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Merger of Appleton and Rutherford Laboratories

IN November 1976 the Science Research Council set up a working party to inquire into the future of the Appleton Laboratory. For some time there had been concern about the problems of providing proper support for the national space science programme required for the 1980s from the relatively limited resources available at Appleton. In the same period the SRC Engineering Board was seeking to widen the scope of the radio propagation work and, in particular, to introduce work on communications systems.

The Council decided that the Appleton and Rutherford Laboratories should be brought together under common management, that the Ditton Park (Appleton Laboratory) site should be closed and as much work as possible should be transferred from there to the Chilton (Rutherford Laboratory) site.

Meeting the Demands of 'High Technology'

Explaining the need for the merger, the chairman of the SRC, Professor Sir Geoffrey Allen, F.R.S., wrote in the *SRC Bulletin* in June:

'We are at a turning point in our space research programmes and the way in which we conduct our experiments in space in the future is now under examination. It is already clear, however, that there will be a growing demand for experiments to be conducted from space and that SRC must have the capability to manage the development of complex projects, both large and small scale, if the UK is to remain a major force in space science. This requires the ability to marshal a wide range of skills ranging from project management through to the design and development of instrument packages, if we are to be able to exploit the new launching systems soon to be available outside the UK.

'Although the Appleton Laboratory has been responsible for most aspects of space research and project management, some work has also been undertaken by the Rutherford Laboratory; both laboratories have also been engaged in high technology areas of great relevance to space. By combining the experience and expertise of the two laboratories we believe we will be able to create a team that can call on the resources required to provide a strong focus for the support of space research in the UK. In the especially demanding field of space science such a team can only be kept in the forefront of developments if some of its members are themselves engaged in space research. Experience shows that space projects can be best managed by a team that has the ability and experience to undertake any part of the activity itself, even if much of the work is contracted out to industry and most of the scientific users are in universities. What is required will vary with time but, by having a space team based on the combined resources of the two laboratories on a site containing a wide variety of projects and skills, we believe we will have the flexibility to meet the varying needs of the future, as well as providing good career prospects for the staff. As the requirements for experiments from space grows, this group will be able to give strong support to the universities, as both laboratories have done in the past. It will also serve other scientific institutions including the support of work sponsored by other bodies besides SRC.'

The Communications Systems of the Future

Sir Geoffrey believed that bringing the two laboratories together would strengthen and extend the existing work in the communication field. He continued: 'We recognize the national importance of the role that the Appleton Laboratory plays as a major centre for radio propagation expertise and we look to see a strengthening of the links between this team and work elsewhere, in particular a build up of propagation work in the universities. A need extending beyond the traditional radio work of the Appleton Laboratory has also been identified. The full exploitation of the radio spectrum will require the sophistication of electronic circuitry, which microelectronics offers, in developing new modulation techniques designed, for example, to combat multipath effects in the ionosphere and to take advantage of adaptive antennas. Many of these new developments are most likely to arise in data communications. The Rutherford Laboratory has considerable experience of data communications gained from the development of the Interactive Computing Facility, and thorough research in the contiguous field of Distributed Computing Systems.

'Moreover, the Rutherford Laboratory provides the focus for the Council's development programme in microelectronics. This programme, developed in close association with the Department of Industry and with industry, rests on four university-based device centres with a major central facility at the Rutherford Laboratory for the production of masks by electron beam lithography. Device designers must be brought together with system designers to realize the full potential of large scale integration.

'So we feel that our communications systems programme will flourish by bringing together the radio expertise at Appleton with the digital communications expertise of Rutherford on a site where advanced microelectronics research is in progress. The main thrust in this programme must come from the university community but the involvement of the merged Laboratory should powerfully amplify that development and link it properly with broader national activities in the relevant fields.'

Timetable for the Move

As a consequence of the Joint European Torus project going to Culham the Appleton's Astrophysics Research Division will move from there to the Chilton site in October 1980. The UK team on the international Infra Red Astronomical Satellite will move to the Chilton site next year and some part of this work is already in progress there. It is expected that the majority of the work on the Ditton Park site will be moved to Chilton in about two years' time when accommodation is ready for them.

The New Management

In September of this year Dr Godfrey Stafford, currently the Director of the Rutherford Laboratory, took over as Director-General of the combined Laboratories on the retirement of Dr Fred Horner, the Director of the Appleton Laboratory for the past two years.

Professor John Houghton, F.R.S., Professor of Atmospheric Physics, Oxford University, has been appointed Director (Appleton) in the combined Laboratories on 1st September in succession to Dr Horner. Professor Houghton, who is on a five-year secondment, will maintain links with the work of his University Department.

He is well known internationally for his outstanding research in the upper atmosphere and his experiments have been carried on a number of space missions including NASA's *Nimbus* series and Venus Orbiter. A member of the Astronomy, Space and Radio Board of the Council 1970 to 1973 and since 1976, many of his experiments have received engineering and project support from the Rutherford Laboratory. Professor Houghton was President of the Royal Meteorological Society for 1976–78.

An Historical Epilogue

Originally a division of the National Physical Laboratory, radio research has been carried out at Ditton Park since soon after the First World War: indeed early experiments into the feasibility of 'radio location' were conducted there in the 'thirties, notably by Watson-Watt and Wilkins. The widening extent of its work was recognized in the change of the name from the Radio Research Station to the Radio and Space Research Station: the present name was adopted in 1973 and commemorates the famous pioneer in ionospheric investigation and former Secretary of the Department of Scientific and Industrial Research, predecessor of the SRC.

The proposed move therefore very much marks the end of an era in British radio research. While Sir Geoffrey's explanation of the thinking behind the merger can be appreciated, many radio engineers will learn of the eventual closing of Ditton Park with some regret.

F.W.S.

ANNOUNCEMENTS

Conference on 'Radio Transmitters and Modulation Techniques'

The Electronics Division of the Institution of Electrical Engineers, in association with the IERE, the Radio Society of Great Britain and with the support of the Convention of National Societies of Electrical Engineers of Western Europe (EUREL), is organizing a Conference on 'Radio Transmitters and Modulation Techniques' to be held at the IEE, Savoy Place, from 24th–25th March 1980.

It is expected that the Conference will bring to light many new techniques suitable for use in modern radio transmitters, and it is proposed to cover the following topics:

Communication transmitters (fixed and mobile)
Broadcasting transmitters
Television transmitters
Navigational aid transmitters
Improvements in transmitting valves
Impact of power semiconductors on transmitter designs
New methods of modulation
Exploitation of Doherty and pulse width modulation and other methods for the purpose of higher efficiency
Transmitter control/tuning, protection, safety
Common antenna working (filters and other means)
Linearity control
Frequency and signal generation
Automatic monitoring and correction
Spurious frequencies and noise radiation

Registration forms and programme details are now available from the IEE Conference Department, Savoy Place, London WC2R 0BL.

Graham Clark Lectures

The Council of Engineering Institutions holds stocks of published Graham Clark Lectures given by a number of eminent engineers and academics over the years. The following are available upon request at 20 p per copy, postage paid, from the CEI offices, 2 Little Smith Street, London SW1P 3DL.

No. 13. *Problems of the Organisation of Science in the Modern World*, by Lord Bowden

No. 14. *The Problems of Engineering and Scientific Manpower and their Implications for National Policy*, by Lord Jackson of Burnley

No. 15. *The Sea of Opportunity*, by Sir Frederick Brundrett

No. 17. *World Sources of Energy in the Late Twentieth Century*, by Lord Hinton of Bankside

No. 18. *Conservation and Use of Water Resources*, by Sir Norman Rowntree

No. 19. *Natural Resources, the Engineer and the Environment*, by Sir Kingsley Dunham

No. 20. *The Diseconomies of Size and the High Cost of Discontinuity*, by Sir Frederick Catherwood

No. 21. *The Voice of the Engineer in Public Policy*, by Sir Alan Cottrell

No. 22. *Engineers and Politics: A Case History*, by Lord Ashby of Brandon.

No. 23. *The Challenge of Microelectronics*, by Dr R. J. Clayton

Government Steering Committee set up to Plan Conference on Education and Training of Engineers

The Secretary of State for Education and Science, the Rt Hon. Mark Carlisle, has appointed a steering committee to organize a national conference on education and training of engineers.

The committee of ten, under the chairmanship of Mr J. R. S. Morris, Deputy Chairman, National Enterprise Board (formerly Group Technical Director, Courtaulds), includes a number of leading engineers drawn from major sectors of the profession, and has the support of the Council of Engineering Institutions and the Confederation of British Industry. Among the members are Sir John Atwell, C.B.E., F.Eng., Chairman of CEI and former Chairman of The Weir Group; Mr K. Corfield, Chairman and Managing Director, Standard Telephones and Cables; and Professor J. C. West, D.Sc., C.Eng., Vice-Chancellor of the University of Bradford.

Its task is to prepare the ground for the conference, which will examine comment and opinion stemming from the publication later this year of the Finniston Committee of Inquiry's report on the engineering profession.

The conference will also consider recent or proposed initiatives, attempt to present a comprehensive picture and identify areas where discussions or further studies are needed. It is expected to take place in 1980.

Institutional murder

'Ten ways to kill your Federation or other similar organization' have come to our notice via CEI Digest via the FEANI Quarterly Newsletter who took the list from 'an English review'. They are as follows:

* Do not attend meetings.
* If by chance you do attend a meeting, arrive late.
* Criticize the officers and committee members.
* Be annoyed if not offered office.
* Never accept office as it is easier to be critical than to participate actively.
* Should you be asked to give an opinion, say that you have none. Subsequently complain that nobody seems to want to hear your views.
* Only do anything that is to your own advantage but, when other members get up off their haunches and give their all to the general good of the organization, grumble about it being run by a vain clique.
* Delay payment of your subscription for as long as possible.
* Never introduce new members.
* Complain that insufficient publicity is given to your organization's activities but never offer to undertake a lecture or agree to being a contributor to your own journal.

Accepting that free translation from English into French and back into English can sometimes have laughable results, somehow we think that this list is alas! only too near the truth for complacency.

Members' Appointments

R. M. Denny (*right*)
D. P. Taylor (*left*)

CORPORATE MEMBERS

R. M. Denny (Fellow 1974, Member 1955) has been appointed Managing Director of Rediffusion. He first joined the Board in 1974 and is also Chairman of several companies in the Rediffusion Group.

T. J. Oliver (Fellow 1978, Member 1970) who has been Telecommunications Engineer with British Rail, Southern Region since 1974, has now joined Logica as Principal Consultant to its Communications Group.

D. P. Taylor (Fellow 1972) has been appointed Chairman and Managing Director of Sinclair Electronics, following the reorganization of the company into three parts at the request of the National Enterprise Board. Mr Taylor, who left Hewlett Packard in 1978 where he had been Managing Director since 1969, and became a non-executive director of Sinclair Radionics last December, now has charge of the part carrying on the scientific instrument business. (The Microvision/consummer electronics part has been acquired by Binatone, while the founder of the company, Clive Sinclair, is setting up a separate research company as a joint venture with an industrial partner to exploit a new flat screen tube.)

M. J. Belben (Member 1973) who recently retired from the Royal Air Force with the rank of Squadron Leader after completing 24 years' service, has been appointed Sales Manager, Defence Systems, with Plessey Radar.

Group Captain D. G. J. Breadner, B.Sc., RAF (Member 1968) has taken up an appointment as Staff Officer Engineering, Royal Air Force, HQ No 11 Group at Stanmore, Middlesex. Previously he was at RAF Sealand as Officer Commanding No. 30 Maintenance Unit.

Major M. D. Cherry, R Signals (Member 1973, Graduate 1970) is now Group Supervisor, Radio Group, Royal Signals Trade Training School, Catterick. He was previously Technical Adjutant, 22nd Signal Regiment.

Lt-Col L. E. Cushion, REME (Ret.) (Member 1971) has been appointed Management Services Executive with Worthington Simpson, Newark, Notts. Before his retirement from the Army on 31st May last, Col Cushion was Officer Commanding 33 Central Workshop REME.

C. W. Dennay (Member 1973, Graduate 1963) has been appointed Chief Engineer, External Broadcasting with the BBC. A lecturer at the BBC Training Establishment at Evesham for some 13 years, he went to the Transmitter Headquarters in 1973 and became Assistant Chief Engineer in 1978.

L. J. Ferrington (Member 1973, Graduate 1971) who has been Export Sales Manager with Negretti and Zambra has now been appointed General Sales Manager.

R. N. Jones (Member 1970, Graduate 1963) has taken up a new post as Chief Engineer with Scitec Corporation, Sydney, N.S.W. Previously he was Technical Director of Microcom Australia, Southport, Queensland.

Brigadier G. R. Oehlers (Member 1971), Commander Corps Royal Signals, 1st British Corps in Germany since 1977, has been posted back to the UK to take up an appointment as Director of Army Telecommunications and Director Operational Requirements (Army) for command, control, communication and computer equipments at the Ministry of Defence.

I. E. Powers (Member 1973, Graduate 1968) has been appointed Marketing Manager, MPU Products with Motorola. Prior to his promotion, Mr Powers was Motorola's Senior MPU Applications Engineer. He was previously with Plessey and was mainly concerned with work on telecommunications, and the design of software and hardware systems in military and data processing systems.

NON-CORPORATE MEMBERS

K. Ariyaratnam, B.Tech. (Graduate 1976) has been appointed Telecommunications Engineer for the Northern Region in the Department of Posts and Telecommunications, Sri Lanka, covering the Trincomalee, Jaffna and Anuradhapura areas where he is in charge of network planning, development and installation.

W. D. Mercer, B.A. (Associate Member 1973, Associate 1970) has taken up an appointment as Chief of Television and Film at the United Nations in New York. Mr Mercer's previous posts were with BBC Scotland as a studio engineer and subsequently with Dubai Radio and Colour Television.

Obituary

James Ronald Alexander, B.Sc. (Member 1956, Graduate 1948) died on 3rd May 1979, at the age of 57 leaving a widow. Following war service and a period in industry, Mr Alexander entered the teaching profession. He was for some six years in the Physics Department at Southend High School for Boys and he subsequently joined the staff of Southend Municipal College as lecturer on electronics.

Gerard William Edward Vincent (Member 1970, Graduate 1969) died on 23rd May 1979, aged 47, leaving a widow and three young daughters. Before emigrating to Canada in 1970, Gerard Vincent had worked for some twelve years in the Post Office, interrupted by two years' National Service in the Royal Air Force. He then joined Unilever Instrumentation and Control Department as a Technical Assistant where he remained for five years before leaving in 1969 to become Sales and Service Manager with the Electronics Corporation of America (UK) and Southern Techniques (Electronics). In Canada he held an appointment initially with Hermes Electronics, Dartmouth, Nova Scotia, and in 1978 he joined Varian Associates of Canada as Regional Manager (Ottawa and Eastern Region) for the Electron Device Group. Since 1971 he had been registered as a Professional Engineer of Ontario.

* * *

A premature report of the death of Mr **F. F. Kemp** (Member) was included among the obituary notices in the September Journal. This had been initiated by the return to the Institution of a journal on which his name had been confused with that of his father. We are pleased to publish this correction and to apologize to Mr Kemp and his family and friends for any embarrassment the mistake may have caused.

Applicants for Election and Transfer

THE MEMBERSHIP COMMITTEE at a meeting on 2nd October 1979 recommended to the Council the election and transfer of the following candidates. In accordance with Bye-law 23, the Council has directed that the names of the following candidates shall be published under the grade of membership to which election or transfer is proposed by the Council. Any communication from Corporate Members concerning the proposed elections must be addressed by letter to the Secretary within twenty-eight days after publication of these details.

October Meeting (Membership Approval List No. 264)

GREAT BRITAIN AND IRELAND

CORPORATE MEMBERS

Transfer from Member to Fellow

ROBERTS, David Edward. *Solihull, Birmingham*

Transfer from Graduate to Member

CUBLEY, Leonard David. *Newcastle on Tyne.*
DUNKLEY, John. *Chesham, Bucks.*
WIGNALL, John Michael. *Gloucester.*

Direct Election to Member

BARRITT, Terence Edward. *Fareham, Hants.*
JAMES, Russell. *Huntingdon, Cambridge.*
MUSTAFA, Mohmed Adam. *Portsmouth, Hants.*

NON-CORPORATE MEMBERS

Transfer from Student to Graduate

HUQ, Mohammed Mahmudul. *London.*

Direct Election to Graduate

OLUSANYA, Temitope. *London.*

Direct Election to Associate Member

BAILEY, Anthony Walter. *Bromley, Kent.*
CLEMENTS, Brian Roger. *Northampton.*
JOSEPH, Gbolahan Olugbemi. *London.*
NNAMUCHI, Louis. *Sheffield.*
OKOBIAH, John Okiemute. *Sheffield.*

Direct Election to Student

GOULDING, Declan Peter. *Dublin.*
WHELAN, Ronan Anthony. *Dunladire Co. Dublin.*

OVERSEAS

CORPORATE MEMBERS

Direct Election to Fellow

ALLOS, Janan Emmanuel. *Baghdad, Iraq.*

Transfer from Graduate to Member

WONG, Hon Shu. *Kowloon, Hong Kong.*

Direct Election to Member

AWODIPE, Abayomi. *Schenectady, New York, U.S.A.*
GOH, Jui Chang. *Singapore.*
PINTO, Hewagama P.A.L. *Trincomalee, Sri Lanka.*

NON-CORPORATE MEMBERS

Direct Election to Graduate

LEE, Joseph Yiu Fai. *North Point, Hong Kong.*

Transfer from Student to Associate Member

AMARASENA, Mahanthe A. *Meegoda, Sri Lanka.*
GAMVARA, Anuradha G.T. *Talangama, Sri Lanka.*
VIJAYAGOPAL, Vathavoor. *Colombo, Sri Lanka.*

Direct Election to Associate Member

KHO, Chiang Boo. *Kuching, Sarawak, East Malaysia.*
LO, Hoo Ting. *Smithtown, New York, U.S.A.*
SAPPOR, Joseph M.T. *Accra, Ghana.*
SAPIENZA, Frank. *St. Julians, Malta.*
SHEIKH, Arjmend Akhtar. *Islamabad, Pakistan.*
WANG, Chuen Wing M. *Wanchai, Hong Kong.*

Direct Election to Associate

YIP, Yee Woo. *Tehran, Iran.*

Direct Election to Student

FOO, Teck Song. *Singapore.*
HAR, Chung Yin. *Kowloon, Hong Kong.*
HUI, Wai Wing. *Yuen Long, N.T. Hong Kong.*
KWOK, Ngai Ming. *Wanchai, Hong Kong.*
LAU, Siu Ki. *Kwun Tong, Hong Kong.*
NG, Mou Kwan. *Hong Kong.*
POO, Hoi Wah P. *Hong Kong.*

Ship Handling Trainer with Computer-generated Visual Simulation

An order, valued at nearly £1 million, has been received by Marconi Radar Systems Limited, a GEC– Marconi Electronics company, from the Department of Industry, as prime contractor for the supply, installation and commissioning of a ship handling trainer to be used in a United Kingdom maritime college. Using the experience and technology gained in the development of a somewhat similar type of maritime simulator, Marconi Radar will employ the company's well-proven computer generated imagery system, TEPIGEN, allied to a Decca ship simulator bridge, to provide an advanced all-weather, day and night training system.

The primary purpose of the new simulator is to provide better training in the interest of safety at sea. The requirements for training have been stimulated by a number of recent incidents at sea, which opened public awareness to the possibility of major ecological disaster. A Government paper was issued in January 1979 outlining measures to prevent collisions and strandings of noxious cargo carriers in waters around the United Kingdom and amongst other recommendations stated that 'Simulators to provide training and refresher courses should be made available to the United Kingdom and that other countries should be encouraged to do so.' It further stated that 'overseas aid should be made available' for the latter purposes and that 'the possibility of using EEC funds should also be explored'.

The TEPIGEN system to be used as the 'all-weather' visual part of the trainer uses life-like projected colour television pictures to simulate the actual visual conditions obtaining from the bridge throughout the course of an exercise. The television pictures are synthesized wholly from a computer without the use of television cameras, models, video tape or film. Scenery produced in this way can be manipulated freely and quickly to give a view which changes as the ship alters its heading and position. The size, location, orientation, visibility and colour of objects in the scene are controlled by the computer in response to the details set up for that particular exercise, and are fully interactive with the ship's manoeuvres in response to helm and engine demands. Moving objects such as other ships, from pilot vessels to VLCC's, can be included in the 'playing' area and can be under the control of an instructor; the scene details can include buoyed channels, approaches to real ports, coastlines or open sea.

Transition from daylight through dusk to darkness is realistic and navigation and flashing lights can be shown including general coastal and port lighting. Visibility can be varied from 'unlimited' to thick fog.

The bridge part of the simulator is a developed version of the Decca Ship Simulator which has been successfully in use at the College of Nautical Studies at Warsash for nearly two years. The bridge contains all the normal 'on-board' instruments, radar and navigation displays and these will be integrated with ship's performance equations developed by the Department of Industry's National Maritime Institute and with the TEPIGEN system to provide a full mission capability trainer with the principal advantages of realism, extreme flexibility and simplicity of operation.

Colloquium Report

Peripheral Components for Microprocessors

Organized by the IERE Components and Circuits Group and held in London on 12th June 1979

Almost 100 delegates filled the lecture hall of the Royal Institution to hear five speakers from manufacturing industries describe new developments in the field of peripheral components for microprocessors. Such peripheral devices, the organizing committee had thought, had been somewhat neglected at meetings in comparison to the microprocessors themselves. Five lively talks helped redress this balance by emphasizing the applications aspects and required peripherals rather than the latest architectural improvement or performance feature of the microprocessor chip.

The chairman for the day's programme was Ken Baker from Plessey Microsystems, who, in introducing the morning session in which all the speakers were from the integrated circuit manufacturers, commented on how important a role the silicon technology has assumed in peripherals. The first paper given by Paul Mayes, the managing director of Mostek UK, discussed the most common microprocessor peripherals of all, namely 'memories'. He reviewed the advancements in circuit design and manufacturing techniques that had enabled a sixteen-fold increase in packing density and a five-fold improvement in performance to be achieved in just six years. The technology described was what Mostek call the 'scaled Poly 5' an advanced form of n-m.o.s. This has enabled dense dynamic r.a.m.s, fast static r.a.m.s, r.o.m.s and p.r.o.m.s all operating from a single 5 V rail to be produced. He described a novel technique to reduce power consumption by switching to standby mode after address signals have settled. The new 4816 d.r.a.m. was described which has built-in refresh multiplexers, refresh address counter and chip enable which has been primarily developed to interface the 280 microprocessor.

The second paper on peripheral controllers was by Phil Pittman of Zilog. He presented the i.c. manufacturers philosophy of increasing circuit density to provide enhanced system performance by distributing more intelligence into the peripheral controllers. These controllers, he said, need to be flexible enough to be used in a wide range of specific application that were described. This concept is achieved by allowing the microprocessor to configure the peripheral controller by issuing software generated control codes during system initialization. This means that a relatively small range of programmable controllers can be used for a multitude of applications, thus increasing manufacturing volume on any one device and consequently bringing down its cost to the user.

The concluding paper in the morning session was entitled 'L.s.i. communication interfaces' and was given by Tony Danbury of Mullard. After a very interesting tutorial section on communication concepts, he dealt with hardware elements associated with telephone, microwave and satellite data communications covering the finer points of simplex, half and full-duplex systems. The practical interpretation of envelope delay, distortion, noise, baud rate and band width requirements were described, and he next discussed various types of formating and protocol systems used in modems and the handling of data transparency, error detection and correction. Having introduced the subject and established some of the requirements, definitions and jargon, he then went on to describe a specific communication interface chip recently introduced by Signetics. Problems of different information codes, such as Baudot, IBM, ASCII and EBCDIC, communication protocols such as character controls, BCP, and BOP and message formats were all outlined. He went on to describe how new integrated circuits could now be used to interface most of the popular microprocessors. The functions covered included the 'programmable communications interface' (p.c.i.), the 'multiprotocol communications controller', the 'polynomial generator checker' and the 'enhanced p.c.i.'.

The first paper in the afternoon session was by Mr T. Watanàbe of Hitachi, who demonstrated typical Japanese dedication by flying from Tokyo especially to deliver his lecture. It turned out that he was one of the most experienced microprocessor specialists in Japan, as a designer of m.o.s. and bipolar l.s.i. and more recently as an applications engineer providing technical and marketing support on microprocessor peripherals. He started off by briefly outlining the Hitachi 4-, 8- and 16-bit microprocessors and then introduced the impressive range of peripheral devices that go with them. The most significant of these included a cassette recorder controller, a floppy disk controller that handles single or double density, single or double sided drives and a very impressive c.r.t. controller which seemed to do almost everything.

After the tea break, Andy Piper of CPU Computers gave a very entertaining paper on 'Floppy disks, controllers and interfaces'. The Shugart floppy disk drives were described, along with the basic principles of storage and formating of floppies. The development of the technology was traced and compared with similar evolutions in tape and rigid disk stores. The current state of the art was presented as giving 1·2 Mbytes on 20 cm (8 in) and 500 kbytes of unformated storage capacity on minifloppies. The pros and cons of hard and soft sectoring were discussed as were their implications on both the hardware and software of a system. Controllers for double density recording were described as requiring up to three times as many components as standard units. Controllers incorporating phase-locked loops and write precompensation electronics were described as a requirement to accomplish double density recording.

The speaker also brought to the audience's attention developments in low-cost Winchester technology which provides up to 10 Mbytes/drive in a physical format compatible with the standard floppy. Interfacing requirements were claimed to be similar to floppies, but the cost per bit for on-line storage much lower.

Examples of hardware were demonstrated to illustrate the functional complexity and ingenuity incorporated in these storage systems. In the ensuing discussions the question of bubble memories as a potential competitive technology to disks was debated with considerable enthusiasm.

Some further similarly lively discussion on the whole day's proceedings left an overall impression that the event had been very worthwhile.

K. F. BAKER

Technical News

New Era in Maritime Communications

At a recent international conference in Brighton plans were made to improve and extend maritime communications by satellite. Twenty of the world's maritime nations—including the UK, America, USSR, Japan and the People's Republic of China, discussed a £100M scheme that will use satellites to improve communications with ships around the world.

