum system just described, is shown in Fig. 4. It is evident from equation (32) that the performance of the sub-optimum system (eqn. (33)) is inferior to that of the optimum system (eqns. (26) and (29)), but as $|y_1|/|y_0|$ increases so the advantage of the latter over the former rapidly diminishes. It is also interesting to observe that, when $|y_1|/|y_0|$ is less than about 1·55, the performance of the sub-optimum system becomes inferior to that of the conventional equalizer, with a maximum degradation of about 3·8 dB (not shown in Fig. 4), whereas the optimum system is always at least as good as the equalizer.

A clearer insight into the relationship between the optimum and sub-optimum systems is given by the following simple representation of the results obtained. In the sub-optimum system the filter C is matched to the 2-component vector

$$Y_2 = [y_0 \quad y_1], \tag{36}$$

whereas, in the optimum system, the filter C is matched to the 2-component vector

$$Z_2 = [y_0 \quad (y_1 \pm y_0)] \tag{37}$$

of the minimum Euclidean length (the Euclidean length being defined as the square root of the sum of the squares of the vector components). Thus, in the optimum system,

$$c_0 = b(y_1 \pm y_0), \qquad c_1 = by_0 \tag{38}$$

where b is a positive constant and the appropriate sign is used in $y_1 \pm y_0$.

A theoretical study has been carried out, along the same lines as that in this paper, on two developments of the basic technique described. In the interests of brevity and since the developments are not considered further here, only the more important results of this analysis are now presented.

When

$$|y_1| > |y_0| \tag{39}$$

and

$$|y_2| > |y_1| + |y_0| \tag{40}$$

a further improvement in performance of the optimum system can be achieved at high signal/noise ratios by using three taps for the filter C, which is now matched to the two values of the 3-component vector

$$Z_3 = [s_i y_0 \quad (s_i y_1 + s_{i+1} y_0) \quad (s_i y_2 + s_{i+1} y_1 + s_{i+2} y_0)] \tag{41}$$

of the minimum Euclidean length. s_i' is here determined at time $t = (i+2)T$ from the sample

$$x_{i+2}' = r_{i+2}' - p_{i+2}' \tag{42}$$

at the detector input, r_{i+2}' and p_{i+2}' being appropriately modified for the new filter C. Again, when the filter C has just two taps but 4-level data symbols are used, such that

$$s_i = \pm 1 \text{ or } \pm 3 \tag{43}$$

it is necessary that

$$|y_1| > 3|y_0| \tag{44}$$

before any advantage in tolerance to noise can be achieved by the optimum arrangement of the modified equalizer over the conventional equalizer. The filter C is now matched to the 2-component vector

$$Z_2 = [y_0 \quad (y_1 \pm 3y_0)] \tag{45}$$

of the minimum Euclidean length.

Since, in the two cases just described, the amplitude distortion introduced by the channel must be very severe indeed before any useful advantage is gained, it is clear that the most important application is that previously studied where binary data-symbols are used and the filter C has two taps.

4 Computer-Simulation Tests

The tolerances to additive Gaussian noise of the optimum and sub-optimum arrangements of the modified equalizer have been compared by computer simulation with that of a conventional equalizer, for two taps in the filter C, binary signals and the seven different channels shown in Table 1. The model of the data-transmission system assumed in the tests is that described in Sections 2 and 3 of this paper and shown in Figs. 1–3.

Table 1

Sampled impulse-responses of the different channels

Channel	Sampled impulse-response Y						
1	0·408	0·816	0·408				
2	0·548	0·789	0·273	−0·044	0·027		
3	0·321	0·620	0·633	0·322	0·087		
4	0·167	0·500	0·667	0·500	0·167		
5	0·29	0·50	0·58	0·50	0·29		
6	0·085	0·289	0·493	0·577	0·493	0·289	0·085
7	0·19	0·35	0·46	0·50	0·46	0·35	0·19

Each vector Y in Table 1 has unit length so that no signal gain or attenuation is introduced by any channel. Channel 1 is a well-known partial-response channel,[15] and Channel 2 was obtained with an actual telephone circuit in the transmission path. Whereas the z transform of the sampled impulse-response of Channel 3 has no zeros (roots) on the unit circle in the z plane, all zeros of the z transforms of the Channels 4–7 lie on the unit circle. Channels 4 and 6 are unfavourable to non-linear equalizers,[13,14] whereas Channels 5 and 7 are particularly unfavourable to maximum-likelihood detectors.[16] The latter channels introduce very severe amplitude distortion. The various channels have been selected to test the different systems over a wide range of values of $|y_1|/|y_0|$, for which equation (10) holds, in order to provide a thorough check for the theoretical results obtained. Furthermore, the tests not only include two practical channels (1 and 2) but also two channels (5 and 7) for which a maximum-likelihood detector is likely to gain only a relatively small advantage in tolerance to additive white Gaussian noise over a conventional non-linear equalizer, at the given high levels of amplitude distortion. A simple and effective method for specifying the level of the amplitude distortion has been described by Clark.[8]

The results of the computer-simulation tests, together with the theoretical estimates given by equations (9), (31) and (35), for the important case where binary data-symbols are transmitted (eqn. (1)), are shown in Fig. 5. The error rate in the detected data symbols $\{s'_i\}$ is here plotted against the signal/noise ratio in decibels. The latter is defined as $10\log_{10}(1/\sigma^2)$, where σ^2 is the variance of the Gaussian noise components $\{w_i\}$ and of the Gaussian noise waveform $v(t)$, the mean-square value of s_i being unity. Each individual measurement of the error rate in the $\{s'_i\}$ has been made with the transmission of either 50 000 or 100 000 data-symbols at the given signal/noise ratio, an average of over 3×10^6 data-symbols being used in plotting each curve (Fig. 5). The 95% confidence limits are up to about $\pm\frac{1}{2}$ dB. However, for the relative performance of the optimum and sub-optimum systems, at all error rates tested, the confidence limits are much better than is suggested by the above figure, through the use of the same sequence of $\{s_i\}$ and the same sequence of $\{w_i\}$ at any given signal/noise ratio.

Before comparing the theoretical and measured results in Fig. 5, it is necessary first to consider the approximations that have been made in the theoretical estimates P_1, P_2 and P_3 of the error rates in the three systems. The approximate theoretical estimate, P_3 (eqn. (9)), of the probability of error in the conventional equalizer assumes the correct values of the last g detected data-symbols $\{s'_i\}$. The incorrect detection of s_i means that its intersymbol interference in the equalized signals $x_{i+1}, x_{i+2}, \ldots, x_{i+g}$ is likely to cause errors in the detected values of some of the following data symbols (an error-extension effect), so that errors tend to occur in bursts. The actual error rate (when this is less than 1 in 10) can now be taken to be P_3 multiplied by the average number of errors in an error burst. The latter number typically lies in the range 1–100 but may occasionally be a little greater. However, at high signal/noise ratios (giving error rates of less than 1 in 10^5) a change of ten times in the error rate, for a given received data signal and a given equalizer, corresponds to a change of less than 1 dB in the signal/noise ratio, so that a discrepancy here in the value of the error rate, of a factor of several times, does not constitute a serious inaccuracy, the error rate being an extremely sensitive measure of the signal/noise ratio.[8]

As the error rate becomes smaller so the value of σ (the standard deviation of w_i) that actually causes this error rate becomes closer to the value estimated theoretically by setting P_3 (eqn. (9)) to the given error rate. The theoretically estimated value of σ, for a given error rate, is an upper bound to the actual value, the bound becoming tighter as the error rate becomes smaller or as the number of the larger components of Y decreases. The latter holds because the greater the level of the intersymbol interference at the output of the filter A, the greater the average length of the error bursts and hence the greater the degradation in performance relative to the theoretical value.

The estimates P_1 and P_2 are approximate lower bounds to the actual error rates in the optimum and sub-optimum systems, respectively. Both P_1 and P_2 neglect a factor of $\frac{1}{2}$ and so, in effect, assume an average of two errors in an error burst, which is in general smaller (and sometimes very much smaller) than that experienced here.