Currently only about 200 of the world's merchant fleet is equipped to make use of satellite communications. These vessels work to Marisat satellites provided primarily for the United States Navy on which there is limited capacity for commercial use. One of the first tasks of the new organization, called INMARSAT (International Maritime Satellite Organization), will be to provide sufficient satellite capacity so that by the mid-1980s more than 2000 ships will be able to use the service.

Members hope to use two series of satellites to relay messages between Earth stations and ships. Some of the capacity will be provided by satellites owned by INTELSAT (International Telecommunications Satellite Organization), in which the British Post Office has a major share. Other capacity will be obtained from satellites supplied by ESA.

To get the most from the new system the Post Office is to build a new dish aerial at its satellite Earth station on Goonhilly Downs, Cornwall. When complete, the 13 m diameter aerial will transmit messages via satellite to and from ships in the Atlantic, while to extend the service to seafarers the Nordic members of Inmarsat—Norway, Denmark, Sweden and Finland—plan to construct a new satellite Earth station in Norway to work to a satellite over the Indian ocean. These two new installations will provide high quality communications and improved safety to life at sea in the Atlantic and Indian Ocean areas for the European shipping community.

New Speaker Cone Material

Two new ranges of loudspeakers incorporating new developments in acoustic engineering have been announced by Wharfedale: of particular interest is a new cone material specially developed by Wharfedale's engineers for use in the bass/midrange drivers. Called 'mineral filled homopolymer' or M.F.H.P. for short, it possesses several desirable benefits such as an increased stiffness-to-weight ratio, an extremely low colouration factor (far lower than bextrene, for example), and does not require doping in any form. The result is a new drive unit that is claimed to be less prone to break-up, will handle higher power levels and is more efficient.

The T.S.R. range (standing for Total Sound Recall) consists of three models, the 108, 110 and 112, and other features include a new highly efficient soft dome tweeter to complement the bass/midrange driver, contour controls plus time delay compensation, achieved not by building steps into the cabinets but by gently sloping the baffle board by 30° and recessing the treble unit into a specially designed foam lined cavity. This treatment effectively brings the acoustic centres into line and also compensates for time delays introduced by the two crossover units. The result is an improved ambience and spaciousness of the stereo image.

The new Laser range (the name alluding to Wharfedale's use of laser holography to observe the behaviour of speaker cones) also employs the M.F.H.P. cone material in the bass drivers, a specially designed perforated midrange driver and the Isodynamic tweeter, a flat field transducer delivering an extremely smooth and extended frequency response.

High Integrity Alarm Link

One method of transporting oil from a platform in the North Sea to the refineries is to use tanker loading buoys remote from the platform. However, major problems can occur in the event of the tanker being unable to accept oil from the interconnecting pipeline due to an on-board equipment failure while the platform is still pumping. Faced with the potentially expensive consequences of pipeline failure under these conditions, the British National Oil Company commissioned Frazer-Nash (Electronics), of Hersham, Surrey, to design and manufacture a high integrity data link between tanker and platform.

The system utilizes various inputs derived from monitoring gas pressure, control valves, oil pressures and safety circuits etc. and uses duplex u.h.f. telemetry. To ensure that information accepted at the platform is not corrupted a correlation technique is used. This involves retransmission of received data for checking at source and requires four valid comparisons for acceptance. In the event of the platform receiving a validated fail signal an automatic interlock inhibits the pumping operation. A voice channel is also incorporated for priority communication.

The complete telemetry system is duplicated, with extensive self-checking facilities, and in the event of failure, an alternative channel is automatically selected. Manufacture is of a rugged standard to cope with the harsh environment encountered on tankers and platforms in the North Sea.

Standard Frequency Transmissions

(Communication from the National Physical Laboratory)

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

September 1979	MSF 60 kHz	GBR 16 kHz	Droitwich 200 kHz
1	−2·8	3·6	65·4
2	−2·7	3·7	65·1
3	−2·9	4·0	64·6
4	−2·9	3·5	64·9
5	−2·9	3·7	64·6
6	−2·9	3·7	64·3
7	−2·7	3·5	64·0
8	−2·9	3·3	63·6
9	−2·9	3·6	63·2
10	−2·8	3·4	62·9
11	−2·8	3·8	62·6
12	−2·8	3·7	62·3
13	−2·8	3·8	61·9
14	−2·9	4·5	61·6
15	−2·9	3·6	61·3
16	−3·0	3·6	61·0
17	−3·0	3·6	60·7
18	−3·0	4·5	60·5
19	−2·8	4·5	60·1
20	−2·9	3·8	59·7
21	−3·1	3·9	59·4
22	−3·1	4·2	59·1
23	−3·1	4·2	58·8
24	−3·1	4·2	58·5
25	−3·1	4·0	58·2
26	−3·2	3·8	58·0
27	−3·2	3·9	57·5
28	−3·2	4·0	57·2
29	−3·2	3·8	56·9
30	−3·3	4·0	56·6

Notes: (a) Relative to UTC scale (UTC_{NPL} − Station) = +10 at 1500 UTC, 1st January 1977.
(b) The convention followed is that a decrease in phase reading represents an increase in frequency.
(c) Phase differences may be converted to frequency differences by using the fact that 1 μs represents a frequency change of 1 part in 10^{11} per day.

Evaluation of 16-bit Micros

A report published recently in the US indicated that the major growth in microprocessor sales in the next few years is expected to be in the area of 16-bit microprocessors. The report went on to say that it expects the dollar value of these processors to increase from their current level of 6% to 23 % by 1983. This represents an annual growth rate of approximately 60% over the next five years.

This percentage increase in the market share of these, the newest of all microprocessor devices, must be considered against the expected increase of microprocessor sales from $430M in 1978 to over $1,300M by 1983.

In an attempt to anticipate the application for the 16-bit microprocessors before many of them have actually been sampled, or are still in the data sheet stage, *Microcomputer Analysis* has put together an evaluation of these microprocessors and examined six of the leading contenders for market leader.

The report begins by asking the question 'Is it worth it—are 16-bit microprocessors really necessary?' It would seem that the question lies in whether or not one considers the increase in word size to be a boon to throughput, or whether the architectural and programming features to be found in the devices at present being sampled are important. It is true that there are attractive advantages to wider words in memory and c.p.u. registers. Also programmers find wider words easier to manipulate, that is one of the appealing properties of a high-level language. The report points out that the reason the 8-bit microprocessor so quickly undermined the market for 4-bit devices was due almost entirely to improved architecture and the inconvenience of having to program multi-position arithmetic for any integer value larger than 15. The question today is whether such an advantage exists for 16 over 8-bit words (or, indeed, 32 over 16).

None of this may be important since the major disadvantage to wider words is the possible reduction in utilization of data storage; if all the data is in ASCII, a 16-bit computer will probably afford no more throughput than an 8-bit design. If the majority of all data variables could fit within 8-bits, then at least half of the bits in memory are wasted—the high-order bits in each 16-bit word. However, data packing techniques could be used to effectively utilize memory area.

It would seem that it would be advantageous if somewhere in the already vast literature of computing, a chart were available which would show the distribution of necessary word sizes. The report indicates how such a chart could be created and also points out that the chart would analyse a large collection of computer programs written in a language that discouraged the programmer from being cognizant of word size. Such a graph—one is reproduced within the text—would be composed from a wide spread of software programs.

The report points out that word widths of memory are growing according to customer demands. The first parts were organized as a one-bit output; however, designs are now emerging with 4- or 8-bit outputs. It indicates that although the larger word sizes require more pins and consume more chip area, it is customer demand that has forced the semiconductor vendors into these configurations. It also indicates that there has been insufficient demand to warrant introduction of a 16-bit wide memory, nor is there likely to be in the near future. It makes an interesting point that the large consumers of memory components, the mini- and main-frame computers, operate on 16- and 32-bit words, the sizes of their memories warranting the use of high density dynamic devices; static devices are used in smaller computer configurations, e.g. microcomputers. It points out that even manufacturers of 16-bit microprocessors (like Texas and GI) have made no effort to introduce 16-bit wide memories in support of their proprietary c.p.u.s.

In concluding this section, the report indicates that it is possible for the microprocessor architecture to operate efficiently on 8-bit memory words while presenting a 16-bit architecture to the machine language program; these kinds of micros are called bridge products, and will be treated separately in future issues of *Microcomputer Analysis*. The report then goes on to discuss those processors which interface to memory sub-systems along a 16-bit wide data bus. The report shows a diagram of the empirical ranking of 16-bit microprocessors and illustrates the power of the various processors in their relation to one another.

The report continues by pointing out that the 16-bit processors perform a little better than any other computer architecture. They seem to serve the narrow performance gap between 8-bit micros and 16-bit minis, but software and documentation support for the micros will continue to be sparse when compared to the breadth offered by the manufacturers of minicomputers.

Finally, the report examines the Motorola MC68000, Zilog Z8000, Intel i8086, Western Digital WD9000, Texas 9900 and National INS8900. A comparison of these devices indicates that the i8086 device will be the dominant 16-bit microcomputer for dedicated applications in the next three to five years. However, the usual problem will still remain that of software support which will continue to lag behind that traditionally supplied for equivalent minicomputers.

Microcomputer Analysis is published by Mackintosh Publications Ltd, Mackintosh House, Napier Road, Luton LU1 1RG, England (Tel. (0582) 417438).

Broadcast Quality Lincompex Equipment

The first system to overcome fading and interference on high frequency radio links to a standard acceptable by radio broadcast operators has been developed by the Electronics Division of Standard Telephones and Cables at Newport, Gwent. Known as radio relay lincompex (linked compressor and expander) equipment, it provides high quality circuits between main radio stations' studios and local broadcast transmitters, or distant studios and the main broadcasting station.

Similar to conventional communications lincompex, which has been in use for many years, STC's radio relay lincompex system is intended for point-to-point transmission over a 6 kHz audio channel, and has been designed to conform generally to British Broadcasting Corporation specifications. The terminal equipment can be integrated easily into radio relay systems, replacing existing control terminals, provided that the associated transmission facilities meet certain minimum transmission requirements.

A lincompex terminal consists of a transmit or a receive channel, each split into speech and control paths. 'Set and forget' controls and levels are designed into the system for simple operation. Improved performance is said to be immediately noticeable with the system, the fast-acting constant-volume amplifier in the receive circuit effectively minimizing rapid fading. When circuit noise is moderate or low and a programme has an inherently quiet background, the signal-to-noise ratio is considerably improved, with reception clarity approaching that from a local station or line feed.

Other benefits of lincompex equipment include the use of a lower power transmitter to achieve a given signal-to-noise ratio. Because the transmitter does not handle such a large dynamic range of signal level, operation nearer the peak envelope power level is possible. This results in increased average transmitted power.

Letters to the Editor

From: **C. Powell,** C.Eng., F.I.E.R.E.
W. Laycock, B.Sc., C.Eng., M.I.E.R.E.

The Engineer-Author

There is a growing problem within the electronics industry that has, to date, received far too little attention: namely the lack of interest, and of competence, in written and graphic communication among the younger engineers. In an industry based on, and still largely concerned with, communication in one form or another, this is surely an ironic situation. Its practical consequences, however, are a more serious matter.

The provision of adequate and accurate operating, maintenance and repair information is an essential support service for any piece of equipment. Indeed it is generally a contractual requirement, often with an associated penalty clause. The demand for competent Technical Authors is therefore high, yet their numbers are falling and there is now an extreme shortage of younger engineers willing or indeed equipped to join their ranks. Even the specialist technical publications firms, which have tended to offer relatively attractive terms of employment for authors, seem to be feeling the shortage of truly 'literate' engineers.

Undoubtedly much of the blame lies with the industry itself (and perhaps the Institutions are not altogether blameless). For far too long, the fact that technical authorship demands competence both in engineering and in communication has been ignored and the status of the Technical Author has been inferior to that of the 'active' engineer on whom no such dual demand is made. In reality he must be literally a writing engineer: the word engineer is paramount and needs no definition here, but writing in this context signifies skill in diagrammatic presentation as well as in the use of words. While these skills are essential for an engineer undertaking technical authorship, the fact is that they also tend to make him a *better engineer*—something that employers should bear in mind when assessing the usefulness and worth of any engineer.

Probably as a result of this lack of appreciation on the part of industry, a corresponding lack of emphasis on communication has become evident in the education and general training of engineers. In some instances even the correct use of English appears to be considered an affectation. As a consequence, many of those who have entered the industry of late would be quite unsuited to the task of technical authorship even if they were prepared to accept it as a career. Some are ill-equipped (and at times even unwilling) to document their own work adequately: the need for an engineer to keep proper records of his work should surely be part of the basic discipline. When a development engineer fails in this, an even greater demand is made on the expertise and time of the already overloaded author.

If those concerned with the education and training of engineers were to make themselves more familiar with the need for technical authorship and with what it involves, they would find that it is a much more creative activity than is commonly supposed. In having to explain a circuit or a system in detail, the author necessarily subjects the design to a degree of independent critical analysis and in consequence his partnership with the development team can be of real benefit to the quality of the final product.

If the industry is to hold its ground, let alone prosper internationally, it must stop viewing the Technical Author's job as that of a 'failed engineer', for nothing could be farther from the truth. Competence in communication must be accepted as an additional engineering qualification and rewarded accordingly. The future position of the Technical Author will have to be assured so that more engineers can be attracted into this exacting profession; but until he can be accorded his proper place in the professional pecking order, few engineers will willingly forgo the interest and stimulus of R and D for the often more demanding and onerous work of technical authorship. Once this place is recognized and established, one may hope that some younger engineers will be encouraged to change course and so help to relieve the now really serious shortage of people capable of producing the technical documentation on which the manufacturer and the customers depend.

This letter expresses thoughts stemming from many years' varied experience in the technical information field on the part of the undersigned who is head of the group of departments concerned with technical information in a well-known electronics company, and of his colleague, David Hillman who is Technical Publications Manager.

CLAUD POWELL

Little Briars, Cobham Road,
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7th September 1979

The Electronic Engineer and the Arms Trade

I have been concerned for some years by the way in which the practice of electronics has been prostituted from being for the good of mankind to being directed towards death and destruction. It seems foolish that engineers should ask for greater respect and recognition from the public and at the same time be employed in such a wasteful and deadly way.

Whilst a large part of the world is suffering and starving I do not see how it can be right to connive at the maintenance or aggravation of this state of affairs by our misuse of resources in manpower or material and our trade in arms and still claim to be a responsible body of people. It seems to me that an engineer with a wider view of the world as we are constantly exhorted to have must see the folly and hypocrisy of having these two conditions at the same time.

If we want respect and if we claim to be responsible we must see that our talents are used fully for the good of mankind as a whole and we must not allow ourselves to be put off by bland statements about complexity or the possibility of redundancy. Ingenuity and goodwill should overcome these problems and these are two qualities that any engineer should have and be prepared to use.

I would be interested to hear the views of other members on this matter.

WILFRED LAYCOCK

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26th August 1979

Contributors to this issue*

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D. K. Kinell is Technical Supervisor of the silicon-on-sapphire product development project and silicon wafer fabrication operations within the Lockheed Microelectronics Center. He received his bachelor's degree in electrical engineering from the University of Washington in 1964 and an M.S.E.E. from the University of California at Berkeley in 1966. He joined the Lockheed Missiles and Space Company in 1966. In the Lockheed Microelectronics Center, he has been engaged in both bipolar and m.o.s. process and circuit development.

Ludek Kitajewski has responsibility for silicon-on-sapphire and m.o.s. integrated circuit design as a member of the technical staff within Lockheed Microelectronics Center. He received his B.Sc. (Eng.) degree in electronics from London University in 1955 and completed courses in semiconductor physics and transistor circuit design at the Graduate School of Northeastern University in Boston. Mr Kitajewski has been involved in semiconductor work and integrated circuit design since 1963. Prior to joining the Lockheed Microelectronics Center in 1978, he was engaged in the design of l.s.i. m.o.s. circuits including calculators, microprocessors and non-volatile memories.

Dexter Girton has responsibility for silicon-on-sapphire process development engineering as a member of the technical staff within the Lockheed Microelectronics Center. He received his B.S.E.E. in 1960 and M.S.E.E. in 1962 from Ohio State University, where he then taught electrical engineering and worked at the Electro Science Laboratory for three years. In 1969 he was awarded the Ph.D. degree in electrical engineering from Ohio State University, and in 1970 he joined the University of California at San Francisco working on biomedical applications of electrical engineering. Joining Lockheed Missiles and Space Company in 1973, he worked on photolithography and thin/thick film in the Lockheed Microelectronics Center. For the past three years, his work has involved the characterization of integrated circuits in radiation environments.

Michael Windram received the B.A. degree in natural sciences in 1966 from the University of Cambridge; he stayed on to obtain a Ph.D. for postgraduate work on radio astronomy. From 1969 to 1971 he worked on radar systems at Marconi Elliott Avionic Systems and he then moved to the Independent Broadcasting Authority, joining the Radio Frequency Section. He has been concerned with the development of the high performance tunable receiver, and of adaptive aerial arrays which led to an operational array in the Channel Islands. Dr Windram is now Head of the Radio Frequency Section at the IBA.

Philip Kelly is a Deputy Director in the Network Planning Department of the British Post Office where his responsibilities cover the planning of switching systems for both voice and non-voice services. He was formerly the head of its Data Systems Planning Division where he was responsible for the planning and implementation of all new data services including packet, circuit and leased line services. He has represented the British Post Office at SG VII of the CCITT and the CEPT's Special Committee on Data Transmission (CSTD). He is also a Vice-Chairman of the CEPT's Data Transmission Group and Chairman of the multi-national committee established to oversee the planning and implementation of the telecommunications network for EURONET. In the late 60s Mr Kelly was head of the Computer Division of the Treasury, now the Central Computer Agency. He has an honours degree in electrical engineering of the University of London.

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* See also pages 556, 580 and 590.

UDC 551.591
Indexing Terms: Visual range monitors

Visual range monitors

M. E. JUDGE, B.Sc., M.Inst.P., C.Eng., M.I.E.E.*

SUMMARY

The determination of the atmospheric extinction coefficient is necessary to allow visual range to be calculated. The limitations of nephelometers (devices which measure light scattered in the atmosphere) are discussed. These limitations lead to the conclusion that optical transmissometers are the only valid devices for assessing extinction coefficients. New techniques for maintaining the accuracy of transmissometers over many months of use are described. The high accuracies achieved allow a wide dynamic range of visibility to be measured with a single base line instrument.

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1 Limitations to the Visual Range

The propagation of light through the Earth's atmosphere is affected by the presence of aerosols, particles of solid matter and even to a small though measurable extent by the air molecules themselves. Both scattering and absorption processes occur as the light interacts with the components of the atmosphere. These processes are responsible for the existence of a limit to the range at which objects and lights can be seen, that is, distinguished from their surroundings. In the case of lights, the limiting visual range is also dependent on the luminance of the background against which the light is being viewed.

The ability to sense visibility conditions at a remote location can be of great advantage in some situations. The most significant application area from the point of view of the work to be described later in this paper is the measurement of visual range on airfields. Other examples of applications where automatic visibility reporting offers considerable benefits are on motorways and in automatic weather stations. Over a long period of time people have struggled with varying degrees of success to automate the process of determining how far they can see. For the purpose of system design, the basic physical parameters involved in determining visual range have been known for many years. Allard in 1876 described how the illuminance from a light varied with distance and with the transparency of the atmosphere. Koschmieder in 1924 showed how the relative contrast between an object and its surroundings was reduced by the intervening atmosphere. Both Koschmieder's and Allard's Law are discussed in Reference 1. In addition, extensive studies of the psychological and physiological effects involved with 'seeing' have been carried out in order to relate the observer's performance to his monitorable environment.[2]

In spite of this thorough groundwork, numerous visibility monitoring systems have been and are still being designed which fail to produce representative visual range data. In all cases, this failure can be attributed to the fact that due account has not been taken of one or more of a number of essential design criteria. These design criteria, which are summarized in the conclusions, will become apparent as the various problems of visibility instrumentation are discussed.

2 Some Useful Formulae

Reference has already been made to work carried out by Koschmieder and Allard and to laws bearing their names. These laws relate the visual range to measurable parameters of the environment, and will be useful when the assessment of visibility systems is made later.

2.1 Transmission of Light through the Atmosphere

A beam of light traversing the atmosphere is affected by scattering and absorption as the electromagnetic

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radiation interacts with the constituents of the atmosphere. In a uniform atmosphere, the fractional loss of power from the beam per unit length of atmosphere traversed by the beam is constant.

If dP is the amount of power lost from a length dx of a beam containing power P at range x from a source, then

$$dP = -P \,.\, \sigma_{ex} \,.\, dx$$

where σ_{ex} is the extinction coefficient. Integrating gives

$$P = P_0 \,.\, \exp(-\sigma_{ex} \,.\, x) \qquad (1)$$

This equation describes the way in which the power in a beam of light decreases with the distance from the source. The scattering and absorption effects can be characterized by individual coefficients, σ_{sc} and σ_{abs} such that

$$\sigma_{ex} = \sigma_{sc} + \sigma_{abs}.$$

2.2 Koschmieder's Law

Koschmieder's work in 1924 showed how the relative contrast between an object and its surroundings was reduced by the scattering of light in the atmosphere between the object and the observer.

Subsequent work[3] extended the treatment to include absorption as well as scattering. For a dark object viewed against a bright background, the initial contrast is unity. Koschmieder's Law states that this contrast will be reduced to a level equal to the transmittance of light for the path length between object and observer. If E_c is the apparent contrast of the object, then using equation (1) we can write

$$E_c = \exp(-\sigma_{ex} \,.\, x)$$

This is a statement of Koschmieder's Law. When an object is at the limit of the visual range, V, it is customary to consider that the relative contrast has been reduced to between 2% and 5% of its initial value. Taking the 5% figure, the previous equation can be written as

$$0{\cdot}05 = \exp(-\sigma_{ex} \,.\, V)$$

which leads to the simple relationship between visual range and extinction coefficient.

$$V = 3/\sigma_{ex} \qquad (2)$$

2.3 Allard's Law

Allard's Law describes the manner in which the illuminance from a light decreases with range, R, from the source. Mathematically the law is

$$E = (I/R^2) \,.\, \exp(-\sigma_{ex} \,.\, R)$$

where I is the luminous intensity of the source and E is the illuminance at a distance R.

The observer's eye will only be able to distinguish the light source from its surroundings if the illuminance at the eye is above some threshold level E_t. The visual range, V, of the light is, therefore, given by the formula

$$E_t = (I/V^2) \,.\, \exp(-\sigma_{ex} \,.\, V) \qquad (3)$$

The eye's illuminance threshold is a function of the background luminance B. Blackwell's work[4] established the relationship as

$$\log E_t = -7{\cdot}0 + 0{\cdot}89 \log B$$

where E_t is measured in lux, and B is measured in $cd.m^{-2}$.

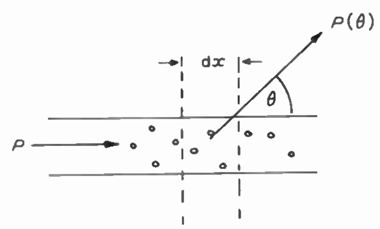
3 Techniques

The previous Section showed how the visual range is related to atmospheric extinction coefficient for Koschmieder's Law, equation (2), and for Allard's Law, equation (3). From these two expressions it can be seen that the instrumental assessment of visual range depends on determining a representative value of σ_{ex}, the atmospheric extinction coefficient. The use of the word representative here is significant, for the concept of visual range as experienced by a human observer relates to the condition of the atmosphere over a distance between the observer and a particular location in a particular direction. It will become apparent from this discussion of techniques that a sensor measures the extinction coefficient in a relatively restricted region of the atmosphere. Fortunately for some sensors and some conditions this difference between an observer and a sensor is less of a disadvantage than it at first appears as it is possible to improve significantly the representativeness of the measurement by using time averaging to replace the observer's spatial averaging. The following Sections consider the two main techniques available for measuring extinction coefficient: nephelometers, which measure light scattered in the atmosphere, and transmissometers, which directly assess the atmospheric extinction coefficient by measuring the transmission of light through the atmosphere.

3.1 Nephelometers

In terms of the complete definition of the visual range problem these instruments are really non-starters in that they try to assess the extinction coefficient by a measurement of the amount of light scattered from a beam traversing the atmosphere. This means that the contribution to extinction made by absorption is ignored. There are some situations, however, in which it is arguable that the absorption contribution to total extinction is negligible. Bearing in mind that they have a restricted application in the field of visibility instrumentation, therefore, their advantages and limitations will be examined.

Nephelometers are attractive to system designers because of their generally compact construction and because the output signal from the device varies linearly with the scattering coefficient. Most of the inherent limitations of nephelometers arise from the impossibility



Total power scattered per unit length of beam $\frac{dP}{dx} = P \cdot \sigma_{sc}$

Angular distribution of scattered power (power per steradian at angle, θ) $P(\theta) = \frac{P \cdot \sigma_{sc}}{4\pi} \cdot f(\theta)$.

Fig. 1. Definition of scattering coefficient, σ_{sc}, and angular scattering function, $f(\theta)$.

of measuring all the light scattered from a beam traversing the atmosphere. The subject of errors introduced by measuring only part of the scattered power has received considerable attention. Most of the analyses that will be referenced show how sensitive this error is to changes in the aerosol distribution. Figure 1 illustrates the scattering geometry convention that will be used, with scattering angle, θ, measured with respect to the direction of propagation, and the scattering function, $f(\theta)$, describing the angular distribution of the scattered power. Computed scattering functions for individual water drops illuminated by radiation of a single wavelength are extremely complex. Figure 3 shows typical forms obtained experimentally for the scattering function. These scattering functions illustrate the averaging effect that results from using a broad spectral source to illuminate a polydisperse aerosol.

Three categories of nephelometers will be considered—integrating nephelometers, which attempt to measure the power scattered from a beam over a wide range of angles; backscatter instruments, measuring power scattered at or around 180°; and instruments measuring power scattered over a narrow range of angles usually in the forward direction.