It is evident now that the discrepancies between the theoretical and measured results in Fig. 5 are caused mainly by the error-extension effects (error bursts) which

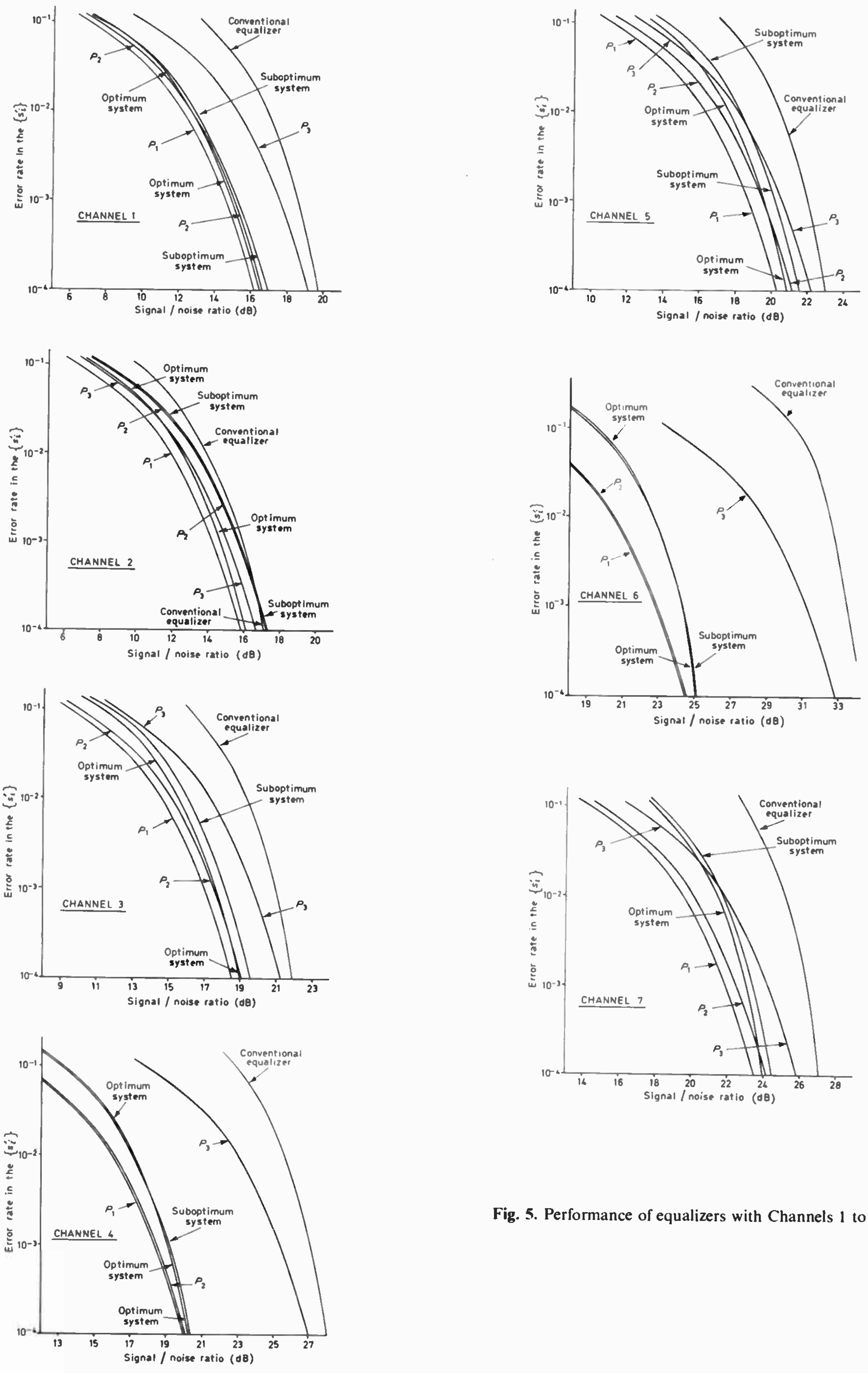

Fig. 5. Performance of equalizers with Channels 1 to 7.

are neglected in P_1, P_2 and P_3 (eqns. (31), (35) and (9)), the influence of these becoming greater as the error rate increases. For any given channel, the discrepancy is greater with the conventional equalizer than with the optimum or sub-optimum system, essentially because of the greater average length of the error bursts in the former case, this being due to the greater relative magnitudes of the intersymbol interference components removed here (with correct detection) by the transversal filter B and subtractor in Fig. 1. The more severe the signal distortion introduced by the channel, the greater the average length of the error bursts with any given system, and hence the greater the discrepancy. At the lower error rates tested the discrepancies are not very serious, and at error rates below 1 in 10^5 they should in most cases be well under 1 dB.

At error rates below 1 in 10^3 the relative tolerances to additive white Gaussian noise of the optimum system, sub-optimum system and conventional equalizer, as given theoretically by P_1, P_2 and P_3, respectively, are quite close to the relative tolerances to noise as measured by the computer-simulation tests. Thus Fig. 4 gives a reasonably accurate comparison of the three systems, at these lower error rates.

Figure 5 shows that, at error rates in the range 1 in 10^3 to 1 in 10^4, the optimum system has in every case a better tolerance to noise than both the sub-optimum system and the conventional equalizer, and it achieves its best performance relative to the conventional equalizer when operating over Channel 6, where it gains a measured advantage of about 9 dB in tolerance to additive white Gaussian noise. With Channel 2 the optimum system gains its smallest measured advantage of only about 1 dB over the conventional equalizer and its greatest advantage of about 1 dB over the sub-optimum system, which, at error rates below 1 in 10^4, has an inferior performance to that of the conventional equalizer. In the case of Channels 4 and 6 the performance of the sub-optimum system is very close indeed to that of the optimum system, and in fact becomes very slightly better at the higher error rates.

Since all zeros (roots) of the z transform of the sampled impulse-response of each of the Channels 1 and 4–7 lie on the unit circle in the z plane, which means that these channels are minimum phase, the linear feedforward transversal filter A in Fig. 1 is omitted here and the conventional equalizer now becomes a 'pure non-linear equalizer'.[8] It can be seen from Figs. 4 and 5 that in the case of Channel 1, which is one of the well-known partial response channels formed by correlative-level coding,[15] an improvement of 3 dB in tolerance to noise over that of a conventional system can be achieved through the inclusion of the appropriate 2-tap filter C (eqn. (38) and Table 1) at the input to the subtractor in Fig. 1.

It is evident from equation (44) and Table 1 that, when 4-level data symbols are transmitted (eqn. (43)), no useful advantage in tolerance to additive white Gaussian noise is likely to be achieved by the optimum system over the conventional equalizer, with any of the Channels 1–7 at low error rates, the performance of the sub-optimum system being now in every case almost certainly inferior to that of the conventional equalizer.

The essential mechanism used by both the optimum and sub-optimum systems is to delay the detection of s_i until the receipt of r_{i+1} and then to operate on the two received samples r_i and r_{i+1}, to give the detected value of s_i, thus involving both $s_i y_0$ and $s_i y_1$ in the detection of s_i. The conventional equalizer uses only $s_i y_0$. Other techniques have been described whereby the tolerance to noise of an equalizer can be improved by delaying the detection of s_i,[8] but these all involve a significant increase in the equipment complexity. The importance of the techniques described here is that, with a known time-invariant channel and a given filter A in the receiver, the only change required in the equipment is the addition of one tap to the filter A and, of course, the appropriate adjustment of all tap gains.

The great improvement in performance achieved by the optimum and sub-optimum systems, when $|y_1| \gg |y_0|$, suggest that under certain conditions the linear feedforward transversal filter A could be omitted altogether leaving only the 2-tap filter C and thereby greatly reducing the complexity of the equalizer. Extensive computer simulation tests have in fact been carried out on such systems over a range of different channels and some surprisingly encouraging results have been obtained. For example, in the presence of the more severe amplitude distortion and in the absence of phase distortion, an improvement in performance may be obtained by replacing the filter A by the appropriate 2-tap filter C. It is hoped that the results of these tests may be published shortly.