3.1.1 Integrating nephelometers

The general principles of this type of instrument are illustrated in Fig. 2. A sample of the atmosphere is illuminated by a source of light. Power scattered from the sample of atmosphere in the field of view of the

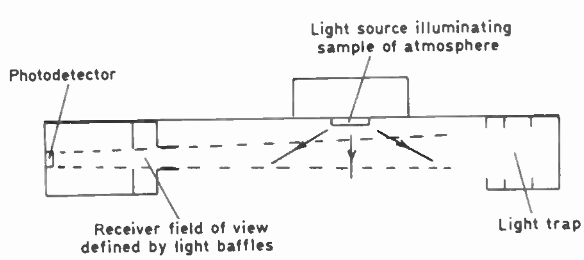


Fig. 2. General principles of an integrating nephelometer.

receiver is detected by the receiver. Geometrical constraints limit the range of scattering angles involved in the measurement to between 7° and 173° for a well designed instrument.[5] Because of the complexities of aerosol scattering and the shape and non-uniform illumination of the scattering volume, a theoretical calibration of this instrument is difficult. Calibration techniques usually depend on comparison with observer estimates of visual range or on the use of scattering surfaces inserted in the beam.[6,7]

A number of studies of the systematic error introduced by omission of the forward and backscattered components have been made.[8–11] The sensitivity of the systematic error to changes in aerosol population is evident with errors varying from ±10% to ±30%[11] for a range of aerosol types which may be only part of the range to occur naturally. Major changes in the aerosol scattering function near 0° and 180° can be seen in Fig. 3 which is taken from Barteneva's work.[12] Moreover, this particular set of scattering functions was obtained in aerosols for which the visual range value was the same.

3.1.2 Backscatter instruments

Measurement of backscatter power provides a convenient instrumental arrangement for transmitter and receiver. Many devices of this type have been developed and a typical configuration is shown in Fig. 4. As in the integrating nephelometer, the sample of atmosphere involved in the measurement is defined by the overlap of the transmitter beam and receiver field of view.

Backscatter instruments, and others using only a narrow range of angles, depend for their successful operation on the existence of a good correlation between total scattered power and power scattered over that narrow angular range. In the case of scatter at or near 180°, the correlation is not good.[13–15] Again, Fig. 3 provides some idea of the extent of the problems. Figure 5, taken from reference 15, illustrates the likely variations to be found.

3.1.3 Forward scatter instruments

A number of workers studying scattering behaviour have reported the existence of a correlation between the scattering coefficient and the amount of power scattered at or near 45°.[16,17] This phenomenon has formed the basis for a number of visibility sensors.[18,19] Theoretical assessments of accuracy suggest visibility values can be determined with r.m.s. uncertainties of ±12%,[19] although these error estimates have been carried out for aerosols of a limited range of drop sizes, and take no account of the possible presence of non-spherical particles (dust, sand, etc.).

The bulk of the experimental support for the existence of a correlation between power scattered around 45° and total scattered power comes from the extensive Russian

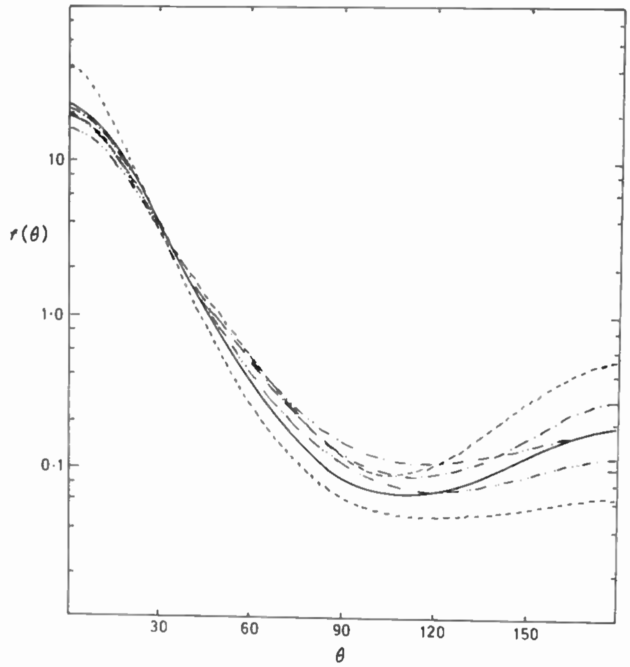


Fig. 3. Scattering functions for a visual range of 1 km (taken from reference 12).

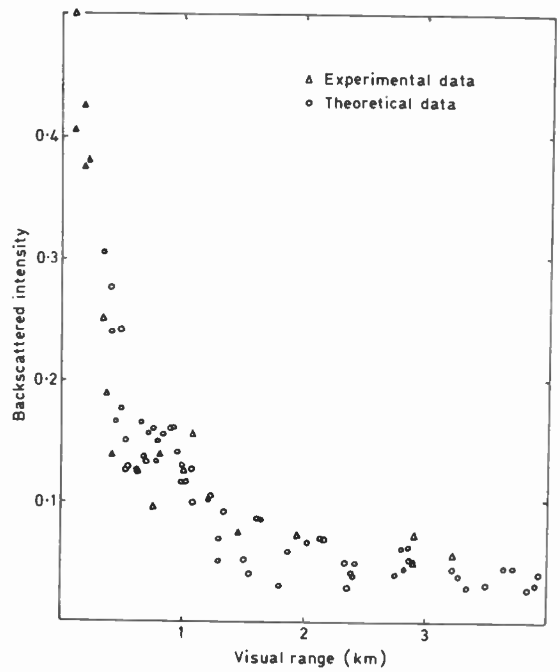


Fig. 5. Correlation between backscatter and visual range (taken from reference 14).

measurement programmes.[12, 20] Barteneva's work studied scattering in fogs for visibilities between 160 m and 220 km. Although a considerable amount of data was collected for the visibility range 160 m to 400 m, no analysis of 45° scatter was carried out by Barteneva for visibilities below 480 m. For visibilities in excess of 500 m the value of $f(45^\circ)$ is given as 1·5 with a standard deviation of $\pm 16\%$.[17] Later measurements by Barteneva and Polyakova[12] show unexplained variations in $f(45^\circ)$ which should increase the uncertainty. For visibilities of 1 km and 2 km, their data give an average value of 1·2 for $f(45^\circ)$. For the range 150 km to 1 km, however, this average value increases to approximately 1·5, in general agreement with the earlier work of Barteneva.

3.2 Transmissometers

In Section 2 the transmission of a beam of light through the atmosphere was described by an equation of the form

$$P = P_0 \,.\, \exp(-\sigma_{ex} \,.\, x)$$

The transmissometer assesses extinction coefficient by measuring the variations in light power transmitted over the instrument base length, B.

$$T_B = \exp(-\sigma_{ex} \,.\, B)$$

where T_B is the ratio of power received in foggy weather to power received when the air is clear.

Using equation (2) from Section 2 enables the following transmissometer equation to be written

$$V = -3B/\ln(T_B).$$

The transmissometer makes a direct assessment of extinction coefficient, including both scattering and absorption contributions in the measurement, and assesses extinction correctly whatever the type of condition causing reduced visibility—snow, dust, sand, rain or polluted fog.

The two forms of transmissometer available are illustrated in Fig. 6. In the first, a transmitter and receiver are used, one at each end of the instrument base line. In the second type, a folded measurement path is provided, with the transmitter and receiver combined in one optical unit. At some distance from this combined unit a retro-reflector is positioned to return transmitted light to the receiver. The method of sorting out the transmitted and received light using a beamsplitter is shown schematically in Fig. 6. Although this method of

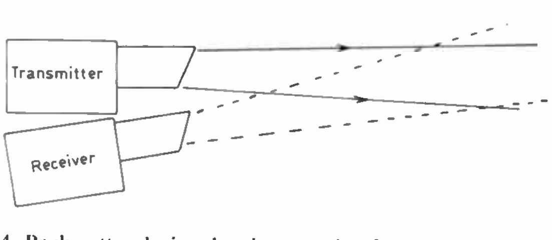


Fig. 4. Backscatter device showing sample of atmosphere defined by overlap of transmitted beam with receiver field view.

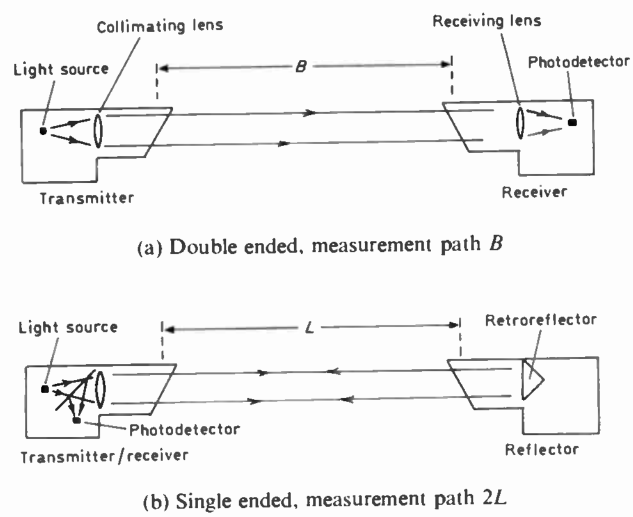


Fig. 6. Basic transmissometer designs.

construction appears the more complex of the two, it is the preferred design, offering distinct advantages when calibration techniques are considered.

4 Transmissometer Problems and their Solution

With transmissometers apparently providing an ideal solution for visibility instrumentation, why should nephelometers have ever been considered for use at all, suffering as they do from limited applicability and uncertain accuracy?

The two main reasons have already been put forward. The first is the compact construction offered by nephelometers, although this is less an advantage than it at first seems when the need for an adequate size of atmospheric sample is considered. The second, more valid, advantage is the linearity that is obtained between sensor output signal and scattering coefficient.

In comparison with nephelometers, practical designs for transmissometers lead to instrument baselines of many metres or tens of metres. In addition the transmissometer equation derived in Section 3 shows that the extinction coefficient is a logarithmic function of the atmospheric transmission. In the past this logarithmic relationship has seriously limited the dynamic range available from a fixed baseline transmissometer.

Fortunately, new developments in data processing techniques have allowed this particular limitation to be overcome.

It must be emphasized here that the techniques to be described are novel. The optical sensor on its own is conventional and inaccurate. By enclosing the optical sensor in a control loop with a digital processor, automatic self-calibration techniques are made possible. This self-calibration maintains high measurement accuracy over long periods of unattended operation and in the presence of the inevitable contamination of external optical surfaces.

4.1 Transmissometer Accuracy

The transmissometer equation was derived as

$$V = -3B/\ln(T_B)$$

Differentiating gives

$$dV/V = (V/3B)\,.\,(dT/T) \qquad (4)$$

It can be seen that fractional errors in atmospheric transmission are multiplied by a factor $(V/3B)$ to give fractional errors in visual range. Equation (4) has been plotted in Fig. 7 for $dT = 0{\cdot}02$, with transmissometer baselines of 20 m and 50 m. The conflicting requirements on instrument baseline are apparent. Long baselines are required to provide useful data at longer visual ranges. With long baselines, rapid increase in errors at shorter visual ranges arises from the low transmission values involved. Smaller errors in transmission allow useful visibility information to be obtained over a larger range of visibilities. The improved performance of a 20 m baseline transmissometer with an error of 0·003 (0·3% of full scale transmission) is shown by the lowest line in Fig. 7.

4.2 Improving Transmissometer Accuracy

Having accepted that transmissometers are the only suitable sensor for use in a visibility monitoring system, then it is apparent that they must be made more accurate to improve their usefulness. Significant improvements in accuracy have been achieved over the last few years by incorporating digital processing electronics into the

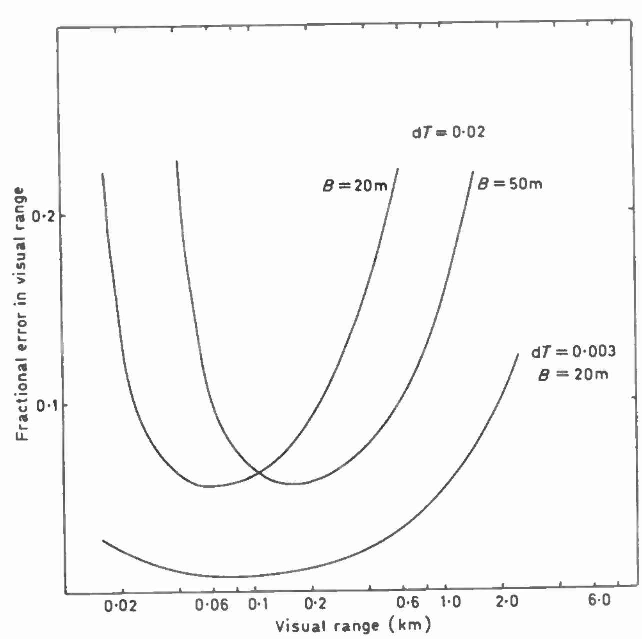


Fig. 7. Visual range error as a function of visual range for various transmissometer baselines and full scale errors.

design of the transmissometer system to operate calibration devices and process signal voltages to produce atmospheric transmission data. For the early system designs, the optical sensor was only one of several sensing devices under the control of a general-purpose minicomputer. Later devices that have been developed incorporate digital processing electronics as part of the field sensor. Details of the calibration processes used in the transmissometer are described in the next Sections. The first is based on the use of a built-in reference reflector for short-term drift correction. The second process is based on the identification of periodically occurring clear air conditions for use as a 100% transmission reference. With these techniques, which were developed for airfield visibility monitoring systems, the optical sensor can provide highly accurate transmission values over many months of unattended operation in the field.

4.2.1 Short term calibration with a reference reflector

Operation with a folded baseline instrument enables all the active components of the transmissometer to be contained in one unit—the transmitter/receiver unit. The basic principles of the optical arrangement of this unit are illustrated in Fig. 6(b). If a reference reflector is now introduced into the optical path outside the collimating lens, a signal will be produced at the output of the receiver. If any changes occur in the output of the light source or in the sensitivity of the receiver, or if contamination of the collimating lens occurs, then the reference reflector will provide a direct assessment of these changes. By recording changes in the reference reflector signal, subsequent voltages resulting from measurement of the atmospheric path can be modified to take account of changes in the instrument characteristic. In a similar way, using a shutter to mask the light source, allowances can be made for variations in the sensor noise level, thereby allowing the instrument to provide useful data down to very low voltage levels. The process is illustrated graphically in Fig. 8. The instrument characteristic is defined by the voltage derived from the reference reflector (V_{ref}) and from the noise level check (V_n). The reference reflector is assigned some arbitrary, but fixed, transmission value, T_{ref}. Transmission values are calculated from signal voltages, V, by the expression

$$T = T_{ref} \cdot (V - V_n)/(V_{ref} - V_n).$$

Changes in the value of V_{ref} and V_n cause the instrument characteristic to be redefined in the computer.

4.2.2 Long term calibration using clear air

The process just described, using a reference reflector introduced into the optical path periodically by the computer, enables compensation to be made for any drift in characteristic in the instrument or the data collection system.

What can be done about changes in the reference reflector itself, and, more important, how can allowance be made for contamination of the retro-reflector at the other end of the instrument base line? To simplify the following explanation, it is worthwhile rewriting the transmission equation in the form

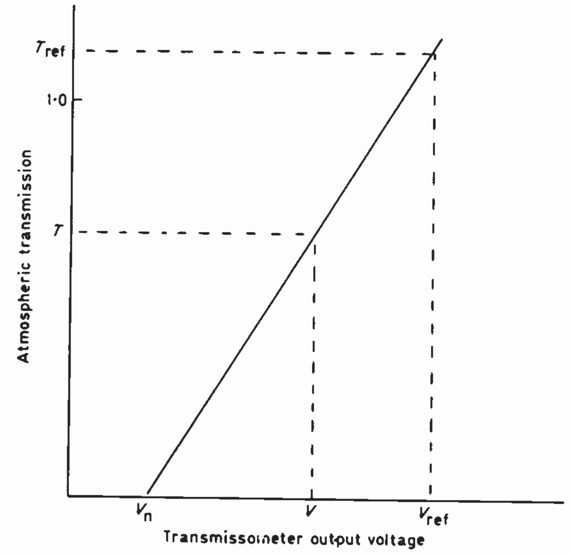


Fig. 8. Transmissometer characteristic.

$$T = \frac{T_{ref}}{(V_{ref} - V_n)} \cdot (V - V_n).$$

$$T = (V - V_n)/k$$

where k is a scaling factor equal to $(V_{ref} - V_n)/T_{ref}$.

Dealing first with the path reflector, it is apparent that any contamination is interpreted by the computer as a reduction in atmospheric transmission. The computer is, therefore, programmed to reduce periodically the value of the scale factor, k. This reduction in scale factor is at a very low rate, but sufficient to be in excess of the actual contamination rate of the path reflector. In this way the transmission values derived from the instrument voltages become slightly optimistic at a rate governed by the difference between the estimated and actual contamination rates. When clear weather occurs, the computer will begin to calculate transmission values in excess of 100%. Each time this occurs the scale factor will be increased to reduce the calculated transmission to just less than 100%, this being the best available estimate of the true value of scale factor, and an improvement on the scale factor previously being used. The magnitude of the optimism introduced by this process is adequately contained by closely matching the contamination correction to the actual contamination rates, and by the frequency of occurrence of short periods of relatively clear air corresponding to visibilities in excess of 10 to 20 km. Results obtained from transmissometer systems installed at a number of sites in the UK show that errors from this process are unlikely to exceed 0·3% of full scale transmission.

Although the reference reflector is kept in a protected enclosure and heated to prevent condensation, it is still possible that over periods of many months some slight contamination may occur. Any decrease in signal from the reference reflector will lead to increases in the calculated transmission value, and will therefore, be corrected by the same overscaling process used to compensate for path reflector contamination. This clear weather rescaling process operates continuously throughout the operational life of the system. Relatively short periods—perhaps only a few seconds—of clear air are capable of being identified by the computer and used to renormalize the instrument characteristic to an absolute standard of 100% transmission.

5 Runway Visual Range (RVR) Measurements

The calibration processes described have been successfully incorporated in two transmissometer systems. The first of these systems, the Marconi Instrumented Visual Range System, IVR–1, for measuring visibility at airfields, will now be described. The second, a short base line transmissometer, will be described in Section 6.

5.1 Visual Range at Airfields

The measurement of visual range at airfields is an important application area for automatic visibility monitoring equipment. Runway Visual Range is specified in the Air Navigation Regulations as the maximum distance in the direction of take off or landing at which the runway or lights delineating it can be seen from a point 15 feet above its centre line. Aircraft operations are grouped into categories dependent on the level of instrumentation in the aircraft and on the airfield. Briefly, clearance for operation in category 1 conditions implies that the RVR value is 800 metres or greater, category 2 is between 400 metres and 800 m, and category 3 is below 400 metres. The RVR definition, based as it is on the estimation of the horizontal visual range from a point close to the ground, cannot account for problems that may arise for a pilot on approach who views the runway along a slant path through the atmosphere. Fog studies (reviewed in reference 21) indicate that the vertical characteristics of fogs have some degree of regularity about them, with extinction coefficients tending to increase with height above the ground until the top of the fog layer is reached. In these conditions the pilot's slant visual range would be less than the RVR value assessed over a horizontal path near ground level. This factor has been taken into account in the decision to fix the category boundaries at the values specified. The instrumental assessment of RVR, therefore, involves the determination of the horizontal visual range value of lights or runway markings from a point as close to the runway as is consistent with safety.

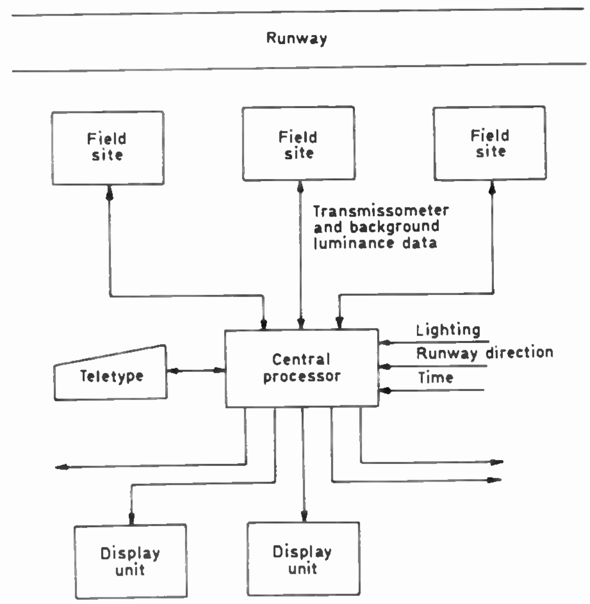


Fig. 9. Block diagram of three field site RVR system.

5.2 System Details

Full details of the hardware used in the airfield equipment can be found elsewhere.[22] Only a brief description will be given here. A block diagram of a typical installation is shown in Fig. 9. Three folded baseline transmissometers are positioned close to the runway. At one of the field sites a photometer provides background luminance data used in the determination of the visual range of the runway lights. One of the field sites is shown in Fig. 10. The transmissometer, operating over a folded baseline of 2×10 m, is housed in the large fibreglass enclosures. The weatherproof box provided for the photometer and data transfer equipment can also be seen. Analogue signals from the field sensors are digitized and passed to the central processor over the data link. The central processor also receives information about the runway light intensity, enabling RVR values to be calculated for each of the three runway locations.

The organization of the transmissometer calibration processes is essentially as described in Section 4. The reference reflector is mounted on a rotary solenoid outside the transmitter/receiver unit and is brought into the main beam of the collimating lens by a command transmitted to the field site from the central processor. The timing of calibration events is arranged so that no significant amounts of drift due to ambient temperature changes can occur in the instrument characteristic between calibrations.

Every 24 hours small adjustments (0·1% of full scale or less) are made to the scale factor value stored in the computer to compensate for optical contamination. Knowledge of the likely contamination rate is necessary as the actual amount of periodic adjustment must be slightly in excess of the expected contamination rate.

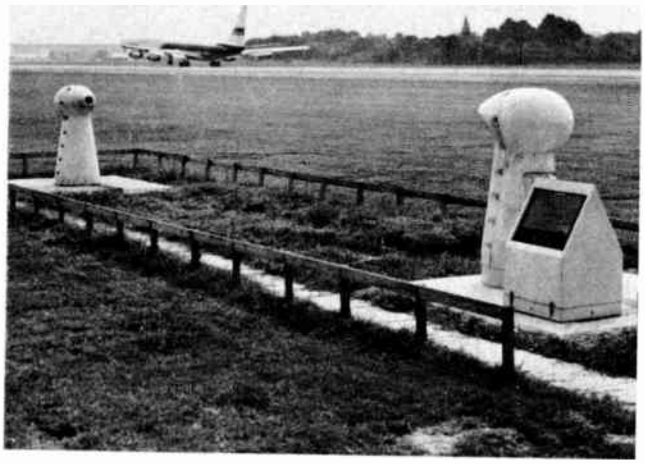


Fig. 10. IVR field site.

5.3 Calculation of Visual Range

After calculating atmospheric transmission using the formula in Section 4.2.1, visibility values according to both Koschmieder's and Allard's laws are derived. The use of Allard's law presents some additional problems as an analytical solution for visual range is not possible. Instead an iterative process is carried out, based on Newton's approximation method, using the Koschmieder visibility value as an initial solution.

5.4 Data Averaging

A potential problem arises with almost all instrumented visual range systems because the volume of the atmosphere sampled by the optical sensor is necessarily small when compared with the sample of atmosphere involved in a human observer estimate of visual range. In stable, patchy fog conditions which can be produced by variations in topography, any type of sensor in a fixed location can only provide a local description of the state of the atmosphere. In some circumstances, however, conditions are favourable for the use of systems with fixed sensors to provide visibility data which is representative of a large area. In particular the airfield environment can be remarkably uniform, comprising as it does large expanses of flat terrain. Prevailing winds move large amounts of the atmosphere past the sensors in relatively short periods of time. With a sufficiently high sampling rate and adequate averaging capability in the data processing system, small scale and large scale inhomogeneities in the fog are adequately sampled and averaged. The success of time averaging to replace the spatial averaging of an observer thus depends on the use of a high initial sampling rate by the optical sensor, and on the use of data processing to vary the averaging time constant in step with the general behaviour of the fog. Observation of fog structure has shown that large variations in fog extinction coefficient can occur from instant to instant in the sample volume. If the sampling rate used by the instrument is too low, the mean value of extinction coefficient derived by the instrument will have a large uncertainty associated with it. In the airfield equipment described, the transmissometer light source is modulated at 3·9 kHz. The received signal is then averaged in a low-pass filter with a bandwidth of approximately 0·5 Hz. The noisy structure of this averaged signal is apparent from Fig. 11(a) which shows variations in visual range calculated from consecutive samples of atmospheric transmission taken at intervals of 1·6 seconds. The effect of further averaging using the central processor can be seen in Fig. 11(b), where individual samples were averaged over 1 minute periods.

To achieve a further significant reduction in 'noise' in the reported visibility data compatible with an acceptable speed of reporting significant changes in visibility, the computer is used to process transmission values with a running, weighted mean.[22] This allows a long time-constant to be used when atmospheric conditions are essentially steady, but introduces a short time-constant when significant changes in the general fog situation are detected. The influence of this more sophisticated processing can be seen in Fig. 11(c), which shows the data of Fig. 11(a) processed in the manner described.

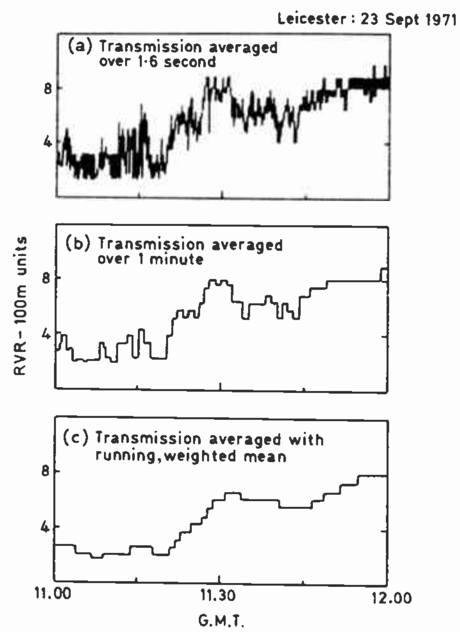
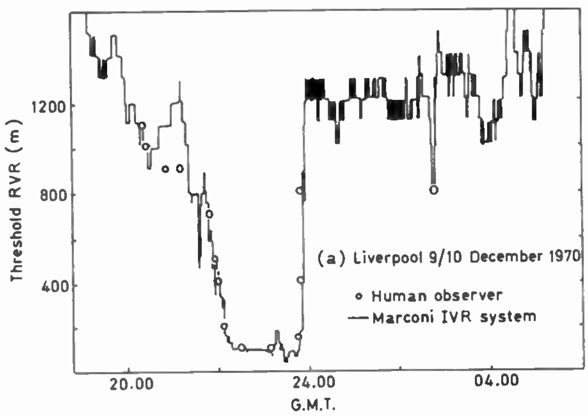


Fig. 11. Visual range variations with different averaging techniques.



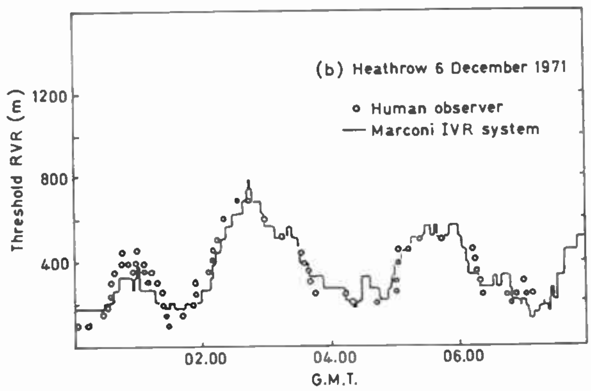


Fig. 12. Comparison of visibility data from Marconi IVR system and human observers.

For systems limited by a low initial sampling rate, is it possible to extend the averaging period to improve the accuracy of the derived visual range values? It would appear from the variability of the basic signals obtained from the RVR equipment, that systems sampling at rates of the order of once per second (pulsed flash lamps for example) would require averaging periods possibly as long as several tens of minutes to produce valid mean values. Such an averaging period would be tolerable in stable fog situations. When visibility conditions were unstable, however, long time constants would cause unrepresentative visibility data to be produced.

5.5 Results

The previous Sections have explained in some detail how much attention has been paid in the design stage to overcoming the known limitations of visibility monitoring systems. Transmissometers were selected as being the only valid technique for determining atmospheric transmission. Computer-controlled calibration processes were devised to ensure that data were obtained of sufficient accuracy to be useful. A high sensor sampling rate and additional computer averaging of results were used to derive representative visibility values. To determine how successful these techniques were, the visibility systems were evaluated over two fog seasons (1970–71, 1971–72) by the Civil Aviation Authority on a number of major UK airfields. Visibility values from the systems were compared with data provided by human observers stationed near the runway. The observers counted the number of runway edge lights that were visible and this information was converted into a runway visual range value.[24]

Some results from this evaluation programme are presented here (Fig. 12) by permission of the Civil Aviation Authority. The degree of correspondence between the man and the machine over a wide range of operating conditions demonstrates the validity of the system design approach adopted.

6 Short Base-line Transmissometer

6.1 Further Transmissometer Developments

After the success of the airfield visibility instrumentation it became apparent that the new transmissometer techniques could be adapted to the solution of other visibility monitoring problems. Numerous areas of application were found which required the same reliable approach to deriving visibility data, but either did not require or could not accommodate the size and sophistication of the airfield transmissometer. This situation provided the incentive to develop a short base line sensor, incorporating the calibration techniques developed for the airfield system. This device has been called the Marconi Environmental Transmissometer, MET–1. By keeping the hardware design simple, it has been found possible to reduce inherent transmissometer instabilities to a level that enables a long-term, r.m.s. transmission error of the order of $\pm 0{\cdot}1\%$ of full scale to be achieved.

6.2 Optical Sensor

The optical transmitter/receiver unit and a reflector unit are mounted on a common rigid framework, (Fig. 13). Measurement of atmospheric transmission is

Fig. 13. MET–1.

carried out using an electronically modulated light emitting diode (l.e.d.) illuminating a retro-reflector at the other end of the instrument base line. The l.e.d. is modulated at a frequency of 1 kHz, and the wavelength of the light emitted is centred on 680 nm in the red region of the visible spectrum. The radiation returned from the retro-reflector is collected by the transmitter lens and focused via an optical beam splitter onto a silicon photodiode (Fig. 14). Electrical signals from the photodiode are amplified and filtered using a phase sensitive detector. A calibration mirror, brought into the optical path by a rotary solenoid, is situated on the transmitter/receiver unit just outside the collimating lens. Excessive contamination of the external optical surfaces is prevented by the use of dust baffle tubes.

6.3 Digitizing and Processing Unit

The digitizing and processing unit provides a means of converting the analogue voltages from the optical sensor into atmospheric transmission values. Compensation for changes in the system characteristic is carried out in the same way as described for the airfield transmissometer. The processor logic for converting input voltages, V, into atmospheric transmission values, T, is shown in Fig. 15.

Input voltages are digitized using a variable gain analogue-to-digital converter. Control signals to operate the reference reflector solenoid are generated periodically. These reference reflector signal voltages are digitized and compared with a reference value held in the calibrate store. Differences between the new and reference value are used to modify the system gain until the difference is reduced to zero. Periodically the reference value held in the calibrate store is increased. This leads to an increase in system gain as a means of compensating for reflector contamination. If transmission values in excess of 100% are calculated, the calibrate store value is decreased until the computed transmission is acceptable. Transmission values are made available from this unit in the form of TTL compatible, 10-bit parallel words. It can be seen that information about the current condition of the system characteristic is held in the calibrate store. As these data are vital to the continued operation of the instrument, a small battery is included in the circuit to retain the contents in the store in the event of a temporary interruption of the power supply.

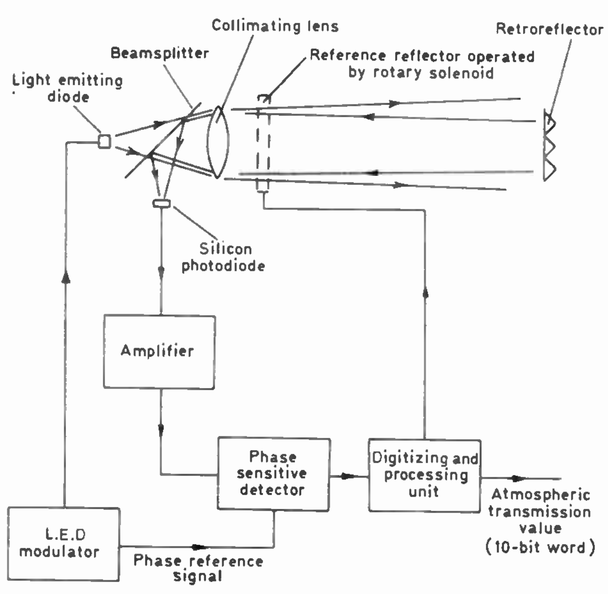


Fig. 14. MET–1 transmissometer system.

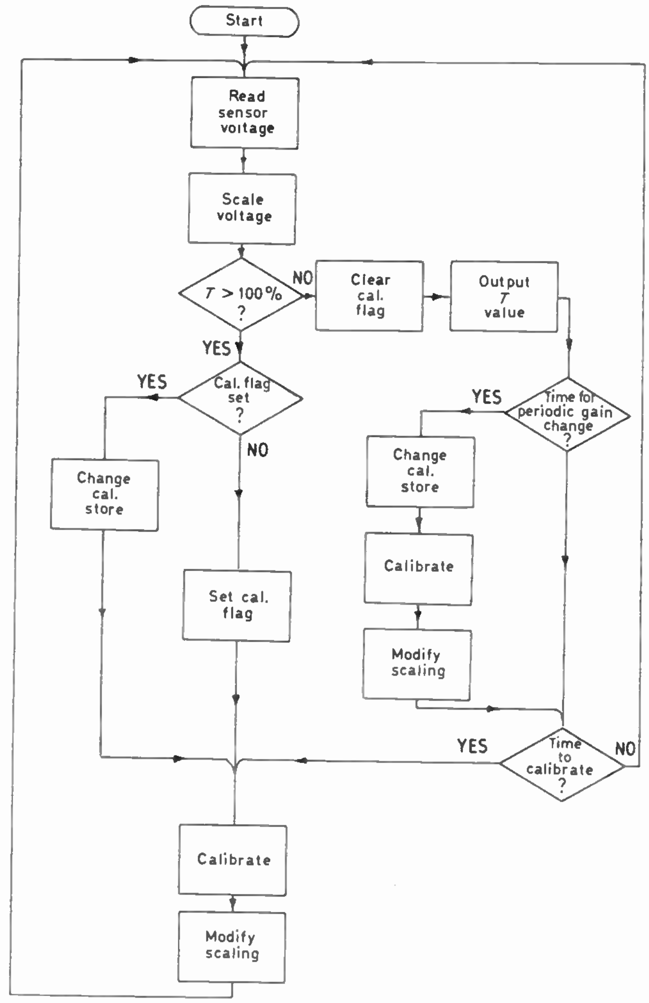


Fig. 15. MET–1 processor flow diagram.

6.4 Results of Evaluation Trials

Evaluation trials were carried out on the Marconi field site at Leicester during the winters of 1974–75, 75–76 and 76–77. The basis of this performance evaluation was a comparison between the MET–1 system and an airfield visual range system. The RVR transmissometer was certified for airfield use by the Civil Aviation Authority.

Typical r.m.s. atmospheric transmission errors for the RVR transmissometer operating over a 2×10 m baseline

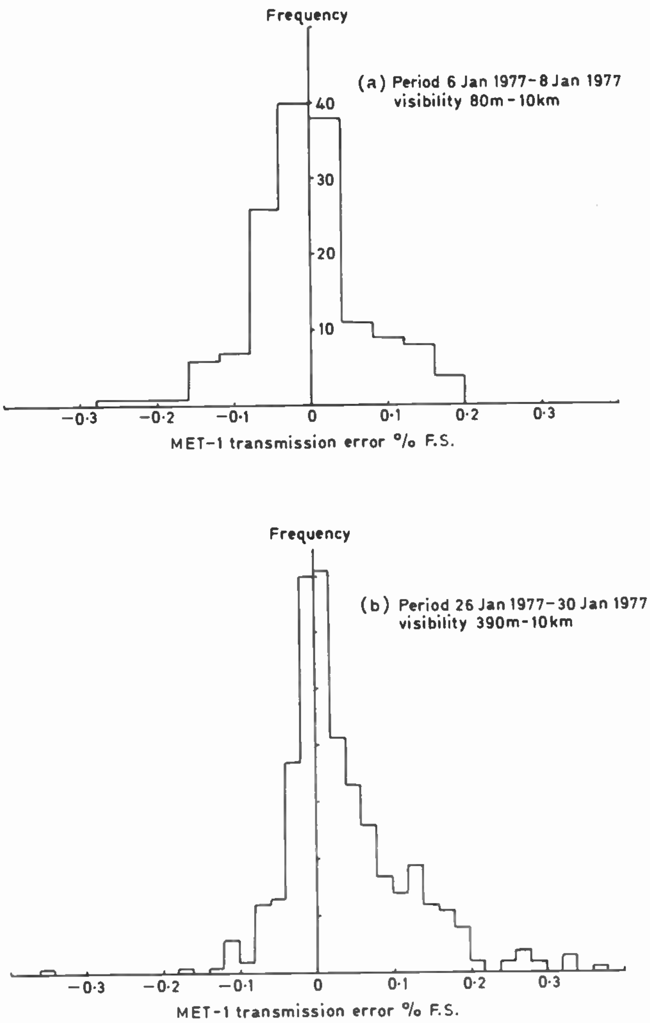


Fig. 16. Distribution of MET-1 transmission errors.

are $\pm 0{\cdot}2\%$ of full scale. This uncertainty in transmission is equivalent to an uncertainty of $\pm 0{\cdot}04\%$ of full scale for the short base line sensor operating over 2×2 m.

The two transmissometers were arranged with their measurement paths horizontal at a height of 1·5 m above the ground. The separation between the two systems was 3 m. Analogue data from the RVR transmissometer were digitized and passed to the computer for processing to give atmospheric transmission values. For the initial phase of the trials, analogue data from MET-1 were treated in the same way, thus providing the opportunity to evaluate fully the characteristics of the optical sensor. When sufficient data on the behaviour of the sensor had been obtained, the various factors determined were incorporated in the hard-wired digital processor unit. The second phase of the trials was carried out with the short base-line sensor fitted with the processor unit. Atmospheric transmission values from the field site were recorded by the computer.

Samples of data from the two phases of the evaluation trials are shown in Fig. 16. The histograms show the distribution of atmospheric transmission errors in the data from the short base-line transmissometer. These errors are assessed on the basis that the RVR transmissometer is providing an accurate value for the extinction coefficient. For the period 6th–8th January 1977, the optical sensor was passing data to a PDP-8L minicomputer for processing into atmospheric transmission values. This particular result was obtained after 60 days of unattended operation of the instrument. In the second sample, taken from the end of January 1977, the digital processing unit was producing atmospheric transmission values. For both periods, atmospheric transmission values were taken every three seconds and averaged over 15 minutes to produce one data point.

The results from 6th–8th January 1977 have been presented in a different way in Fig. 17. This graph shows the correlation between the visibility values (Koschmieder's law) derived from the two transmissometers. The other interesting feature that can be seen in these results is that the visibility values derived using red light centred at 680 nm, are the same as those from a C.I.E. corrected system (within the transmissometer accuracy specification) over a wide range of visibilities. This confirmation is particularly important for visibilities in excess of 1 km, where small deviations might have been expected to occur. Work is now in hand on improvements to the MET-1 design. In order to extend the useful dynamic range, longer base-line versions—still on a rigid framework with single-point, site installation—are being produced. Microprocessors are now replacing the discrete digital logic circuits in the processing unit, enabling more sophisticated data

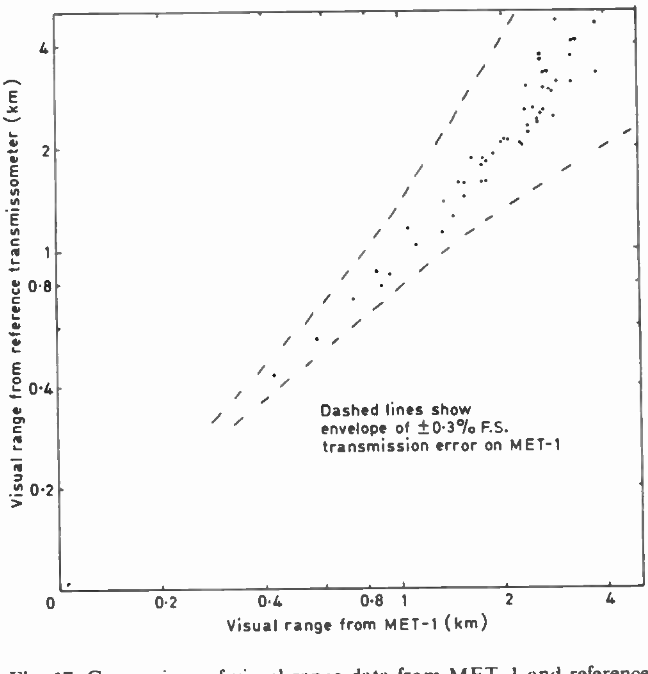


Fig. 17. Comparison of visual range data from MET-1 and reference transmissometer; 6th January 1977–8th January 1977.

averaging to be incorporated together with both Allard and Koschmieder visual range calculations.

7 Conclusions

The design of visibility monitoring equipment has been described. The various techniques available for measuring the relevant properties of the atmosphere were considered. The inherent limitations of nephelometers meant that the use of transmissometers was the only valid solution to measuring the extinction of light in the atmosphere. Difficulties with transmissometers associated with accuracy at longer visual ranges were overcome by the development of automatic calibration techniques, controlled by a minicomputer. The availability in the system of the processing capability of the computer enabled relatively complex data averaging techniques to be incorporated. This additional averaging used a variable time constant to bring about a correspondence between the temporal averaging of the transmissometer and the spatial averaging of the human observer. Finally, these transmissometer calibration techniques have been applied to the design of short baseline sensors, enabling transmissometers to be used for a wider range of visibility monitoring applications where site location problems previously restricted their use.

8 Acknowledgments

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Indexing Terms: Television equipment, testing, Transmitters, Receivers, Modulators

The design and use of high performance u.h.f. transmitters, receivers and demodulators for television r.f. measurement

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Based on a paper presented at the IERE Conference on Television Measurements in London on 21st to 23rd May 1979.

SUMMARY

As the number of transmitters in operation in the broadcasting network is increased, so there is an ever-increasing requirement to provide more sophisticated test and monitoring equipment. To meet this need, the IBA has developed three equipments, the tunable receiver, tunable demodulator and tunable test transmitter. The requirements, equipments, design and application are discussed in this paper.

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1 Introduction

The total number of television transmitters operated by the Independent Broadcasting Authority is at present around 400 and is expected to rise to about 650 by completion of the broadcasting network. The problems associated with assessing, commissioning and servicing transmitting and monitoring equipment on multiple channels increase in proportion to the number of channels. With this in mind, the IBA has developed a range of high-quality precision u.h.f. test transmitters and demodulating equipment for performance measurement of transmission and reception equipment within the UK Independent Television network.

A major requirement of the design was to eliminate the need to change costly and heavy fixed channel modules within the test equipment and make the equipment switch or remotely tunable to all channels in the u.h.f. band. A further requirement was that the demodulation equipment should use synchronous detection as first introduced by the IBA to the u.h.f. network for re-broadcast receivers in 1971 because of the superior waveform performance, and that the modulation equipment should be compatible with equipment using synchronous detection.

To achieve the tunable characteristics required, it was necessary to use a synthesized local oscillator in each of the equipments. The performance requirements of synthesizers when used with synchronous detectors is described in Sections 4 and 5 of this paper, and is also of considerable importance to the specification and performance measurement of both synthesizers and synchronous detectors when used elsewhere in a broadcasting network.

2 The Requirements

There are three main requirements for tunable test equipment within the IBA. These are:

(a) Service planning: the tunable receiver is required for assessment of the signal reception at potential transmitter or transposer sites where the off-air signal is required for re-transmission. The tunable test transmitter is used in conjunction with a 1 watt amplifier and a portable tower and aerial to help determine the service area of proposed low-power transposers. This is particularly useful because of the need deliberately to limit coverage to the required area and avoid interference in other areas.

(b) Installation and commissioning: the tunable demodulator and the tunable test transmitter are required for the testing of transmitters and transposers during acceptance, installation and commissioning.

(c) Maintenance: the tunable demodulator and tunable test transmitter are required for the

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maintenance of transmitters and transposers. The tunable receiver is a particularly useful aid in tracking down the origin of distortions etc. in long chains of transmitters and transposers where the cause could lie at any stage in the chain.

Table 1

Acceptance and maintenance limits

Parameter	Acceptance Limits	Maintenance Limits
VISION		
$2T$ pulse K rating	2% K	2% K
$2T$ pulse bar ratio	100% ± 4%	100% ± 4%
bar tilt (10 μs bar)	1%	1%
chrominance/luminance gain ratio	± 10%	± 7%
chrominance/luminance delay	± 20 ns	± 20 ns
chrominance/luminance cross-talk	—	2%
luminance linearity	7%	7%
differential gain (error)	6%	5%
differential phase (error)	3°	4°
intermodulation products (3-tone test)	−50 dB	−50 dB
incidental phase modulation	5°	10°
h.f. noise (r.m.s. unweighted)	−54 dB	−54 dB
l.f. noise (pk − pk unweighted)	−50 dB	−48 dB
SOUND		
noise (weighted—vision) modulated)	−66 dB	−58 dB

Acceptance, commissioning and maintenance limits for transmitters and transposers are shown in Table 1. This list must for reasons of space be limited to only the main parameters.

3 The Equipments

The tunable equipments, the receiver, demodulator and test transmitter are required to be capable of as many of these tests as possible and this indicates the facilities required. The performance limits must also be better than those of the equipment under test. The requirements for a tunable receiver and tunable demodulator were found to be sufficiently different that it was decided not to compromise the performance by producing a single equipment.

Because an off-air signal to be measured or monitored need not necessarily be the strongest signal at that site, the receiver must be capable of operating correctly in the presence of an unwanted group of channels of signal strength higher than that under investigation. Particular features required are therefore low noise figure, high dynamic range, high selectivity and good a.g.c. range.

The tunable demodulator is designed for measurements on transmitter output signals, and therefore very little gain or selectivity is required, but the very high linearity of the receiver must still be maintained. The demodulator is therefore a simplified version of the receiver, but with extra facilities including optional sound rejection to enable transmitter measurements to be made over a wider bandwidth, and facilities for the measurement of transmitter incidental phase modulation.

The tunable test transmitter is designed to generate a video modulated carrier of adjustable modulation depth from an external video feed. The sound carrier may be modulated either in frequency or amplitude by an internal or external audio source. The equipment also includes facilities for 2- and 3-tone tests with the carrier levels correctly calibrated without adjustment.

All three equipments are required to be rugged as they are for mobile applications and can therefore be expected to be carried by car to remote sites along unmade roads or tracks. Table 2 shows some of the specifications of the three equipments. Again, as for Table 1, the complete list would be far too lengthy to include here.

4 The Designs

The tunable equipments all employ the same design of frequency synthesizer and share a common choice of 1st and 2nd intermediate frequencies. The basic block diagrams are shown in Fig. 1.

The receiver, demodulator and test transmitter are all of double-superheterodyne design with intermediate frequencies of 1191·75 MHz V_c, 1185·75 MHz S_c, 22·5 MHz V_c and 16·5 MHz S_c. Several of the modules of the receiver and demodulator such as the sound trap, vestigial side-band filter, f.m. demodulator and sound amplifier are also common to the IBA fixed tuned equipments, and will therefore not be discussed here. The synchronous detector, although also common to fixed tuned receivers and demodulators, is discussed here because of the effect on phase noise from synthesizers.

The receiver, demodulator and test transmitter heads are shown in Fig. 2. The image reject and l.o. reject filters which are printed on the various amplifier boards are omitted to simplify the diagrams.

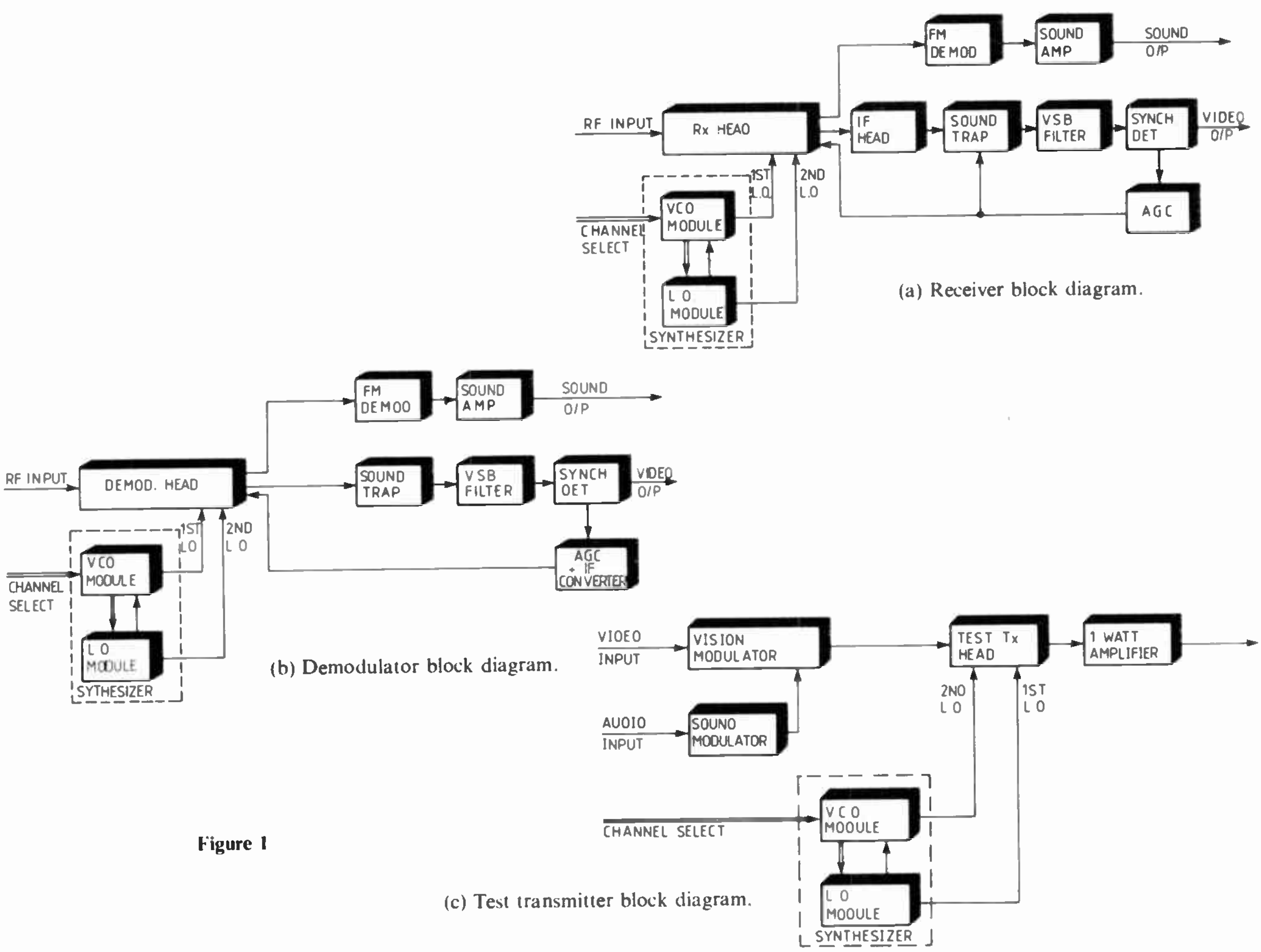
The three r.f. heads are very similar in design. The receiver and test transmitter use the same 5-pole bandpass filter for the 1191·25 MHz i.f. The demodulator, however, does not have the same filtering constraints as the receiver or test transmitter and therefore uses a wider filter bandwidth with consequential easing of alignment and manufacture. A detailed study has been made of the effects of intermodulation and unwanted mixer products. The most complicated case is that of the receiver because of the possibility of many signals being present at the input, with the wanted signal not necessarily the strongest. It was for this reason that the i.f. of 1191·25 MHz was chosen.