When the filter A is omitted and the channel introduces significant phase distortion (which, of course, is not now removed), the corresponding arrangements of the optimum and sub-optimum systems do not normally operate well. This is because there now tends to be a steady but slow increase in the magnitudes of the components of the sampled impulse-response of the channel, with the time delay of these components, the largest component no longer being among the first two or three. However, for channels formed by coaxial cables or 600 Ω pairs, the impulse response usually rises quite rapidly to its peak and then decays at a progressively decreasing rate, this shape being ideally suited to applications of the optimum and sub-optimum systems without the filter A. When transmitting data over such a channel at a relatively high rate, the time interval between the start of the impulse response and its peak is appreciably greater than the sampling interval, which means that the second non-zero component of the sampled impulse-response of the channel is normally greater than the first and often very much greater (depending upon the phase of the sampling instants). Now, even with the optimum sampling phase, the performance of a conventional pure non-linear equalizer (decision-feedback equalizer with no linear transversal filter ahead of the detector) will fall well below the optimum obtainable with a more sophisticated detection process. The addition of the 2-tap filter C (with the appropriate tap gains) ahead of the detector should now give a very useful improvement in performance with no significant increase in complexity. There is clearly considerable potential in such applications for the techniques described in this paper. Further details of these systems are however beyond the scope of the paper.

5 Conclusions

When a non-linear (decision-feedback) equalizer is used with a binary signal that has been subjected to severe amplitude distortion, a useful improvement in tolerance to additive white Gaussian noise can be achieved by appropriately modifying the tap gains of the equalizer without changing its basic structure. The equalizer does not now attempt to achieve the accurate equalization of the channel or even to minimize the mean-square error in the equalized signal, but instead deliberately introduces an appreciable level of intersymbol interference into the equalized signal in return for which the noise level here is considerably reduced, giving an overall improvement in tolerance to noise. Unfortunately, the arrangement is not suitable for use with multi-level signals. An important application of the technique appears to be at very high transmission rates, in excess of 1 Mbit/s, where it is required to achieve the highest available transmission rate, with binary signals and using only simple equipment.

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Contributors to this issue

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A. Clark

Ken Freeman (Member 1961, Graduate 1958) obtained his B.Sc. with honours in mathematical physics from Birmingham University in 1953. After National Service he joined Philips Research Laboratories where he has carried out research into various aspects of television, particularly colour displays. He is currently engaged in studies of high-quality colour systems and displays. Mr Freeman has served on the Papers Committee since 1973, and has contributed a number of papers to the Journal; in 1959 he received the Associated Rediffusion Premium (now called the Paul Adorian Premium) for a paper on a gating circuit for single-gun colour television tubes.

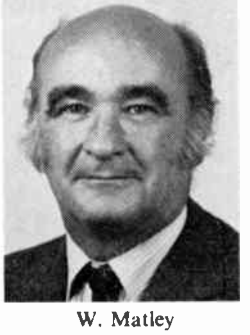
Robert Bywater graduated from Battersea College of Technology, London, with a B.Sc. degree in electrical engineering in 1964. After spending a further year reading computing engineering, he joined the Computer Systems Development Group of ICL, then English Electric Computers. Currently, he is a Reader in the Department of Electronic and Electrical Engineering at the University of Surrey with research interests in special-purpose computers. He gained his doctorate in 1975 for a thesis arising from this work on various aspects of which he has contributed papers to this Journal.

Wallace Matley is a graduate of London University and the first eleven years of his professional career were spent in the Royal Naval Scientific Service on all aspects of radar research and development from systems research to post-design development. From 1954 to 1956 he was with the Australian Department of Supply designing and developing rocket range instrumentation; he returned to the RNSS in 1956 and for the next five years worked on microwave-aerial research, missile-system command and display development. From 1961 to 1964 he was Chief Engineer at the Royal Observatory, Edinburgh, implementing the application of electronics to astronomy, and in 1964 he went to the Government Communications Headquarters to do communication system research. Since 1969 he has been a Senior Lecturer and Head of the Industrial Electronics Research Unit in the Department of Electronic and Electrical Engineering, University of Surrey.

Adrian Clark (Member 1964) who graduated from the University of Cambridge in 1954, is now a Reader in the Department of Electronic and Electrical Engineering at Loughborough University of Technology. Before joining the University in 1970, Dr Clark spent 12 years in industry at Plessey Telecommunications Research, Taplow, working on various aspects of digital communication systems: this period was interrupted for three years when Dr Clark held an Industrial Fellowship at Imperial College, London, as a result of which he was awarded his doctorate. He has numerous papers in the Institution's Journal and conferences to his credit and has published two books on Data Transmission. Dr Clark is a member of the Communications Group Committee and has also served on conference organizing committees.