Adjacent channel interference (a.c.i.) is a particularly severe test of a high-performance receiver as intermodulation can occur in stages prior to the a.c.i. filter. It is not possible to design a stable a.c.i. filter at first i.f. The early

Table 2

Tunable Equipment Specification

Parameter	Receiver	Demodulator	Test Transmitter
input level	50 μV to 7 mV	100 mV + 3, − 10 dB 330 mV + 3, − 10 dB	—
output level	—	—	100 mV ± 1·5 dB
noise figure	8 dB	—	—
VISION PERFORMANCE			
2T pulse K rating	1% K	1% K	0·5% K
2T pulse/bar ratio	100% ± 1%	100% ± 1%	100% ± 1%
bar tilt (10 μs bar)	1%	1%	1%
chrominance/luminance gain ratio	2%	2%	1%
chrominance/luminance delay	± 10 ns	± 10 ns	± 5 ns
chrominance/luminance cross-talk	1%	1%	1%
luminance linearity	1%	1%	0·75%
differential gain (error)	1%	1%	1%
differential phase (error)	1°	1°	1°
intermodulation products	− 60 dB	− 60 dB	− 65 dB
incidental phase modulation	1°	1°	1°
h.f. noise (r.m.s. unweighted)	− 50 dB	− 53 dB	− 52 dB
l.f. noise (pk − pk unweighted)	− 56 dB	− 56 dB	− 54 dB
SOUND PERFORMANCE			
weighted signal/noise ratio (with vision modulated)	60 dB	60 dB	65 dB



(a) Receiver block diagram.

(b) Demodulator block diagram.

(c) Test transmitter block diagram.

Figure 1

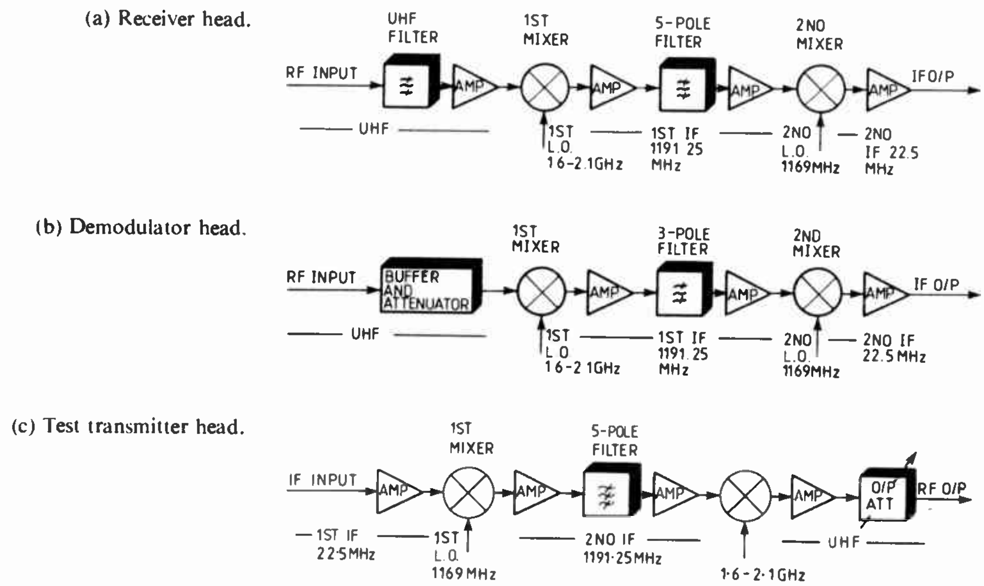


Figure 2

stages of the receiver to second i.f. prior to the a.c.i. filter are therefore designed with high dynamic range, and the a.g.c. in the earlier stages is limited for this reason.

Figure 3 shows the block diagram of the synthesizer which is common to all three equipments.

The synthesizer frequencies have been chosen to ensure that spurious signals generated in the mixers fall outside the passbands of the i.f. and r.f. filters of the three equipments. The phase-locked loop has been optimized to give minimum phase noise on the output signal. In particular, the unusually high reference frequency of 1 MHz was chosen to make possible a high loop bandwidth so that phase noise sidebands out to ~50 kHz from carrier for the voltage-controlled oscillator are significantly reduced in amplitude. For all channels, the final peak-to-peak phase noise at the synchronous detector output has been reduced to ~1°. This ensures that the video signal/noise requirements of the three equipments can be met, even for transmissions with high levels of incidental phase modulation. Phase noise is discussed in more detail later. Acquisition of the synthesizer during channel changes take place in less than ~100 ms, due to the presence of a powerful frequency lock circuit.

The tunable receiver and demodulator both use synchronous detection for demodulation. This technique has been in use in IBA fixed tuned receivers since 1971 to avoid the quadrature distortion produced as a result of envelope detection of vestigial sideband transmissions.

Figure 4 shows the block diagram of the synchronous detector used for both the receiver and demodulator. The detector phase locked loop has a bandwidth of around 750 Hz. A gated loop is used with the gating at the sync pulse tips, and this sets an upper limit to the stable loop bandwidth available. The use of a gated loop is essential if the detector phase is to get a correct phase reference, essential for both re-broadcast video and for measurement of incidental phase modulation. Non-gated loops are sensitive to modulation and incidental phase modulation and tend therefore to have a varying phase reference.

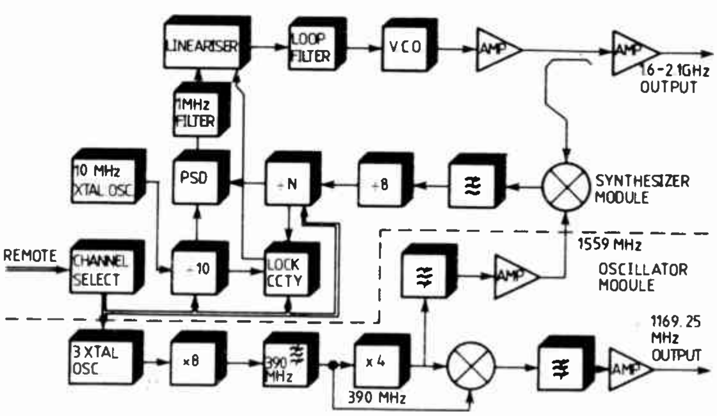


Fig. 3. Synthesizer block diagram.

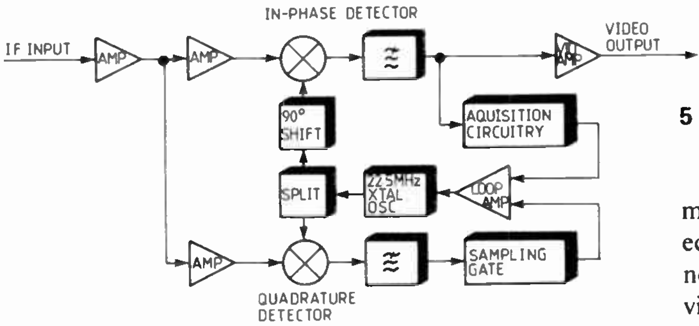


Fig. 4. Synchronous detector block diagram.

Figures 5(a) and (b) show block diagrams of the vision and sound modulators of the tunable test transmitter.

The vision modulator accepts a 1 volt signal and via a peak sync clamp, modulates the 22·5 MHz carrier. Filtering is employed to reduce harmonics on the i.f. output to a very low level and group delay equalization is employed to reduce in-band errors to <3 ns. A third tone oscillator is included for 3-tone tests and can be variable in frequency to test for frequency dependence of intermodulation products, and also provides a convenient method of conducting frequency sweeps within the video passband.

The sound modulator generates the 16·5 MHz sound carrier at a spacing of 384 × line rate (6 MHz exactly), or 5·9996 MHz from vision carrier. A narrow-band phase-locked loop is used to provide accurate frequency stabilization, and can be modulated from an audio input with either a.m. or f.m., the latter with switchable 50 µs pre-emphasis. The a.m. facility is useful for tests of a.m. rejection of discriminators in receiving and monitoring equipment.

5 Phase Noise in Synthesizers, Synchronous Detectors and V.S.B. Filters

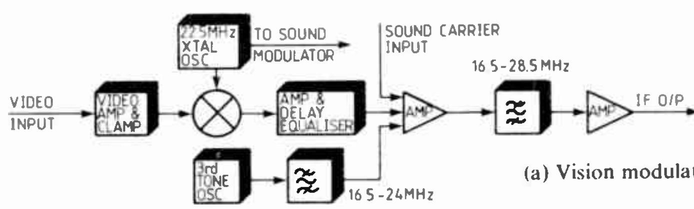
Phase noise is not generally a directly specified or measured parameter of transmission or measurement equipment. If significant, it manifests itself in the form of noise on sound or as unpredictable levels of noise on vision waveforms, especially on chrominance as a result of synchronous detection or as a result of slope detection through the v.s.b. filter and detector.

Fixed tuned equipment generally employs crystal oscillators as local oscillator sources. These generally have excellent phase noise performance so that there is no problem for synchronous detection. This is the situation in the existing IBA network where all transmitters use crystal oscillators and multipliers as frequency sources and all permanent demodulators and receivers are fixed tuned.

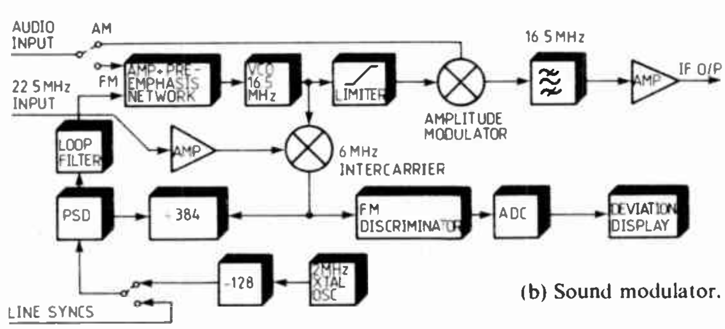
Synthesizers have considerably higher levels of phase noise in general. Synchronous detectors contain some form of limited bandwidth recovery (a phase locked loop in IBA detectors) and are therefore phase sensitive.

Consider the phase locked loop shown in Fig. 6. θ_ε is the phase error from in phase in the synchronous detector. The effect of this can be seen in Fig. 7(a).

If θ_ε has a total deviation ϕ_ε peak to peak, then the voltage peak to peak at the output of the detector is $V\phi_\varepsilon^2/8$ where V is the voltage of the particular part of the waveform of interest with respect to the blanking (zero-carrier) level. For example a video signal-to-noise ratio at black level (measured pk − pk with respect to 0·7 V) of −52 dB is produced by phase noise of 7° pk − pk.



(a) Vision modulator.



(b) Sound modulator.

Figure 5

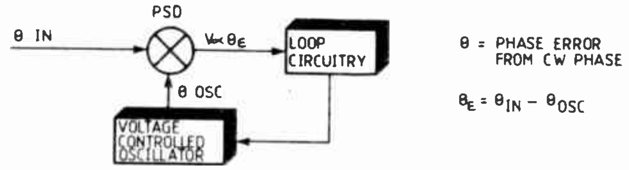
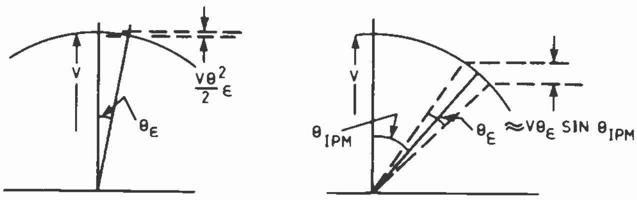


Fig. 6. Phase locked loop.



(a) Effect of phase deviation. (b) Effect of incidental phase modulation.

Figure 7

Incidental phase modulation can cause a significant difference between the reference phase and the phase at the appropriate point in the waveform. The detector situation can then be as shown in Fig. 7(b). In this case, if $\theta_{IPM} \gg \phi_\varepsilon$, the phase noise pk − pk, then the phase noise is given by $V\phi_\varepsilon \cdot \sin\theta_{IPM}$, so that the signal-to-noise ratio is closely related to the amount of incidental phase modulation. For example, for i.p.m. of 10°, then the same noise level of −52 dB at black level is produced by phase noise of only 0·7° peak to peak.

For the synthesizer/detector combination used in the tunable equipments described here, care has been taken to minimize the phase noise. In particular, it was necessary to balance the phase noise tracking which results from high synchronous detector loop bandwidth with the aliasing effect of the sampling gate in the detector loop which results in a noise component which increases with increased bandwidth. The optimum has been found to be around 500 Hz–750 Hz loop bandwidth. The total phase noise for each tunable equipment has been reduced to typically ~1° pk − pk of which ~0·5° pk − pk is low frequency noise. This enables us to meet the l.f. and h.f. noise performance specifications laid down in Table 2 even when the transmitter or other device under test has incidental phase modulation ~10°.

The theory above shows that the introduction of synthesizers into transmitters and transposers in conjunction with synchronous detectors in reception and measuring equipment can lead to severe problems if phase noise calculations are not considered, and in particular the effect of other transmitter defects such as i.p.m. are not considered.

Vestigial sideband filtering is a source of h.f. noise for both envelope and synchronous detection. This is because the v.s.b. slope and detector act as an f.m. discriminator and are therefore sensitive to wideband f.m. noise. This noise is not a limitation in the IBA tunable equipments, but can again be important where synthesizers are used in transmitters and transposers.

6 Measurement Features

The tunable equipments provide the means of carrying out virtually all the acceptance and maintenance tests on transmitters and transposers when used in conjunction with video and audio generation and measurement equipment, and, for two and three tone tests, in conjunction with a spectrum analyser. Table 1 lists many of the parameters which are measured using the tunable equipments. In particular, the following facilities for the three equipments should be noted.

(a) Tunable test transmitter:
- (i) Clamped vision modulator with calibrated or variable modulation depth.
- (ii) Full sound f.m. modulator with or without pre-emphasis. a.m. facilities are also available for tests on limiters and discriminators. A display of modulation depth is given.
- (iii) Switchable 2- or 3-tone tests with calibrated setting of correct carrier levels and fixed or variable frequency third tone.
- (iv) Precision output levels and frequencies on all channels from 21 to 70 at −5/3, zero and +5/3 line rate offsets. The channel selection may be made remotely or on the front panel.
- (v) In conjunction with the 1 W amplifier, the tunable test transmitter can act as a complete 1 W transmitter for selection of suitable low power transmitter sites.

(b) Tunable receiver:
- (i) Good noise figure (typically better than 8 dB) and high input sensitivity (signal range 50 μV to 7 mV in 50 r.m.s. peak sync vision carrier).
- (ii) High selectivity—a switchable filter with >30 dB a.c.i. rejection is fitted.

(c) Tunable demodulator:
- (i) High dynamic range input to give good signal to noise ratio. A limited range a.g.c. is incorporated so that levels of +3 dB to −10 dB of the nominal 100 mV or 330 mV r.m.s. pk sync input are correctly demodulated to 1 V video.

Other features common to both the receiver and demodulator are:

- (i) Input on all channels from 21 to 70 at −5/3, zero or +5/3 line rate offsets at frequencies up to ±5 kHz of nominal.
- (ii) F.m. sound can be demodulated directly or in the 'inter-carrier' mode; 50 μsec de-emphasis can be switched in or out.
- (iii) Synchronous or envelope detection—switch selectable.

The importance of synchronous detection can be seen from the list of distortions in Table 3 which are associated directly with envelope detection. Hence

Table 3

Distortions associated with 'perfect' envelope detection—System I

	Synchronous	Envelope
pulse to bar ratio	100%	~95%
chrominance/luminance cross-talk	0%	~8%
chrominance/luminance gain inequality	0%	~2%
differential gain (140 mV subcarrier)	0%	~4%
line-time non-linearity	0%	~4%

measurements made with a synchronous detector do not have to be 'corrected' for these distortions, and therefore can be made with considerably greater accuracy.

Synchronous quadrature video can also be selected for incidental phase modulation (i.p.m.) measurements. I.p.m. on the transmitted vision carrier causes 'intercarrier buzz' when the sound is demodulated in the normal intercarrier f.m. demodulator of a domestic television set. Although synchronous detection is used to avoid the distortions inherent in envelope detection, i.p.m. can itself be a source of waveform distortions when used with synchronous detection. I.p.m. enhances the noise from oscillators in transmitters and transposers when demodulated with synchronous detectors as part of a programme chain. As shown above, i.p.m. for this must be limited to $<10^\circ$.

At present, phase noise is not a directly measured parameter. It is assumed to be satisfactory if the l.f. and h.f. signal-to-noise ratios are within specification. It may be that with increasing use of synthesized local oscillators in transmission equipment, this will need to become a specified and measured parameter in its own right, as indeed it already is with the synthesizer used in the tunable equipments.

The provision of full clamping circuitry in the tunable test transmitter makes it possible to use the equipment for the testing of transposers and fixed tuned receivers and demodulators for teletext performance, and because full sound modulation circuitry is included, it is possible to check for any untoward effects occurring on the sound channel.

7 The Future

The existing test equipment used by installation and maintenance teams is still heavy and bulky, although the tunable equipments have reduced significantly the amount of equipment compared with that which would otherwise have been carried. In the future, we intend to reduce the number of equipments carried by moving away from general-purpose test equipment such as spectrum analysers towards test equipment designed specifically for installation or maintenance.

Specific maintenance equipment is under investigation by the IBA at present. A suggested configuration is shown in Fig. 8. The unit being studied at present is the central unit containing all the r.f. processing for all the maintenance tests normally carried out. Microprocessor control will be used to remove the time-consuming setting of controls and calibration of the test equipment. In this way, the total amount of equipment will be reduced and maintenance will be eased, especially for sites with difficult access.

8 Conclusions

The tunable equipments, the test transmitter, receiver and demodulator provide the means of carrying out much of the transmitter installation and maintenance work of the IBA broadcast network. These equipments use low noise frequency synthesizers, and synchronous detectors to provide the high standard of performance required.

In this paper, the significance of phase noise of synthesizers where synchronous detection is employed later either in the equipment or in the transmission chain, has been discussed. In particular, the effects of incidental phase modulation have been demonstrated. From this, it can be seen that the phase noise performance of synthesizers used in transmission equipment must be good if noise on demodulated video is not to be a problem. The derivation of a specification is, however, not a simple exercise because of its close dependence on the type and performance of synchronous detector used.

For the future, the tunable equipments will be improved, condensed and automated for a comprehensive light weight test set for operational use.

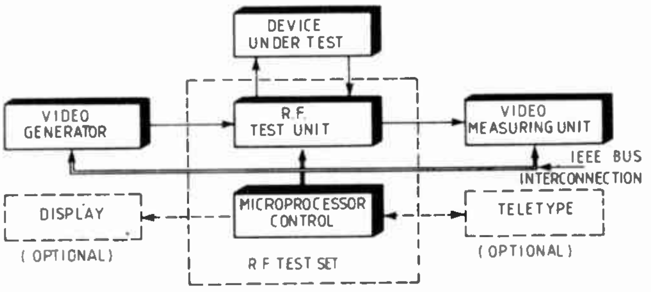


Fig. 8. Comprehensive transmitter/transposer maintenance equipment.

9 Acknowledgments

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The EURONET Telecommunications and Information Network

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Based on a paper presented at a meeting of the Communications Group in London on 23rd November 1978.

SUMMARY

In 1971 the Council of the European Economic Community agreed in principle to the planning of a computer network to enable data terminals located in any of the member countries to gain on-line access to scientific, technical and socio-economic information held on data bases at specific locations. In 1974 the Commission issued a three-year action plan which included the establishment of a computer network which became known as EURONET. More recently the term EURONET has become associated with the Telecommunications Network alone and, for aspects concerning the terminals and the data bases, the name DIANE—Direct Information Access Network EUROPE—has been adopted. The EEC EURONET organization is described in the paper and also the involvement of the CEPT (Conference of European Postal and Telecommunications Administrations) and of the individual Administrations.

The basic specification, and the network configuration to meet it, are given together with those factors affecting the choice of packet-switching.

Following a description of the hardware and software to be used, including alignment with CCITT (International Telegraph and Telephone Consultative Committee) Recommendations, an assessment is made of the likely future enhancement of EURONET, after its planned opening in 1979, and of the work undertaken by the Commission to standardize data base access and retrieval procedures.

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1 Introduction

In the late sixties the Council of the European Economic Community (EEC) foresaw the need for the exchange of a large volume of information between data terminals and data bases, located on host computers, in the member countries of the Community. The data bases cover a wide range of information on medicine, chemistry, engineering, textiles, electronics, physics, metallurgy, aerospace, nucleonics and patents. The EEC agreed in principle, in 1971, to plan a computer network giving on-line access to data bases within the Community and, in 1974, issued a three-year Action Plan which envisaged a computer network consisting of a telecommunications network called EURONET serving data bases now known collectively as DIANE (Direct Information Access Network—Europe). EURONET is currently planned to come into full commercial operation in the autumn of 1979 when, in its initial phase, it will connect to some 20 host computers containing over 100 data bases. Significant hosts for early connection to the network are the Space Documentation System of the European Space Agency at Frascati (Italy), the German Medical Documentation and Information System at Cologne, the British Library Information System (BLAISE) in London, Infoline, and the Commission's data bases in Luxemburg. Access to DIANE by terminals located in non-EEC countries is envisaged in due course, as also is access by EEC terminals to data bases in non-EEC countries.

2 The EURONET Organization

The Directorate-Generale XIII of the Commission, in Luxemburg, through its Committee for Information and Documentation on Science and Technology (CIDST), has established an influential sub-committee, the Technical, Economic and Financial Group (ETAG) to oversee the project. In early 1975 the Commission made an initial approach to the nine Community-based Telecommunications Authorities (PTTs as they are known) and in December 1975 these nine PTTs formed a legal Consortium and empowered the French PTT, on their behalf, to sign a contract with the Commission to implement the telecommunications network—EURONET. At about the same time of the approach made to PTTs, the Conference of Postal and Telecommunications Administrations (CEPT), on which all the European PTTs are represented, established a new organization to co-ordinate planning and policy in the data transmission field. Its Special Committee for Data Transmission (CSTD) quickly realized the significance of EURONET and agreed to co-operate fully with the Commission to establish the network as soon as possible.

Within the Consortium a management committee has oversight of EURONET with two sub-committees, one dealing with planning and implementation (realization) and the other dealing with commercial aspects. A project

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team controls the detailed implementation aspects of the network, liaising with individual national implementation groups. Individual national groups also exist to cover marketing aspects. The cost of providing the overall telecommunications network has been co-financed by the Commission and the PTTs jointly on approximately a 50 : 50 basis but with the Commission covering all contractual costs for the international element.

3 The Network

The contract signed between the Commission and the PTT Consortium stipulated certain objectives which were of considerable significance and influenced the design of the network.

(a) The network should be implemented in accordance with agreed CCITT/CEPT standards[1,2] in so far as they apply.
(b) The design of the network should be such that its reliability can be enhanced to that needed for a public network.
(c) Access to the network should be via national public data networks wherever such networks exist or as soon as they exist.
(d) The design of the network should not preclude the carrying of non-EURONET traffic.
(e) In the longer term the services and facilities provided by EURONET should be gradually integrated into an emerging European public data network.

The network configuration currently envisaged for the initial phase—that is in 1979—is shown in Fig. 1. Four packet-switched exchanges (PSE) have been established in Frankfurt, London, Paris and Rome with remote access facilities in Amsterdam, Brussels, Copenhagen, Dublin and Luxemburg. These remote access facilities will enable not only low-speed data terminals connected to the public switched telephone networks in each country to gain access via suitable multiplexed links to the packet switched exchanges but also hosts to be connected. A central Network Management Centre (MNC) is located in London. The design of the network is such as to ensure that the facilities available to users connected via remote multiplexors are identical to those for users who are connected directly to the PSEs. For example, the distance of a user from a PSE or multiplexor site should not prejudice the facilities or performance offered.

4 Packet Switching

The term packet switching describes a system where users' data are formed into one or more blocks (packets) of a maximum size, 128 bytes having now been agreed for international working, and the packet is switched as a complete entity at appropriate points in the network.

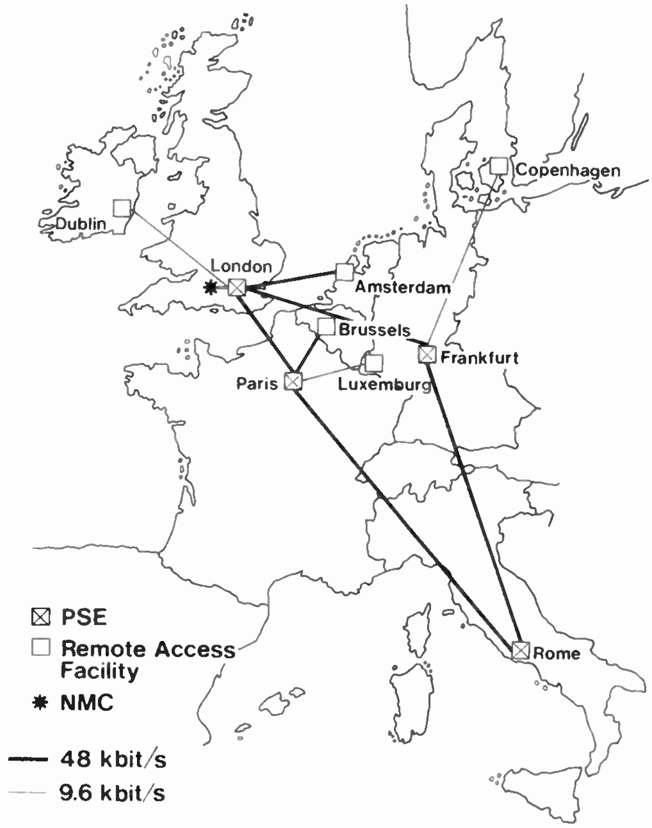

Fig. 1. EURONET network at opening date.

Each packet carries address and control information contained in a header and is followed by an error-check field. Following international Recommendations, each packet is put inside a High Level Data Link Control (HDLC) frame with the check bytes added on a frame rather than a packet basis as shown in Fig. 2. Each packet is transmitted individually and, on any given part of the network it traverses, time-shares with other packet based traffic as shown in a simplified form in Fig. 3. Arrangements are made for the correct sequential delivery of all packets associated with a particular call. Thus, elements of the network are time-shared by a number of users each of whom appears to have exclusive use of the route. Details of the basic concepts and techniques of packet switching can be found in a series of articles, published in the Journal of the Institution of Post Office Electrical Engineers, relating to the Post Office Experimental Packet Switched Service (EPSS).[3,4]

Prior to the choice of a method of data transmission for EURONET, studies had shown that a common telecommunications network covering all the Community countries would cost the European taxpayer between one-third and one-tenth of what it would cost if separate networks were set up alongside each other, e.g. one star network for each host computer. The adoption of the packet mode of operation stemmed from a need to accommodate a wide range of often incompatible terminals on the single network and

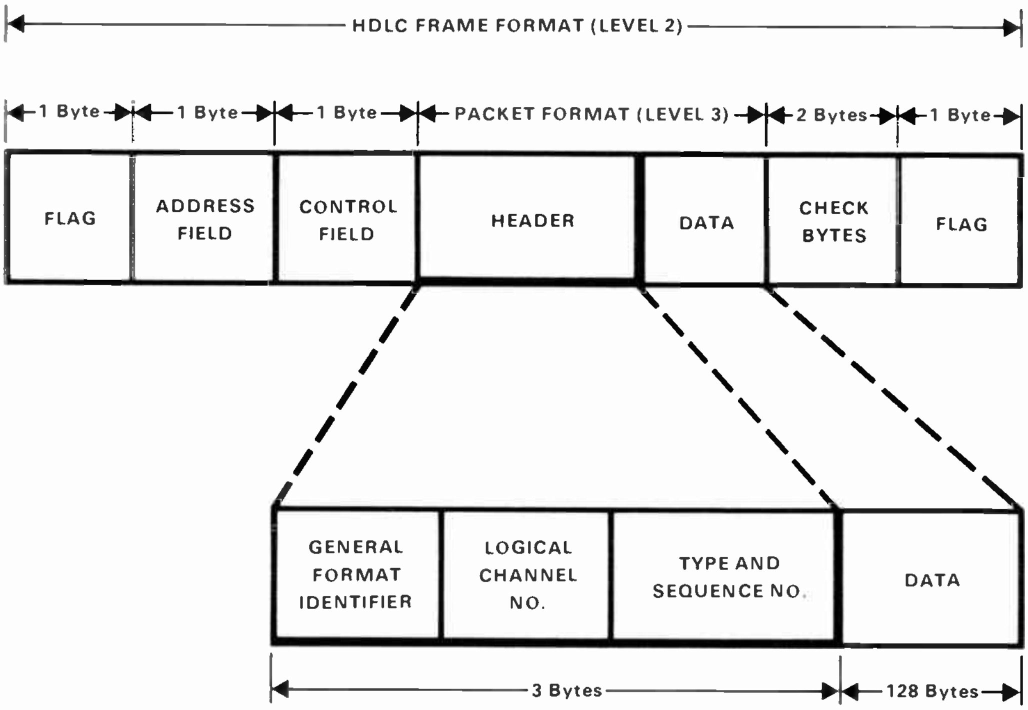

Fig. 2. Frame and packet structure.

(perhaps the more significant point) from a need to provide low cost access in an interactive mode, to the data bases connected to the network.

The conventional international telephone network could, of course, have been used but it was not considered appropriate for interactive searches because of the following shortcomings:

Restriction of working speeds.
Long connect time.
Restricted facilities.
Expensive long-distance transmission.
Different traffic pattern of data transmission from telephony traffic.
Relatively poor error performance.

The packet mode of working lends itself naturally to volume based charges which can, within certain limits, be distance independent and thus meet the Commission's objective that all users should be treated on an equal footing throughout the Community and that no one should be penalized tariffwise because of location.

5 EURONET Packet Switching

A user of the EURONET packet-switched network will be able to gain access to the hosts containing the data bases in either of two ways. Either the network will establish, at a user's request, a communication path known as a virtual call between a user's terminal and any one of the hosts or a user may use a specially allocated route, called a permanent virtual circuit, through the network to a specific host. The term virtual call is used because, although the call appears to provide a direct connection between a user's terminal and a computer, transmission facilities within the packet-switching network are only assigned to the connection when packets are being transferred over it. A permanent virtual circuit provides such a connection on a permanent basis; it resembles a point-to-point leased or private circuit in that it allows data transfer only between two predetermined terminals but, as for virtual calls, transmission capacity is allocated only when packets are being transmitted. A permanent virtual circuit, once established on EURONET, has a permanently allocated route through the network, whereas a virtual call has a route allocated between two terminals only for the duration of that call.

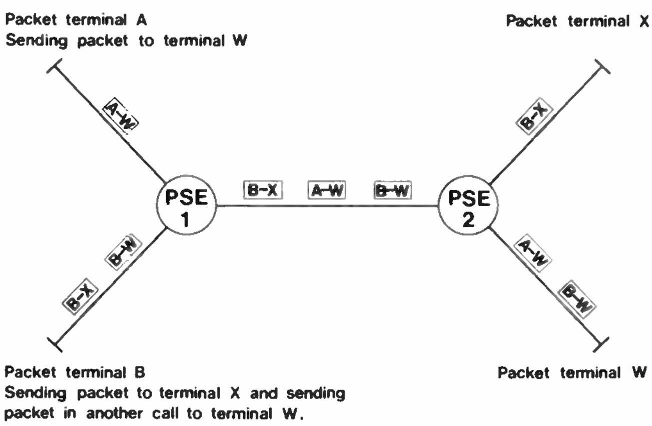

Fig. 3. Principles of packet switching.

When a virtual call is made, the optimum routing of the packets through the EURONET packet-switching network is decided by the PSEs in accordance with routing tables, for a virtual call can be established over a choice of inter-PSE routes. Once the route is established it is used for the duration of the call and, on completion, it is cleared down. The routing tables are held in each PSE and can be changed remotely via the NMC. The system of fixed routing adopted for EURONET has certain advantages over the adaptive routing method used by many other packet-switching networks, including the Post Office Experimental Packet Switched Service (EPSS). With the adaptive routing method, the PSEs choose the optimum route for each packet during a virtual call, thereby adapting the routing of the packets to a continually changing traffic situation. In an international network involving routings through several countries the transit accounting arrangements of PTTs for transit virtual calls, each packet of which could take a different route, would present excessive network management overheads if adaptive routing of packets were to be adopted. EURONET, however, uses the fixed routing method which minimizes the network management overheads in this respect. With the fixed routing method the individual PTTs preferences for certain routes, i.e. direct or via a third country, can be taken into account when the routing tables are established or changed.

Over a physical link between a EURONET packet-mode data terminal and its serving PSE a number of virtual calls can be established simultaneously. To enable the terminal and the PSE to distinguish between packets associated with different virtual calls, a logical channel number is assigned to each virtual call when it is established; during the call all packets carry this logical channel number. A permanent virtual circuit is also assigned a logical channel number although this remains fixed as long as the permanent virtual circuit is established. Over a single physical link, several permanent virtual circuits and virtual calls can be continued simultaneously if required. In fact one 9600 bit/s packet port can handle several hundred remote 110 bit/s terminals at a time.

One of the advantages of the packet-mode of working is the network's ability to interconnect terminals having different data signalling rates. This requires that the network be able to control the flow of packets between the two terminals. The network acts as a temporary absorber (or buffer) of data. Depending on the degree of buffering available, and the switching capability of the network, a limit on the data throughput transmitted by a terminal has to be imposed and the maximum value allocated to a terminal is referred to as the 'throughput class'.

The network is able to provide an error-checking facility on a link-by-link basis. This detects transmission errors on inter-PSE trunks and on links between PSEs and packet terminals. If a frame (which contains a packet) is corrupted during transmission over a link the receiving PSE or packet-mode terminal requests the sending packet terminal, or PSE, at the other end of that link to re-transmit the frame. Once a frame has been satisfactorily received it is no longer retained at the sending end of the link. A frame can consist of an opening synchronization character (known as a flag), a frame address field, a frame control field, a packet, a frame check sequence field and a final flag. A frame containing an information field, which could for example be a packet containing user's data, is shown in Fig. 2.

6 Protocols

One of the most important aspects of EURONET is the establishment of appropriate protocols or procedures to enable calls to be established between the data terminals and the host computers and for subsequent transmission of information after the completion of the call establishment phase. To establish communication between two terminals it is necessary to set up a virtual call or use a permanent virtual circuit. This necessitates the adoption of specified terminal-to-network interface procedures, or protocols as they are known by users. Different protocols are used depending on whether the terminal operates in the packet or character mode. Figure 4 shows the logical network topology, the major interfaces and also indicates the appropriate CCITT X-series Protocol Recommendations. Also shown are two Application Protocols 'H' and 'G', which are network independent. These are known as High Level Protocols, and those for EURONET, but really for DIANE, will be developed and initially standardized by the Commission on behalf of EURONET users and may eventually become full standards of the International Standards Organisation (ISO).

6.1 Recommendation X25

The protocols applicable to a EURONET user-to-network packet interface are those given in CCITT Recommendation X25 which deals with the interface between data terminal equipment and data-circuit terminating equipment for terminals operating in the packet mode on public data networks. Figure 5 shows the protocols required for data transmission between packet mode terminals, and it includes the three levels specified by CCITT Recommendation X25. They are:

Level 1—the physical, electrical, functional and procedural characteristics to establish, maintain and disconnect the physical link between the terminal and the network;

Level 2—the link access procedure (LAP) based on HDLC procedures for data interchange across the

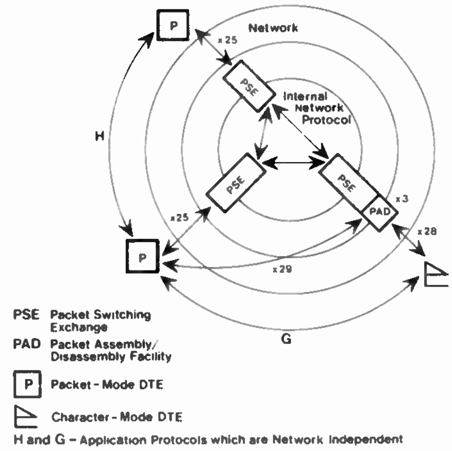

Fig. 4. Logical network topology.

link between the terminal and the network (recently the CCITT has proposed an alternative procedure known as LAP 'B' and it is expected that this will also be offered in due course on EURONET); and

Level 3—the packet format and control procedures for the exchange of packets containing control information and user data between the terminal and the network.

Figure 5 also shows the network independent protocols applicable to the user's system and these are shown connected to the network-dependent protocols by means of bridging software.

6.2 Recommendation X3

To enable a start-stop character-mode terminal to use a packet switched system, and to communicate with a packet-mode terminal, provision must be made to convert the output/input of these terminals to/from a packet format. This function is performed by a packet-assembly/disassembly (PAD) unit, and is defined in CCITT Recommendation X3. A PAD is an integral part of the network and is implemented in each PSE.

6.3 Recommendations X28 and X29

CCITT Recommendations X28 and X29 define, respectively, the interface protocols for a start-stop character-mode terminal working to a PAD and for a packet-mode terminal working to a PAD, as illustrated in Fig. 4. The X28 and X29 Recommendations have been further extended to meet EURONET needs for extra facilities and data signalling rates.

6.4 Recommendation X75

CCITT Recommendation X75 is for an international inter-network signalling protocol. Although the inter-PSE protocol adopted for EURONET will initially be that used for TRANSPAC (the French national packet switched data network),[5] it is expected that in due course this protocol could be modified to the proposed international inter-network signalling protocol, CCITT Recommendation X75, which will enable the connection

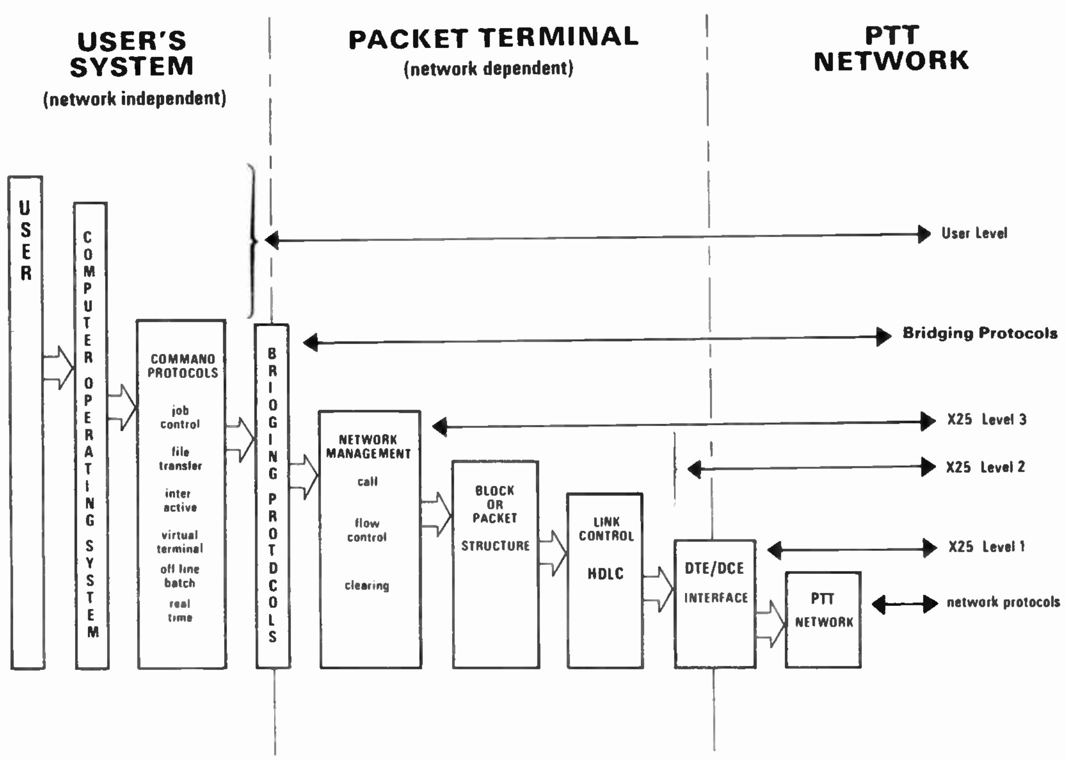

Fig. 5. Protocols for packet-mode terminals communicating via the network.

of EURONET not only to other international public packet-switching data networks but also to national networks such as the new Packet Switched Data Service (PSS)[6] in the United Kingdom, planned to be operational in early 1980.

7 Traffic and Tariffs

Recent studies carried out at the request of the Commission indicate that the ratio of searches within a given Community country and to other countries could be of the order shown in Table 1. Naturally, as might be expected in the case of the larger countries, many of the searches will be national in nature and will not need to flow via the international telecommunications network. Taking the United Kingdom as an example, some 65% of all searches could well remain within the United Kingdom with the majority of external searches being made of the data bases in the Federal Republic of Germany. On the other hand it is probable that the majority of incoming searches to the United Kingdom will be coming from France.

One factor of considerable interest to the PTTs concerned with designing and planning the telecommunications network is the traffic flow. This flow can be affected by the number of hosts on the network and the extent to which the data bases may be duplicated on several hosts. Whereas two years ago only two hosts had been named for connection to the network initially, EURONET has now attracted over 20 hosts and 100

Table 1

Estimated ratios of country to country searches in 1979

	Demand									
Information Service Suppliers	BELGIUM	DENMARK	FRANCE	GERMANY	IRELAND	ITALY	LUXEMBURG	NETHERLANDS	UNITED KINGDOM	Supply Ratios
BELGIUM	21	2	24	0	1	11	0	16	23	98
DENMARK	66	105	214	27	5	125	1	128	163	834
FRANCE	70	20	1273	54	4	136	6	144	163	1870
GERMANY	123	37	410	2536	9	250	1	240	324	3931
ITALY	49	16	170	27	4	455	1	104	140	966
UNITED KINGDOM	82	25	286	27	45	159	1	168	1508	2301
Demand Ratios	411	205	2377	2671	68	1139	11	800	2321	10000

6.5 Network User Identification

In addition to the protocols given in CCITT Recommendations X3, X28 and X29, one of the other protocols required for EURONET is for a facility known as Network User Identification (NUI). This relates to the charging procedures adopted for virtual calls established when character-mode terminals gain access via the Public Switched Telephone Network (PSTN). Most users desire an option whereby they can elect either to pay for each virtual call themselves or to have all charges billed to the called terminal, say a computer bureau. For calls established through the PSTN to the EURONET packet-switching network, it is thus essential to be able to identify the calling user. On EURONET, this is achieved by the user accessing his allocated PAD and inserting an NUI code. This is checked by the PSE against a table of NUIs, and a network-user address is passed to the NMC for charging purposes. For a subsequent virtual call the user can opt to pay for all the packets or request all charges to be made to the distant terminal, if the distant terminal is prepared to accept the charge. This option will apply for use of the packet-switching network; the user will be charged for access using the PSTN or Public Data Network (PDN), according to national practice. Recently, however, CEPT Administrations have agreed not to offer the reverse charging facilities on international calls via EURONET because of inter-accounting (fund transfer) problems, but the reverse charging facility will be offered on national networks such as PSS.

Table 2

Provisional initial distribution of EURONET host computers and data bases

Country	Host Computers	Data Bases
BELGIUM	1	2
FRANCE	7	25
GERMANY	5	44
ITALY	4	25+
LUXEMBURG	1	3+
UNITED KINGDOM	2	14+
Totals	20	113+

Table 3

The UK proportion of terminal interfaces to be provided at the opening date

	Character Mode			
	Asynchronous		Synchronous	Packet Mode
	PSTN	Direct	Direct	Direct
UNITED KINGDOM	49(13)	9(7)	4(3)	5(1)
COMMUNITY TOTAL	205(91)	67(40)	17(13)	37(5)
UK percentage	24	13	24	14

Note: The quantities include those terminals, enclosed in parentheses, to be initially unequipped.

data bases as shown in Table 2, the main result of which has been to reduce the amount of international traffic which adds emphasis to the PTT's wish to see the network enhanced so as to carry non-EURONET traffic.

The number of terminal interfaces (ports) and their distribution throughout the network, are shown in Table 3 although, as also indicated, a number of ports are not required to be equipped initially. The ports have been allocated on a per-country basis to meet the anticipated traffic demands. Whereas host computers will be directly connected to the four packet switching exchanges over direct lines at 2400, 4800 or 9600 bit/s, in some cases utilizing multi-line procedures, i.e. utilizing more than one physical circuit, low-speed terminals will generally access the network via the PSTN in each country at speeds up to 1200 bit/s with the majority expected to be at 300 bit/s. In due course it is expected that all terminals will gain access via their respective national data networks which will, in turn, interface to EURONET. However some countries envisage implementing circuit-switched data networks and others packet-switched data networks initially, although nearly all European countries now intend in due course to provide a national packet-switched data service or at least a circuit-switched access to a packet-switched service.

The EURONET tariffs were announced in 1978 and have attracted considerable interest. They are based on the following structure:

(a) distance independent within the EEC;
(b) a time element for the duration of each virtual call, dependent on the user class, i.e. up to 1200 bit/s, 2400–9600 bit/s and at 48 k bit/s;
(c) for access via the PSTN, a virtual call duration charge in addition to the normal PSTN access charge;
(d) a volume charge based on the segment, rather than the packet, where a segment is a fixed number of bytes (the segment size had been set initially as 64 bytes in length, that is about half a packet);
(e) for a permanent virtual circuit, i.e. the equivalent of a leased line where data are transmitted in packet format, there will be a duration charge, once again depending on user class and based on an anticipated hourly usage/month plus a volume-based charge as for virtual calls;
(f) considerable reduction in virtual-call duration charges and volume-based charges for night and possibly weekend traffic;
(g) national access charges for direct connection and for dial-up connection appropriate to the originating country.

Since the tariff proposals have been agreed by all the Community based PTTs it is obvious that they will set the pattern for the structure for future international public packet switched data services. There could well be, however, some minor variations for the charges to apply to national networks. There is no doubt that the announcement of the actual tariffs has helped to identify a significant demand for international data services via EURONET rather than via the existing international PSTN, or via leased lines, but mainly for the low-volume user. The tariff being volume-dependent is unlikely to be attractive for bulk data transfer unless volume discounts are given by the PTTs and this is under active consideration. In addition to the charges for the use of the telecommunications network, charges will be levied by the data base operators for searches and the return of information but the charges have not yet been announced.

The EURONET telecommunications network is based on an adaptation of TRANSPAC. It differs from TRANSPAC in only two areas, in that the EURONET telecommunications network includes facilities for NUI to enable users accessing via the PSTN in each country to be billed for the calls they make (this is necessary because European telephone systems do not provide Automatic Number Identification (ANI)), and a facility for international accounting to enable the Community-based PTTs to apportion costs between themselves. This latter facility is essential if, as expected, non-EURONET traffic is carried on the network.

8 Network Hardware

Many of the earlier types of packet-switching network used conventional minicomputers, programmed to perform packet switching: EURONET, however, uses purpose-designed hardware to perform the packet switching function. Each PSE in the network will be based on two types of processor that have complementary functions. One, the CP50 processor, shown in Fig. 6, has been specially developed in France to perform the many repetitive functions involved in packet switching and is based on a number of specialized microprocessors. The other type of processor, called the Command Unit (CU), is a Mitra 125, a general-purpose

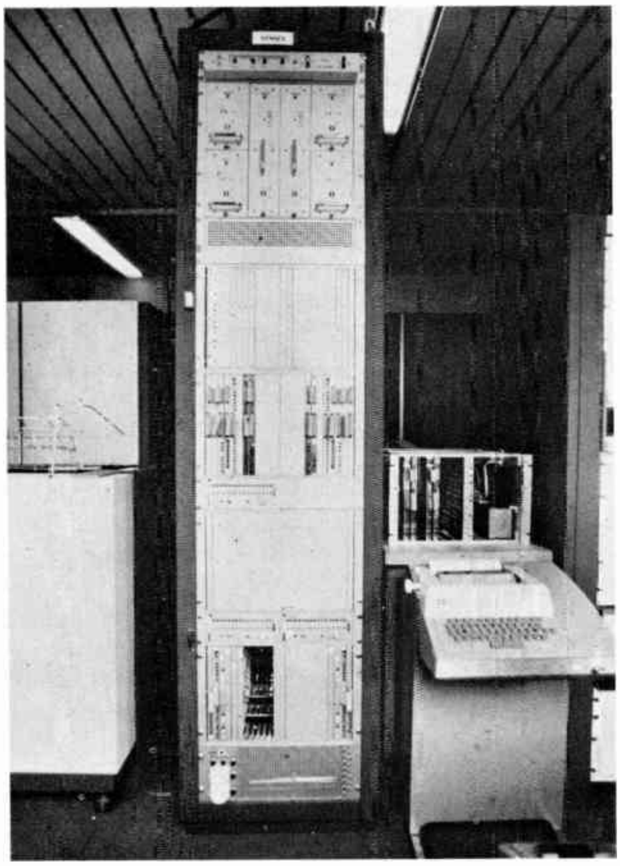

Fig. 6. CP50 switching unit. (Courtesy of SESA; photograph by B. Mandin.)

minicomputer once again of French manufacture. The CU controls the CP50 processor and is connected to it by a high-speed time-division bus: these two processors form the basis of a packet-switching exchange.

The performance of all PSEs in the network will be monitored, and overall control of the network, together with the international accounting and charging information, will be managed by an NMC. An additional Mitra 125 processor will form the basis of the NMC equipment. As already mentioned, at the opening date of EURONET there will be a single NMC, located in London, which will control four PSEs, three of them remotely.

In the interest of reliability, and with the exception of a few sub-units in the CP50 switch modules and the NMC, all the processor equipment will be duplicated. In each PSE, the second CU and the duplicated parts of the CP50 processor will operate in a 'hot stand-by' mode, ready to take over in case of failure of the operational equipment.

One of the features of the packet-switching equipment to be used for EURONET is its modular nature, enabling the expansion of a PSE in a number of steps by the addition of various modules. The CP50 processor is composed of one or more self-contained CP50 switch modules, each switch module occupying one equipment rack. Initially, the EURONET PSEs will each have only one duplicated CP50 switch module although the CU is capable of controlling several switch modules. The connection of additional CP50s to the duplicated high-speed time-division bus will increase the number of interfaces available. Each CP50 will provide up to 480 terminal interfaces. Of these, 240 will be for terminals operating in the synchronous packet-mode of transmission in accordance with CCITT Recommendation X25; some of these packet-mode interfaces will however provide connections for trunks to other PSEs. There will also be 240 interfaces for terminals operating in the asynchronous character-mode of transmission in accordance with CCITT Recommendation X28. Character-mode terminals will be connected directly to the PSE, or via the PSTN or, where available, national PDNs. In addition to those interfaces catering for the connection of users' terminal equipment, data bases, and inter-PSE trunks, the CP50 in the London PSE also provides connections to the NMC. The NMC, whose interface appears to the network as that of a standard packet-mode terminal in accordance with CCITT Recommendation X25, will have links to two synchronous packet-mode interfaces on the London CP50.

All packet-mode data terminals will be connected directly over links to individual synchronous packet interfaces on a CP50. Where distances are considerable, use will be made of synchronous time-division multiplexors to economize on line plant. In this case, the distant packet-mode terminals will be directly connected over links to the input of a synchronous multiplexor at a location remote from the PSE. A synchronous multiplexor at the PSE will then provide individual demultiplexed links corresponding to those of the distant packet terminals, and these will be connected to individual synchronous packet interfaces on the CP50. Synchronous time-division multiplexors will be used to enable packet-mode terminals in Belgium, Denmark, Ireland, Luxemburg and the Netherlands to gain access to their parent PSEs. Start-stop character-mode terminals will be connected in a different way. Terminals in countries having a PSE will be connected to an asynchronous time-division multiplexor directly, or via the PSTN, or via a PDN if available. For terminals sited in a country not having a PSE, access is available via an asynchronous time-division multiplexor whose aggregate bit stream from the terminals can be transmitted directly to the asynchronous multiplexor at the PSE. Alternatively, if there are remotely sited packet-mode terminals connected via a synchronous time-division multiplexor to the PSE, the bit stream can be transmitted from the synchronous multiplexor. The asynchronous multiplexors are of the type used in the Post Office Dataplex 2 service[7] and manufactured in the UK.