Paul Howser (Member 1971) was awarded the Diploma in Electrical Engineering at the Borough Polytechnic in 1964 and graduated from the Open University in 1973 with a Bachelor of Arts degree. His engineering training started in 1955 with a technical apprenticeship at J & E Hall, Dartford, Kent, followed by graduate training at Associated Electrical Industries, Rugby. After a time with Canadian Vickers, Montreal, he rejoined J & E Hall to introduce solid-state techniques into lift control system design. In 1966 he entered the radar and flight simulation fields at Redifon, Crawley, where as Principal Engineer he was engaged in project management on large digital simulation systems. Six years later, after a period in quality assurance with Stone-Platt (Crawley), he joined the University of Surrey. In addition to his current research interest in digital signal processing, his present responsibilities include the general management of the Industrial Electronics Research Laboratories.

R. Bywater



W. Matley

P. Howser

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.

MARCH

MT '82 *15th to 18th March*
Materials and Testing Exhibition, to be held at the National Exhibition Centre, Birmingham. Information: John Payne, Hampton Mill, Evesham, Worcestershire

Computers *16th to 18th March*
The Scottish Computer Show, will be held in the Albany Hotel, Glasgow. Information: Beverley Dellow, Couchmead Ltd, 42 Great Windmill Street, London W1V 7PA. (Tel. 01-437 4187)

Electro-Optics *23rd to 25th March*
International conference on Electro-optics and Lasers organized by Kiver Communications will be held in Brighton. Information: Kiver Communications, Millbank House, 171–185 Ewell Road, Surbiton, Surrey.

***Instrumentation in Flammable Atmospheres** *25th March*
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).

***Microprocessor Control** *28th March to 2nd April*
Vacation School on Industrial Digital and Microprocessor Based Control Systems, organized by the Institution of Electrical Engineers, to be held at Balliol College, Oxford, Information: The Secretary, IEE, Savoy Place, London WC2R 0BL. (Tel. 01-240 1871).

Electronic Components *29th to 31st March*
International conference on new applications of Passive Electronic Components, organized by the Société des Electriciens, Electroniciens et des Radioélectriciens in association with the Electronics Industries Group will be held in Paris. Information: SEER, 48 Rue de la Procession, 75015 Paris.

Reliability *29th March to 2nd April*
One week international CBO-seminar on Reliability Engineering—Advanced Technology and Industrial Applications will be held in Rotterdam. Information: CBO Management and Technology Systems Centre, P.O. Box 30042, 3001 DA Rotterdam, Netherlands. (Tel. (010) 13 90 20)

***Microprocessor Control** *29th March to 2nd April*
Vacation School on Advanced techniques for microprocessor systems, organized by the Institution of Electrical Engineers, to be held at UMIST, Manchester. Information: The Secretary, IEE, Savoy Place, London WC2R 0BL. (Tel. 01-240 1871).

***CAD** *30th March to 1st April*
Computer Aided Design Conference and Exhibition, organized by IPC Science & Technology Press, to be held in Brighton. Information: IPC Science & Technology Press, P.O. Box 63, Westbury House, Bury Street, Guildford GU2 5BH. (Tel. (0483) 31261).

***Peripherals** *31st March to 2nd April*
Peripherals Suppliers Exhibition, organized by IPC Exhibitions Ltd, to be held at the West Centre Hotel, London. Information: IPC Exhibitions Ltd, Surrey House, 1 Throwley Way, Sutton, Surrey SM1 4QQ. (Tel. 01-643 8040).

APRIL

***Electronic Components** *1st to 7th April*
International Exhibition of Electronic Components, organized by French Trade Exhibitions, to be held in Paris. Information: French Trade Exhibitions, 54 Conduit Street, London W1R 9SD. (Tel. 01-439 3964).

Control Systems in Medicine *5th to 7th April*
Meeting on Control Systems, Concepts and Approaches in Clinical Medicine, organized by the Institute of Measurement and Control, to be held at the University of Sussex, Brighton. Information: Mr M. J. Yates, Deputy Secretary, Institute of Measurement and Control, 20 Peel Street, London W8 7PD (Tel. 01-727 0083/5)

On-Line Control *5th to 8th April*
International Conference on Trends in On-line Computer Control Systems, organized by the Institution of Electrical Engineers, to be held at the University of Warwick, Coventry. Information: Conference Department, IEE, Savoy Place, London WC2 0BL (Tel. 01-240 1871, ext. 222).