9 EURONET Software

Software for the Mitra 125 and CP50 processors for EURONET is based on that being developed by the French software house SESA for TRANSPAC. The production of software is being considered as a part of the development of the whole system rather than as a separate entity. In this way it is hoped to avoid any uncertainties of responsibilities at the boundary of analysis and programming. The 'system' is taken to mean the techniques that constitute EURONETs particular packet-switching design and which provide the desired facilities. The system includes the implementation of these techniques as software together with the connection of users and the handling of the appropriate protocols. These will include the following: line and link level management, including packet assembly/disassembly which will be performed by the CP50s; management at the packet level, which will be performed by the CUs and the CP50s, and the control of the network, which will be performed by the CUs and the NMC. The system functions, classified as transmission, call management or operating functions, to be implemented as software are as follows:

(a) in the CU,

Transit point management (call management)
Signalling management (call management)
Operating functions;

(b) in the CP50 'group unit',

Transit point management (call management)
Multiplexing/demultiplexing (transmission)
Link management including protocol conversion (transmission);

(c) in the CP50 'synchronous and asynchronous line units',

Line management (transmission).

The operation of the system software is a highly complex matter and is beyond the scope of this paper.

10 The Future of EURONET

At the commercial opening date of the network in 1979 EURONET will cater solely for the initial needs of the EEC Commission. Extra terminal interfaces will, however, be provided early in 1980 to cater not only for the additional requirements of the EEC Commission but also to meet requirements of possible non-EURONET users. In fact, one of the early non-EURONET users of the telecommunications network will be the European Informatics Network (EIN).[8] By the end of 1979 it is expected that EIN Network Centres, listed in Table 4, will be connected to EURONET as X25 terminals and at the same time, when EURONET is extended to Switzerland, they will again be joined by the Zurich centre (who withdrew from EIN late in 1978). The EIN centres will then not only be able to intercommunicate as before but they will also be able to gain access to the various information data bases connected to EURONET.

One can of course speculate as to what will happen in regard to data transmission services in Europe and, within the CEPT, the preparation of an Action Plan[9] for the establishment of a European Data Network is in hand. The role that EURONET might take in the establishment of such a network will be a key factor. Already it is known th. Switzerland will join the network with a node in Zurich in 1980; Spain and Sweden are anxious to join the network as well, so that users in these countries can access the EEC-based data bases, and recently Austria, Yugoslavia, Norway and Greece have made requests to be connected to EURONET. Since access to EURONET is planned to be via national data networks as and when they exist, it is possible that EURONET could form the basis of an international transit data network linking national networks in the various countries. Figure 7 illustrates the European scene as it might appear in 1984.

However an alternative approach is possible and one which is gaining considerable momentum in Europe. In order to connect national packet-switched data networks or non-EEC networks to EURONET, it has recently been agreed that this should be by means of the CCITT Recommendation X75, the protocol for network interworking. The availability of networks having this interface in, say, the United Kingdom and in the Netherlands will enable the two national networks to be interconnected directly. DIANE traffic between the UK and the Netherlands could also flow on this new link when it is available and the EURONET link could be dispensed with. Similar situations could occur between other countries and the success of EURONET could thus be its gradual disappearance as the traffic is gradually transferred to the emerging European Data Network.

Table 4

EIN—initial network centres

France	Paris	Institut de Recherche d'Informatique et d'Automatique
Italy	Ispra	EURATOM Centre
Italy	Milan	The Polytechnic Institute of Milan
Switzerland	Zurich	The Federal Institute of Technology (ETH)
United Kingdom	London	The National Physical Laboratory

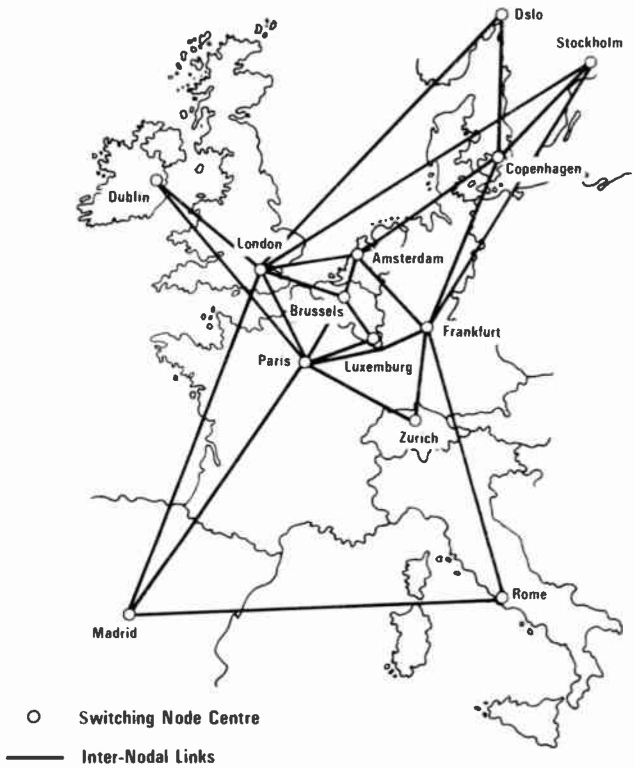

Fig. 7. EURONET—Possible interconnections—1984.

This concept is in line with the agreements reached between the Community and the PTTs which clearly states that in the longer term EURONET will gradually be integrated into the planned European Data Network. Based on current plans it is envisaged that this could well occur by about 1984 but users should not know that they have in fact been transferred to a basically different network since their network interfaces, X25, X28 and X29, will have remained unchanged. Already plans exist for users in the United Kingdom and in France, who initially will access the London and Paris EURONET nodes directly, to be transferred gradually to the respective national networks once these are connected to EURONET. This should result in a reduction of national access charges to the advantage of the user, international charges remaining the same.

The Commission is also anxious to enable users in the USA, in Japan and in the Third World countries to be able to access the Community data bases. As international data services become available throughout the world such connections will be readily available and the opening in 1978 of international data services between the United Kingdom and the USA via IPSS[10] will ease this access once IPSS and EURONET are linked together. To meet the requirements of a public service, however, certain elements of EURONET may need to be enhanced. In particular a second NMC could be needed, and additional transmission links will be required to increase the network's availability to users. The Community-based PTTs are also looking at other ways to enhance the network once it is in service. For example, the remote multiplexor sites could have switching equipment installed, thus providing PSEs in those countries which previously had none. As already mentioned consideration is also being given within CEPT to the extent to which the network can be integrated within a European Data Network to provide an international packet-switching data service and this could well be the preferred approach eliminating the need for special EURONET nodes in some countries.

Once the CCITT has standardized the international inter-working protocol between national packet-switching networks (such as the protocol defined in CCITT Recommendation X75), the EURONET packet-switching network could, as already mentioned, be connected to other data networks in the world. However, while this will enable calls to be established, satisfactory communication can take place only if the user-to-user application protocols, which were referred to earlier, are developed and standardized. The EEC Commission has initiated development of EURONET data base access and retrieval protocols which, when available, will enable any user to use the same protocols to gain access to any of the available data bases connected to the network. Amongst the protocols which are of considerable importance are those related to the Users' Systems. The standardization of retrieval languages will no doubt present problems and, if it can be achieved in time, a single standard command set will eventually facilitate on-line interactive searching of many different sets of data. Consideration is also being given to automatic language translation facilities to enable, at a later phase, a user's enquiries to be handled in his own language irrespective of the data base he is searching.

The user-to-host protocols for EURONET will need to embody all the rules specific to the information retrieval operations of EURONET such as accommodating differing retrieval procedures as well as a standard command set. The protocol will be developed in close consultation with the Commission and the CIDST, taking into account the relevant recommendations of international bodies such as the ISO, the Intergovernment Programme for Co-operation in the Field of Scientific and Technological Information (UNISIST) and the International Council of Scientific Unions Committee on Data for Science and Technology (CODATA). The guidelines will be that standard methods should be optional and exclude as few of the presently used systems as possible, leaving it to the user to decide on a standardized or a systems specific approach to any particular enquiry. The Commission claims this protocol, with its opportunity for standardization, will eventually enable the users to use a large variety of data bases and retrieval systems in an easy and uniform way.

The planning and implementation of the EURONET packet-switching network has been a major task involving the close co-operation of the Community-based PTTs and the Commission of the EEC. Progress has been significant since the announcement of the plans to establish the network and this has served to concentrate the activities of the PTTs in Europe, and the users, on establishing and agreeing a whole range of standards for data communications which should be of world-wide benefit.

Although being set up initially as a 'public' network with limited access, 'public' because many users share common facilities, EURONET will have the capability to be expanded to meet not only the specific needs of the Commission but also those of commercial enterprises that wish to use international data transmission facilities.

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12 Appendix 1—Abbreviations

CCITT	International Telegraph and Telephone Consultative Committee
CEPT	Conference of European Postal and Telecommunications Administrations
CIDST	Committee for Information and Documentation on Science and Technology
CODATA	International Council of Scientific Unions Committee on Data for Science and Technology
CSTD	Special Committee for Data Transmission
CU	Command Unit
DIANE	Direct Information Access Network—Europe
EEC	European Economic Community
EIN	European Informatics Network
EPSS	Experimental Packet Switched Service
ETAG	Technical, Economic and Financial Group
HDLC	High-level Data Link Control
IPOEE	Institution of Post Office Electrical Engineers
IPSS	International Packet Switched Service
ISO	International Standards Organisation
LAP	Link Access Protocol
NUI	Network User Identification
PAD	Packet Assembly/Disassembly
PSE	Packet Switching Exchange
PSS	Packet Switched Service
PSTN	Public Switched Telephone Network
PTT	Post, Telegraph and Telephone Administration
UNISIST	Intergovernment Programme for Co-operation in the Field of Scientific and Technological Information

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SUMMARY

Serial two's complement pipeline multipliers are the basic module in the serial arithmetic implementation of digital signal processing algorithms. These multipliers accept the data serially in two's complement notation and generate a serial output product in two's complement notation as well. The designs, however, presented up to now lack in modularity; in addition, they have the problem of the internal overflow. In this paper a new two's complement multiplication algorithm is worked out, that allows cellular implementation of serial-pipeline multipliers and does not limit the dynamic range of the data input. In addition, by using the proposed cell, useful functions for digital signal processing can be performed. Methods are proposed for overflow detection, and comparison is made with previously proposed realizations.

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1 Introduction

The serial-pipeline multiplier was first proposed by Jackson *et al.*[1] for digital signal processing applications. This multiplier accepts the data serially and the coefficients in parallel form, and generates a serial output having the same length as the input. The basic characteristic of a pipeline multiplier is that there is no delay between consecutive input words or the corresponding consecutive output words. This first serial pipeline multiplier handled only positive numbers. For signed numbers in two's complement notation, a converter from two's complement to sign and magnitude was used at the input. The results were converted back to two's complement at the output.[2] For the conversion at the input, one would have to wait for the most significant bit which contains the sign information. This is the main disadvantage of this modified pipeline multiplier, because the process is delayed by one data word. The second disadvantage is the non-modularity of the design, because of the separate sign handling.

For direct handling of two's complement data, pipeline multipliers were proposed by Lyon.[3] In his method the coefficients are in two's complement or sign-magnitude notation or recorded in Booth's algorithm representation. The implementation of these multipliers is based on the addition of two's complement partial products by sign extension. The realizations that result from this approach, however, have a few disadvantages. The first is the non-modularity of the last two stages for sign-magnitude and two's complement representation of the coefficient. For the Booth's algorithm representation, non-modularity exists only at the last stage.

The second disadvantage of the above implementations is that they do not work correctly for the whole dynamic range of the input data. This problem arises from the sign extension method where the last carry of each partial addition is truncated and internal overflow could occur. In order to avoid such a condition, the sign bit must be repeated and thus the dynamic range of the input is reduced by a factor of two.

In this paper a new two's complement multiplication algorithm is developed, that offers a fully cellular implementation. At the same time it overcomes all the other disadvantages and limitations of Lyon's approach. The basic cell used in the implementation of this algorithm is presented. This cell is then used in the implementation of two distinct cases:

(a) The coefficient magnitude is less than one, which is the case in fast Fourier transform applications.
(b) The coefficient magnitude is less than two, which is the case for a wide range of digital filter applications.

In the first case the new algorithm leads to a fully modular implementation. In the second case the modularity is maintained by inserting a zero bit before the least significant bit of the incoming data if the data

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magnitude is less than 0·5, or by adding an external modification to the last cell only. For the second case we also examine the overflow problem and obtain the conditions for overflow detection.

Before presenting the main results of this work we give a brief discussion of the operation principles of serial-pipeline multipliers.

2 Operation Principles of Serial-pipeline Multipliers

For this general consideration the data and the coefficient are restricted to be positive and less than two. The expressions for the multiplier and the multiplicand are

$$X = \sum_{i=0}^{N-1} x_i 2^{-i} \quad x_i = 0, 1$$

$$Y = \sum_{j=0}^{M-1} y_j 2^{-j} \quad y_j = 0, 1$$

Suppose that there is no overflow, that is the product is in the same range as the data. Then the product can be written as follows:

$$X \,.\, Y = \sum_{k=0}^{M+N-2} P_k 2^{-k} \tag{1}$$

where

$$P_k = \sum_{j=0}^{k} x_j y_{k-j}$$

and

$$x_i = 0 \quad \text{for } i \geqslant N$$
$$y_j = 0 \quad \text{for } j \geqslant M$$

The serial-pipeline multiplier is based on eqn. (1). The P_k products are the quantity that is implemented at each step. The process begins from the l.s.b. of the product, that is from $k = M+N-2$. The P_k term and the carries from the previous step are added to produce the product bit of weight k as well as to generate the carries for the next step. The product bits with weight $k > N-1$ are truncated. In Fig. 1 the hardware implementation of this product is shown. Truncation is provided by the control signal that is propagated along with the data. It is one for all data bits except the m.s.b. for which it becomes zero.

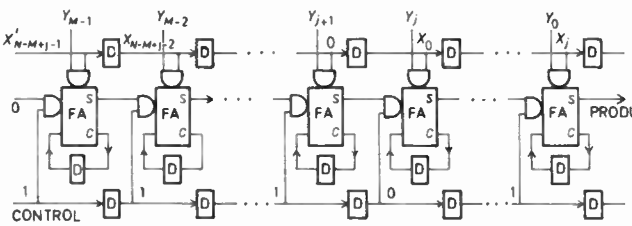

Fig. 1. The serial-pipeline multiplier for positive numbers.

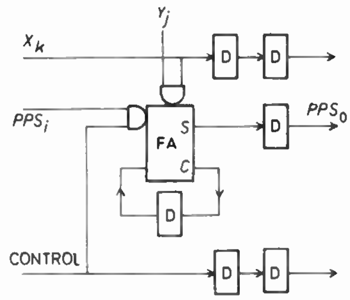

Fig. 2. The rudimentary cell of a pipeline multiplier.

The truncation prevents the mixing of partial products from different multiplications.

One disadvantage with the serial pipeline multiplier of Fig. 1 is that there is a delay before each bit appears at the output, that is equal to the propagation delay through all the modules. To speed up the process a one bit delay is inserted between the adders. For synchronization one more delay must be added in the data line of each stage. So the basic module takes the final form shown in Fig. 2.

3 The Proposed Two's Complement Serial-pipeline Multiplier

The new two's complement multiplication algorithm is introduced in this Section. This algorithm will be applied for the implementation of two's complement serial pipeline multipliers. The handling of the negative terms has been treated by other workers in the past. Lyon's work was mentioned in the Introduction. In Ref. 4 the addition of corrective terms is suggested. Booth[5] proposed an algorithm that is based on signed digit coding of the coefficient. In Refs 6, 7 and 8 use is made of the sign extension technique and the implementation is done by an iterative array. Finally, there are algorithms that handle the negative terms by the addition of their complements.[9,10,11] The proposed method belongs in this last case.

As pointed out in the Introduction, two cases are distinguished for the size of the coefficients Y: $|Y| < 1$ and $|Y| < 2$. These two cases are investigated separately.

3.1 Coefficient Magnitude Less Than One

Consider now the data and the coefficient represented in two's complement notation,

$$X = -x_0 + \sum_{i=1}^{N-1} x_i 2^{-i} \tag{2a}$$

$$Y = -y_0 + \sum_{j=1}^{M-1} y_j 2^{-j} \tag{2b}$$

The data can always be normalized to be less than one in magnitude. Since we have assumed $|Y| < 1$, the magnitude of the product will be less than one. On the basis of this observation the product is written in the equivalent form,

$$A = (X \,.\, Y + 2) \bmod 2 \tag{3}$$

This is the two's complement representation of the product $X \,.\, Y$ because,

(a) if $X \,.\, Y > 0$, then the term 2 is an overflow
(b) if $X \,.\, Y < 0$, then the term A is the two's complement of $X \,.\, Y$ by definition.

Carrying out the operations specified by the definition of A in eqn. (3), the following expression for A is obtained:

$$A = \Bigg[\sum_{i=1}^{N-1} x_i 2^{-i} \sum_{j=1}^{M-1} y_j 2^{-j} + \left(1 - \sum_{j=1}^{M-1} y_j x_0 2^{-j}\right) + + y_0\left(1 - \sum_{i=1}^{N-1} x_i 2^{-i}\right) + \bar{y}_0 + x_0 y_0 \Bigg] \bmod 2$$

where the identity $y_0 + \bar{y}_0 = 1$ was used.

The above expression for A can be rewritten by using the following two identities that can be easily verified:

$$\left[1 - \sum_{i=1}^{M-1} a_i 2^{-i}\right] = 2^{-M+1} + \sum_{i=1}^{M-1} \bar{a}_i 2^{-i} \tag{4a}$$

and

$$x_0 y_0 + \bar{y}_0 = \overline{\bar{x}_0 y_0} \tag{4b}$$

Substituting these two identities in the expression for A, we obtain

$$A = \Big[\sum x_i 2^{-i} \sum y_j 2^{-j} + y_0\left(\sum \bar{x}_i 2^{-i} + 2^{-N+1}\right) + + \left(\sum \overline{y_j x_0}\, 2^{-j} + 2^{-M+1}\right) + \bar{x}_0 y_0 \Big] \bmod 2 \tag{5}$$

The mod 2 notation in eqn. (5) means that the carries will not be considered.

The algorithm of eqn. (5) is implemented with the basic cell of Fig. 3. The complete implementation of the two's complement serial-pipeline multiplier is shown in Fig. 4.

The basic proposed cell has two switches and one EX-OR gate added on to the structure of the cell for positive numbers, shown in Fig. 2. The EX-OR is used to realize the third term of eqn. (5). The switches are controlled by the control bits. The first one, SW1, is used to send the last carry to the additive input of the next cell, thus truncating the partial product sum and avoiding one step delay. The second one, SW2, provides one additive input S_i at the same time with the l.s.b. of the data. These inputs S_i are set to zero except for the last two stages. In the next to last stage S_i is made equal to one, thus adding a term of weight 2^{-N} and converting the truncation to rounding. In the last stage S_i is made equal to y_0, to realize the term $y_0 2^{-N+1}$ of eqn. (5). As shown in Fig. 4, a NOT gate is added at the data input of the last cell to realize the term $y_0 \sum \bar{x}_i 2^{-i}$ as well as the term $\bar{x}_0 y_0$ of eqn. (5). The term 2^{-M+1} of eqn. (5) is added to the first stage by setting the carry input C_i to that stage equal to one. The control line bits are ones for all data bits except for the m.s.b. for which it becomes zero.

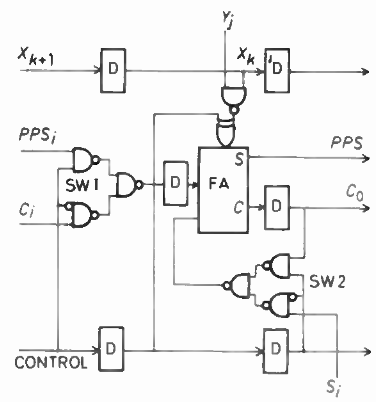

Fig. 3. The proposed two's complement multiplier cell.

3.2 Coefficient Magnitude Less Than Two

In this case coefficients in two's complement and sign-magnitude representation are considered. The cell of Fig. 3 is still the basic building block and the pipeline realization of Fig. 4 is utilized with a small external modification. Since the magnitude of the coefficient may be greater than one, this will cause an overflow product for some data range. Overflow detection circuits are suggested for the cases considered.

3.2.1 *Coefficient in two's complement*

The coefficient in two's complement notation is

$$Y = -2y_0 + \sum_{j=1}^{M-1} y_j 2^{-j+1} = 2\left(-y_0 + \sum_{j=1}^{M-1} y_j 2^{-j}\right) \tag{6}$$

The term in parenthesis has the same representation as in eqn. (3), the only difference being the factor 2. So, if the multiplier of Fig. 4 is used for generating the product with the coefficient of eqn. (6), the output N bits will have the fixed point format $P_0 P_1 \,.\, P_2 \ldots P_{N-1}$. If this number is to be used in subsequent stages, as is usually the case, one must see under what conditions it can be converted to the input data format, that is, $P_1 \,.\, P_2 \ldots P_N$. If the product is not greater than one, the bit P_0 is redundant information and can be discarded. At the same time the last truncation (P_N) of the next multiplication must be inhibited in order to provide at the output the same fractional bits as at the input. These two operations are performed with the bypass circuit of Fig. 5. This circuit truncates the m.s.b. of a product and replaces it with the P_N of the next multiplication. To convert the truncation to rounding, the input S_i in the cell of coefficient bit y_2 is set to one.

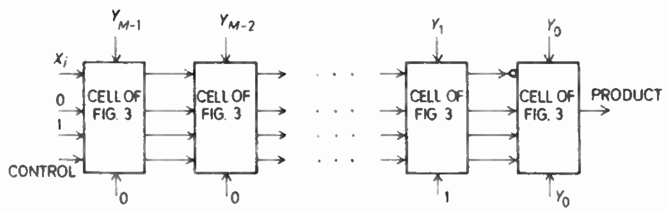

Fig. 4. The two's complement pipeline multiplier for $|Y| < 1$.

If the product magnitude is greater than one, then the discarded bit, P_0, carries significant information and we have an overflow situation which must be detected. The overflow condition is that the last two bits (P_0, P_1) be different. This condition is detected by the circuit shown also in Fig. 5.

An interesting observation that can be made from eqn. (6) is that the data can be multiplied by the factor 2 instead of the product, provided that the magnitude of the data is less than 0·5. This multiplication by 2 can be implemented by inserting a zero bit before the l.s.b. of the input data word, while at the same time discarding the m.s.b. of the preceding data. This is equivalent to shifting the data one bit to the left. With this reforming of the input data, the multiplier of Fig. 4 can be directly applied. Thus, a fully modular realization is obtained.

3.2.2 *Sign-magnitude coefficient*

In this representation the coefficient can be expressed as follows:

$$Y = \operatorname{sgn}(Y)\,.\,|Y| \tag{7a}$$

where

$$|Y| = \sum_{j=0}^{M-2} y_j 2^{-j} \tag{7b}$$

and

$$\operatorname{sgn}(Y) = (-1)^{y_{M-1}} \tag{7c}$$

The y_{M-1} bit operates on the data and converts it to a new sequence defined by

$$\hat{X} = \operatorname{sgn}(Y)\,.\,X = -\hat{x}_0 + \sum_{i=1}^{N-1} x_i 2^{-i} \tag{8}$$

Using eqns. (7) and (8), the two's complement representation of the product $\hat{X}\,.\,|Y|$ in mod 4 can be written as follows:

$$(\hat{X}|Y|+4)\bmod 4 = \left[\left(\sum_{j=0}^{M-2} y_j 2^{-j}\right) \times \times \left(-\hat{x}_0 + \sum_{i=1}^{N-1} \hat{x}_i 2^{-i}\right) + 4\right]\bmod 4$$

$$= \left[\sum_{j=0}^{M-2} y_j 2^{-j}\,.\sum_{i=1}^{N-1} \hat{x}_i 2^{-i} + +2 - \sum_{j=0}^{M-2} \hat{x}_0 y_j 2^{-j} + 2\right]\bmod 4$$

Applying eqn. 4a, the previous expression takes the form,

$$(\hat{X}|Y|+4)\bmod 4 = \left[\sum_{j=0}^{M-2} y_j 2^{-j} \sum_{i=1}^{N-1} \hat{x}_i 2^{-i} + + \sum_{j=0}^{M-2} \overline{y_j \hat{x}_0}\, 2^{-j} + 2^{-M+2} + 2\right]\bmod 4$$

$$= (P+2)\bmod 4 \tag{9}$$

Suppose that the quantity P gives a carry C of weight 2. Then, the last digit (m.s.b.) of the product will become,

$$(2C+2)\bmod 4 = 2\bar{C} \tag{10}$$

That is, the last digit of the product is the complement of the carry of the quantity P. The non-overflow condition is $\bar{C} = P_0$ (the m.s.b. of P). The overflow detection circuit is shown in Fig. 6.

Let us assume now that there is no overflow. Then the product can be computed in mod 2 and is equal to the quantity P. This product is easily implemented with the arrangement of Fig. 7. In this realization the cell of Fig. 3 is used as the controlled complementer. This configuration gives us the additional capability of adding a word Z synchronously with the input data X, thus performing the function $(X+Z)Y$.