IFFSEC '82 *19th to 23rd April*
International Fire, Security & Safety Exhibition and Conference, to be held at Olympia, London. Information: Victor Green Publications Ltd, 106 Hampstead Road, London NW1 2LS. (Tel. 01-388 7661).

Electronics *20th to 22nd April*
The All-Electronics/ECIF Show, sponsored by the Electronic Components Industry Federation, to be held at the Barbican Centre, City of London. Information: 34–36 High Street, Saffron Walden, Essex CB10 1EP (Tel. 0799 22612 Telex: 81653).

★ **Recording** *20th to 23rd April*
Fourth International Conference on Video and Data Recording, organized by the IERE with the association of AES, IEE, IEEE, IoP, RTS and SMPTE, to be held at the University of Southampton Information: Conference Secretariat, IERE, 99 Gower Street, London WC1E 6AZ (Tel. 01-388 3071) *EE, 18th June, p. 2.*

● **Communications '82** *20th to 23rd April*
Conference organized by the IEE in association with the IEEE and the IERE, to be held at the National Exhibition Centre, Birmingham. Information: IEE Conference Department, Savoy Place, London WC2R 0BL (Tel. 01-240 1871).

***Instrumentation in Flammable Atmospheres** *22nd April*
(See item for 25th March)

Reliability *26th to 30th April*
Seminar on Techniques in Reliability Engineering, organized by Plessey Assessment Services, to be held at the Red House Hotel, Barton-on-Sea, Hampshire. Information: Richard Morgan, Plessey Assessment Services, Titchfield, Fareham, Hants, PO14 4QA (Tel. (03924) 43031).

● **Acoustics** *29th to 30th April*
International conference on Spectral Analysis and its use in Underwater Acoustics, organized by the Underwater Acoustics Group of the Institute of Acoustics in association with the IEE, IERE, IMC, IMA, ASA and the IEEE to be held at Imperial College, London. Information: Dr T. S. Durrani, Department of Electronic Science & Telecommunications, University of Strathclyde, Royal College, 204 George Street, Glasgow G1 1WX.

MAY

Acoustics, Speech & Signal Processing *3rd to 5th May*
International Conference on Acoustics, Speech & Signal Processing, sponsored by the IEEE, to be held in Paris. Information: Prof. Claude Gueguen, Département Systemes et Communications, Ecole Nationale Superieure des Telecommunications, 46 Rue Barrault, 75634 Paris, Cedex 13 France.

***Computers** *4th to 6th May*
Compec Europe Exhibition, to be held in Brussels. Information: IPC Exhibitions Ltd, Surrey House, 1 Throwley Way, Sutton, Surrey, SM1 4QQ. (Tel. 01-643 8040).

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

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

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

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

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

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

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

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

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

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

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

JUNE

***Digital Audio** *4th to 6th June*
Conference on The New World of Digital Audio, organized by The Audio Engineering Society, to be held at the Rye Town Hilton, Rye, New York. Information: Audio Engineer-

ing Society, 60 East 42nd Street, New York, NY 10165, USA. (Tel. (212) 661-2355/8528).

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

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

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

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

Fisheries Acoustics *21st to 24th June*
Symposium on Fisheries Acoustics organized by the International Council for the Exploration of the Sea with the collaboration of the United Nations Food and Agriculture Organization, to be held in Bergen, Norway. Information: General Secretary, ICES, 2-4 Palaegade, 1261 Copenhagen K, Denmark
(Papers by 31st March 1982)

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

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

JULY

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

Simulation *19th to 21st July*
1982 Summer Computer Simulation conference will be held at the Marriott-City Centre, Denver, Colorado. Information: Lawrence Sashkin, 1982 SCSC Program Director, The Aerospace Corporation, P.O. Box 92957, Los Angeles, California 90009. (Tel. (213) 648-5934)

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

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

AUGUST

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

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

SEPTEMBER

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

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

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

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

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

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

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

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

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

***Instrumentation in Flammable Atmospheres** *30th September*
(See item for 25th March)

OCTOBER

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

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

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

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

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

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

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

***Instrumentation in Flammable Atmospheres** *28th October*
(See item for 25th March)

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

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

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

NOVEMBER

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

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

1983

FEBRUARY

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