The controlled complementer could be located just as well at the end of the multiplier, thus redefining the

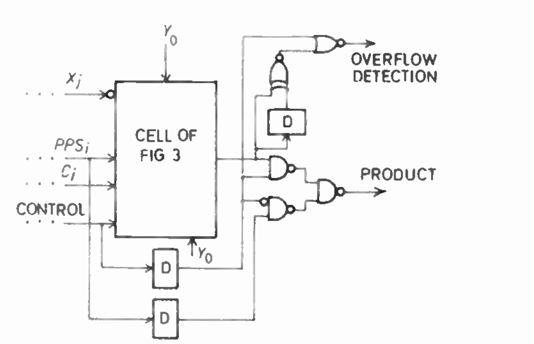

Fig. 5. The last stage for a coefficient with $|Y| < 2$ and in two's complement representation.

output word in a manner similar to that of eqn. (8). As before, an input word Z can be added synchronously with the product thus giving a function of the form $XY+Z$. The functions $(X+Z)Y$ and $XY+Z$ are very useful, because, by repetitive application of either one of these operations, any digital signal processing algorithm can be realized.

4 Performance

In serial digital data processing, pipeline multipliers are widely accepted as the basic elements mainly because

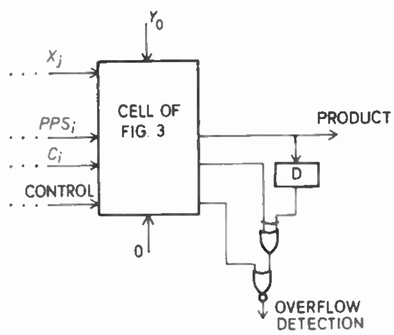

Fig. 6. The overflow detection circuit for $|Y| < 2$ and in sign-magnitude representation.

they provide high data bit rates. In m.s.i. realizations of the two's complement pipeline multiplier,[12,13] the algorithms proposed by Lyon have been used up to now. A comparison of these realizations with other existing realizations based on the serial-parallel multiplication approach has been presented by Baldwin *et al.*[13] This comparison is repeated here in Table 1 for convenience.

In this Table a figure of merit is computed from the expression,

$$F = \frac{f}{24} \cdot \frac{1}{PN_p} \cdot \frac{1}{K} \cdot 10^{-6}$$

where F represents the number of million multiplications per second per watt. The scaler K is the output data length relative to 24-bit data. From this Table it is seen that the realization based on the two's complement/sign magnitude pipeline multiplier[13] is superior to all others in terms of speed. The c.m.o.s./s.o.s. realization of the serial-parallel approach gives a high figure of merit with a great sacrifice, however, of speed.

The two's complement pipeline multiplier that is proposed here has essentially the same number of gates per cell and the same propagation delay as Lyon's basic cell. Specifically, it has 52 gates and a delay of $6\tau_g$, where τ_g is the gate delay. In the final realization the delay will be reduced to $4\tau_g$ by moving the switch SW1 of Fig. 3 after the D flip-flop, thus introducing further synchronization within the cell.

From the above discussion it follows that the proposed pipeline multiplier will have the same performance characteristics as that of Lyon, if it is fabricated with the technique that is explained in Ref. 13. This technique employs a high-speed bipolar process,

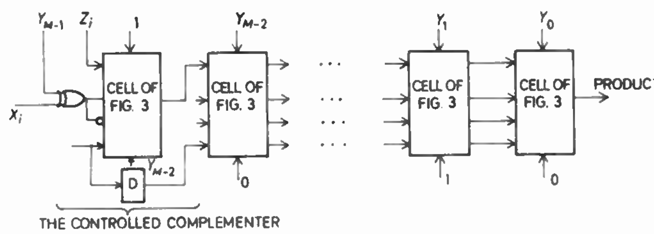

Fig. 7. The full multiplier for $|Y| < 2$ and in sign-magnitude representation, including the capability to generate the function $(X+Z)Y$.

using a combination of emitter-function logic, stacked current switches and two-level metallization. Fabrication with this technique will give an m.s.i. package of four cells with power dissipation of 140 mW and a speed of 40–50 Mbit/s. This package will require 16 pins.

The above realization, while keeping all the performance figures of Lyon's realization, will offer to the engineer certain functional advantages which improve significantly the operation of the pipeline multiplier. These functional advantages are summarized below:

(a) The algorithm works for the whole dynamic range of the data input without the internal overflow possibility that shows up in Lyon's realization.

Table 1

Comparison of several approaches to serial multiplication of 24-bit data by a 12-bit coefficient

Multiplier	Size	Number of Packages Required Np	Data Rate f (Mbit/s)	Power per Package P (mW)	Scaler K	Figure of Merit F
Two's complement/ signed multiplier[13]	$4\times N$	3	44	140	1	4·4
Two's complement two's complement[12]	$4\times N$	3	20	150	1	1·8
C.M.O.S./S.O.S.[14]	$24\times N$	0·5	10 (5 V)	20	1·5	28
C.M.O.S./S.O.S.[14]	$24\times N$	0·5	18 (15 V)	420	1·5	2·4
T.T.L.[15]	$8\times N$	1·5	25	455	1·5	1·02
N.M.O.S.[16]	$12\times N$	1	7	245	1	1·2

(b) As explained in the text, the required rounding operation is performed automatically. This is in sharp contrast to Lyon's approach, where it can only be performed by adding one at the input of the first cell. The timing of this addition requires additional external circuitry. Thus, it cannot be realized with an iterative structure and for this reason it was not included in the practical realization of Ref. 13.

(c) A synchronous addition of the form $XY+Z$ can be performed by introducing a small modification in the first or the final cell of the proposed design. This modification of the basic cell can be easily implemented with the fabrication technique explained in Ref. 13, by using the two-level metallization process. Such an operation is of major interest in the implementation of digital signal processing algorithms.

5 Conclusions

A new serial-pipeline multiplication module is developed, based on a new two's complement multiplication algorithm, which works correctly for the whole dynamic range of the data input. This module, besides having a relatively small circuit complexity, is flexible and can be used as a basic building block for the construction of various multipliers. Also, useful functions can be realized with it for digital signal processing applications. Finally, this module lends itself to m.s.i. realization that will yield a chip with improved overall performance in comparison to existing similar realizations.

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Indexing Terms: Semiconductor devices, Silicon-on-sapphire, Integrated circuits, Transistors, metal oxide field effect

Very-high-speed silicon-on-sapphire integrated circuits

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SUMMARY

Very-high-speed silicon-on-sapphire (s.o.s.) n-channel m.o.s.f.e.t.s have been fabricated which demonstrated a maximum oscillating frequency of approximately 6·4 GHz. These m.o.s.f.e.t.s—test transistors on an integrated circuit—were fabricated with a process which can be used to fabricate high-speed integrated circuits.

Spacecraft data processing requirements are evolving to indicate the need for very-high-speed devices which are capable of being integrated into large-scale circuits. Improvements in high-speed data processing communications systems are badly needed. Many of these applications will be met be very-high-speed s.o.s. integrated circuits.

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1 High-frequency Performance of S.O.S.

The silicon-on-sapphire isolation technique is especially attractive for high-speed l.s.i. because the sapphire dielectric isolation process results in less stray capacitance which, in turn, results in lower power and higher speed.[1] A silicon-on-sapphire m.o.s.f.e.t. is illustrated in Fig. 1. P-type regions were formed by implanting boron into the channel region. Phosphorus ion implantation was used to form a self-aligned-gate structure. Thin silicon dioxide (500 angstroms) was used for the gate insulator. Aluminium gate metallization was used to enhance the speed performance by eliminating gate contact resistance normally encountered when silicon-gate construction is used.

Figure 2 shows the s.o.s. m.o.s.f.e.t. device geometry. Very-high-speed performance was achieved using a short-channel device structure, i.e. a gate length of 1·5 μm. Figure 3 shows the s.o.s. m.o.s.f.e.t. 3-terminal electrical characteristics.

Table 1 shows the speed calculations for this s.o.s. m.o.s.f.e.t. in comparison with a gallium arsenide m.e.s.f.e.t. having an identical geometry. Using measured values of gate-to-source capacitance and transconductance, the cut-off frequency for the s.o.s. m.o.s.f.e.t. is 6·4 GHz.

S-parameter measurements were used to characterize the high frequency performance of the s.o.s. m.o.s.f.e.t.s. Measurement parameters included frequency, S_{21}, stability factor (k), power gain (G_{max}), matched input impedance and matched output impedance. Figure 4 illustrates the *S*-parameter test configuration used to characterize the transistor high frequency performance. Table 2 gives the s.o.s. m.o.s.f.e.t. high frequency test data. Data was obtained over the frequency range from 2 GHz to 6 GHz.

Figure 5 shows the gain versus frequency characteristics for typical n-channel s.o.s. m.o.s.f.e.t.s. This plot shows a cut-off frequency of approximately 6·4 GHz. The cut-off frequency of transistors in a digital integrated circuit can be related to the practical upper limit of system clock frequency for a digital i.c. Using good design practice, the cut-off frequency of transistors in the integrated circuit should exceed the system clock frequency by approximately five times. Using this design guideline, digital integrated circuit operation at clock frequencies up to 1 GHz are expected from integrated

Table 1

Speed calculations

	C_{gs} (fF μm^{-2})	g_m (mS μm^{-1})	$f_t = g_m/2\pi C_{gs}$ (GHz)
GaAs m.e.s.f.e.t.	1·0	0·08	12·7
S.o.s. m.o.s.	0·5	0·02	6·4

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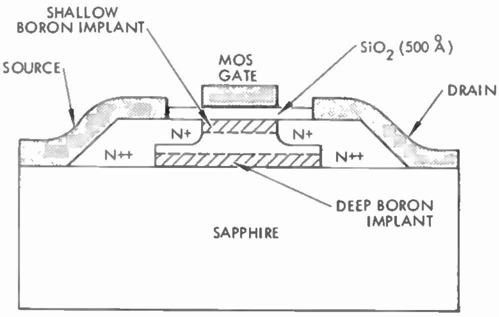


Fig. 1. Silicon-on-sapphire m.o.s.f.e.t.

Epi = 0·7 μm   Aluminium gate
$N^- = <8\times10^{14}$ cm^{-3} $L_G \sim 1{\cdot}5$ μm

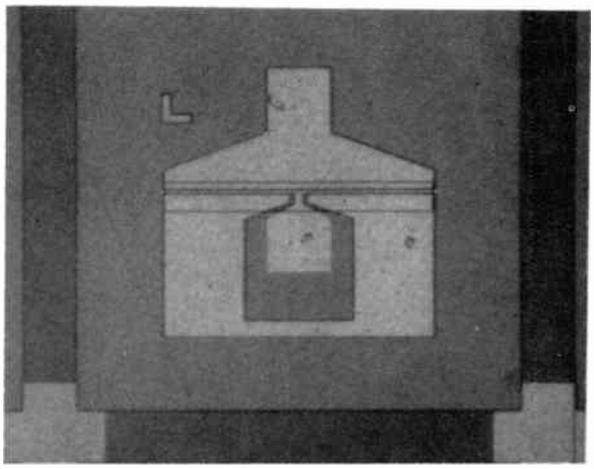


Fig. 2. Silicon-on-sapphire m.o.s.f.e.t. device geometry.

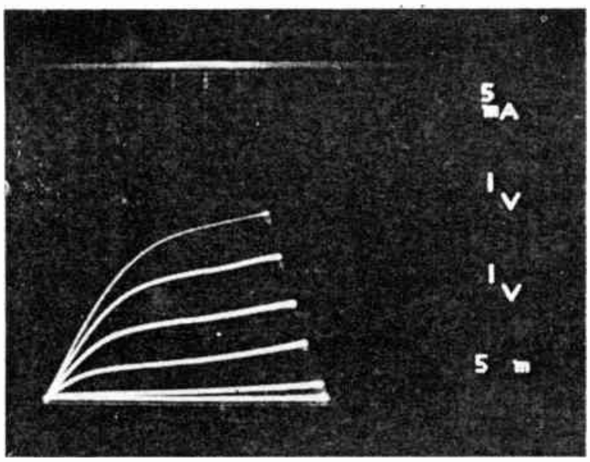


Fig. 3. S.o.s. m.o.s.f.e.t. characteristics.

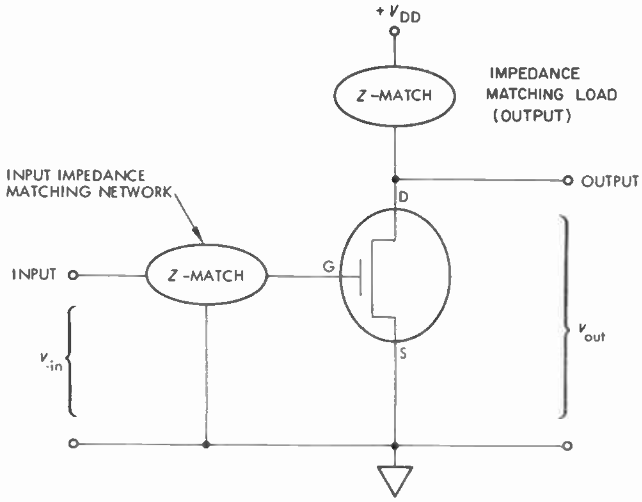


Fig. 4. Silicon-on-sapphire transistor high frequency test.

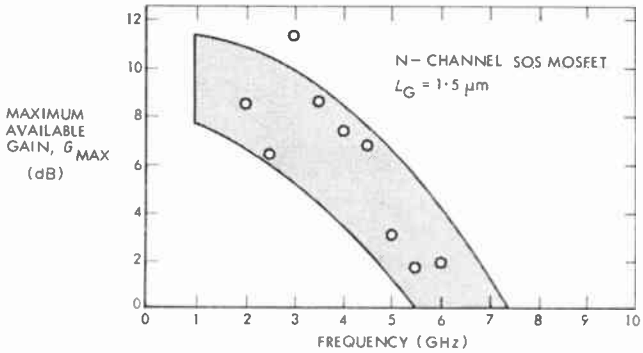


Fig. 5. Gain versus frequency for n-channel s.o.s. m.o.s.f.e.t. showing cut-off frequency = 6·4 GHz.

circuits using transistors having speed performance characteristics described by the data in Fig. 5.

2 Integrated Circuit Technology Development

Programmes are under way in U.S. semiconductor companies to develop high-speed integrated circuits. These programmes have clear-cut goals, despite the fact that these development programmes are looking out as far as 1985 and beyond.

S.o.s. has been proposed as one of the baseline technologies for the Department of Defense programme for very-high-speed integration, or v.h.s.i. The U.S. Department of Defense plans to spend $200M over 6 fiscal years, beginning in 1979, to develop v.h.s.i. circuits with stringent military specifications such as failure rates of 0·1% or less per 1000 hours at 125°C.

Very-high-speed integrated circuits will utilize the v.l.s.i. techniques currently being developed. From tens of devices per chip to tens of thousands—the increase over the last ten years in the number of transistors in an integrated circuit has been truly explosive. The growth process has been evolutionary—continual process improvement and the confidence gained with production has enabled manufacturers to shrink transistor dimensions and increase producible numbers of dice per wafer.

The increasing complexity of logic circuits, layout topology, and lithographic processes have compounded to create technical challenges of such proportions that multi-year development programmes are required.

Table 2

S.o.s. m.o.s.f.e.t. high frequency S-parameter test data

Frequency (MHz)	S_{21} (dB)	K	G_{max} (dB)	Z match in $R+jX$		Z match out $R+jX$	
2000	−3·9	1·12	8·4	7·7	99·3	15·5	99·9
2500	−2·3	1·19	6·3	9·9	71·4	15·8	97·6
3000	−3·8	0·96	11·4	17·7	110·5	6·5	76·4
3500	−3·2	0·87	8·4	7·0	57·6	10·4	53·4
4000	−2·1	0·97	7·4	8·5	36·3	7·1	36·5
4500	−1·2	0·38	6·8	11·1	23·6	5·0	13·5
5000	−1·1	1·14	3·0	6·2	5·7	9·5	2·9
5500	−1·1	1·44	1·9	13·3	−9·1	12·6	1·8
6000	−1·5	1·81	2·0	14·4	−5·1	12·9	−6·4

Lockheed Test Report; tests performed by independent test laboratory operated by Western Automatic Test Service, Inc

Anticipating unprecedented process development and design problems—and greater risks of investment capital, manpower, and development time than ever before—semiconductor R&D laboratories and manufacturers have initiated long-term programmes to address the challenge of what has come to be called very-high-speed integration (v.h.s.i.).

3 Technology Comparison

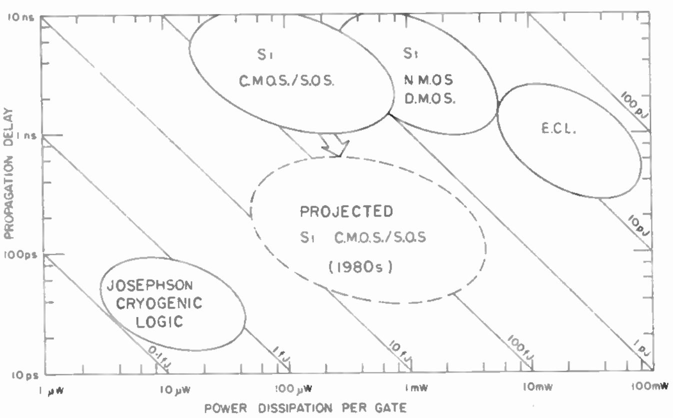
Gallium arsenide has been used for years to build microwave transistors because it is a better semiconductor than silicon for very-high-speed applications—at electric fields less than 3×10^6 V cm^{-1}, electrons in GaAs can move at nearly 10 times the speed of electrons in silicon. Gallium arsenide will provide a twofold to sixfold speed advantage over silicon in l.s.i. and v.l.s.i. circuits, based upon current projections, as shown by the comparison of data in Figs. 6(a) and 6(b).

The device under development for integrated circuit use is the n-channel metal-semiconductor field-effect transistor, or m.e.s.f.e.t. M.e.s.f.e.t.s, unlike metal-oxide-semiconductor f.e.t.s, have no gate-insulating oxide.

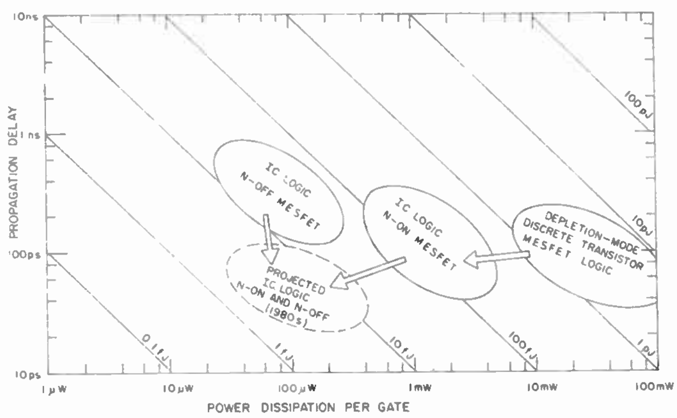
Gallium arsenide integrated circuits are being fabricated using f.e.t.s with channel lengths of one micron or less. Discrete microwave transistors with such small dimensions have been produced for several years in the form of depletion-mode f.e.t.s. These normally-on devices have extremely fast switching speeds—delay times less than 100 ps—but gates using these transistors dissipate 20 to 40 mW each. This high power dissipation of depletion-mode f.e.t.s limits the level of integration to no more than a few hundred transistors on a single chip.

The enhancement-mode, or normally-off, f.e.t. is the key to l.s.i. gallium arsenide circuits. Though not quite as fast as depletion-mode devices, enhancement-mode f.e.t.s dissipate much less power, occupy less space, and can be operated with a single power supply voltage. Moreover, enhancement-mode devices plus depletion-mode loads are similar to the relatively simple direct-coupled f.e.t. logic designs used on silicon.

S.o.s. is a lower-cost, more-mature technology than gallium arsenide. Several thousand gates have already



(a) Silicon.



(b) Gallium arsenide.

Fig. 6. Silicon high-speed logic delay-power comparisons (1979).

been fabricated on a single chip, and both digital and analogue circuits have been combined on a single chip. Because of its advanced maturity, s.o.s. is expected to precede gallium arsenide in very-high-speed applications requiring large-scale integration. But, because gallium arsenide enjoys a fundamentally superior speed performance, gallium arsenide integrated circuits will find increasing use in very-high-speed applications. And, as technology matures, gallium arsenide l.s.i. circuitry is expected to find use in applications such as high-reliability satellite on-board signal processing systems, where the emphasis will be on speed performance, not low cost.[2]

Based upon process and circuit design calculations, we have established a projection for the development of an advanced n-channel s.o.s. integrated circuit technology using one-micron channel lengths. This projection is shown in Fig. 6(a). It is based upon extensions of the existing c.m.o.s./s.o.s. technology and detailed analysis of the process and circuit parameters involved in high-speed, low-power c.m.o.s./s.o.s. design. This analysis of propagation delay and power dissipation is as follows.

Propagation delay, t_{pd}, of a c.m.o.s. inverter is defined as the average value of rise-time, t_r, and fall-time, t_f, or

$$t_{pd} = \frac{t_r + t_f}{2}. \tag{1}$$

Rise-time is the time interval required for charging the load capacitance, C_L, by the load transistor's drain-to-source current, and is given by

$$t_r = \frac{2C_L}{\beta_L} f_L(V). \tag{2}$$

Fall-time is the time interval required for discharging the load capacitance, C_L, by the driver transistor's drain-to-source current, and is given by

$$t_f = \frac{2C_L}{\beta_D} f_D(V). \tag{3}$$

Combining (1), (2) and (3) we get

$$t_{pd} = \frac{C_L}{\beta_L f_L(V) + \beta_D f_D(V)}. \tag{4}$$

Load transistor current, I_L is given by

$$I_L = \beta_L f_L(V), \tag{5}$$

where

$$\beta_L = \frac{\mu_L \varepsilon_{ox}}{2t_{ox}} \left(\frac{W}{L}\right)_L, \tag{6}$$

$f_L(V)$ = function of V_{in}, V_{out}, V_{DD} and V_{TL} (threshold voltage of the load transistor)

and

$\left(\frac{W}{L}\right)_L$ = load device width-to-length ratio.

Driver transistor current, I_D, is given by

$$I_D = \beta_D f_D(V), \tag{7}$$

where

$$\beta_D = \frac{\mu_D C_{ox}}{2t_{ox}} \left(\frac{W}{L}\right)_D, \tag{8}$$

$f_D(V)$ = function of V_{in}, V_{out} and V_{TD} (threshold voltage of the driver transistor)

and

$\left(\frac{W}{L}\right)_D$ = driver device width-to-length ratio.

It can be seen from (4) that in order to minimize circuit delay, C_L must be made as small as possible and β_L, β_D must be made as large as possible but within the limitations set for power consumption.

In c.m.o.s./s.o.s. technology, C_L has been minimized through a reduction of junction capacitance which is achieved by replacing the p–n junction interface with a silicon–sapphire interface, by self-aligned gates which disposes of gate-overlap capacitances and by using smaller geometries which result in the reduction of gate oxide capacitance.

To make β_L and β_D large, t_{ox} and L_L and L_D must be made small (assuming W_L and W_D are constant).

Making t_{ox} small results in larger current but also in larger C_L. Some gain in speed can be obtained, however, since with thinner t_{ox}, C_g increases but C_g is only a part of the total load capacitance C_L; the other major part being the metal and polycrystalline silicon interconnect capacitance.

Reducing device channel length would result in a major speed increase. This is because β_L and β_D would increase proportionally and at the same time C_L would be reduced since gate geometries will be smaller.

As the channel length is reduced, V_T, μ and I_{DS} will change due to short channel effects.[4,5] This effect is very small at 5 µm channels but becomes increasingly more significant and cannot be ignored for channel length of 1 µm. As the channel length decreases V_T decreases, and in general I_{DS} will also decrease as a result of V_T and μ changes.

Some speed improvement can also come from $f_L(V)$ and $f_D(V)$, which are complex functions of supply voltage, input/output levels and threshold voltages. These functions can be made larger by reducing threshold voltages and using a higher-voltage supply. But both parameters are subject to limitations: Reducing the threshold voltages below presently used values will result in an increase in the percentage variation of its value and also lower noise immunity. If the supply voltage is increased, the power will increase by the square law. Also, the power supply voltage upper limit may be limited by the drain-to-source breakdown voltage if small geometry devices are used.

Power dissipation, P_d, of a c.m.o.s. inverter circuit is given[3] by

$$P_d = I_{DD} V_{DD}, \tag{9}$$

where

$$\begin{aligned} I_{DD} &= \text{total power supply current} \\ &= I_{LEAK} + I_C + I_{COV} \end{aligned} \tag{10}$$

I_{LEAK} = leakage current

I_C = capacitance current (dynamic current due to charging and discharging of the circuit capacitance)

and

I_{COV} = cross-over current (current flowing through the n-channel and p-channel devices during switching transitions).

To minimize power dissipation I_{DD} and V_{DD} must be made as small as possible. In c.m.o.s./s.o.s. technology, I_{LEAK} is made low by using a dielectrically-isolated sapphire substrate to minimize total junction area, by using special processes to prevent back channels and side channels (along the island edge), and by avoiding parasitic transistors formed due to field inversion.

The dynamic current of a capacitor is given by

$$I_C = CVf, \tag{11}$$

where C = capacitive load,
V = voltage across the capacitance
and f = operating frequency.

For a given voltage level and at a specified frequency, I_C can be minimized by reducing circuit capacitance. This can be achieved by using vertical-junction s.o.s. device construction to minimize junction capacitance, and by gate overlap capacitance elimination in the self-aligned-gate process of the c.m.o.s./s.o.s. technology, and by reducing device geometries to a minimum.

Cross-over current, I_{COV}, can be minimized by selecting lower values for V_{DD} power supply.

From (4) and (9), the (propagation delay) × (power dissipation) product can be obtained and is given by

$$t_{pd} P_d = \frac{C_L f(V)}{(\beta_L + \beta_D)} I_{DD} V_{DD}. \tag{12}$$

Taking into account all parameters involved in the propagation delay and power dissipation expressions discussed above the projection for the (delay) × (power) product for the c.m.o.s./s.o.s. technology has been made and is shown with a dotted line in Fig. 6(a). The major improvement came from the reduction of channel length from 5–10 μm as presently used to 1 μm and below for devices currently in development. This reduction is considered feasible and is expected to be achieved with current and future technological advances in process control. New process control methods including projection printing, direct stepping of patterns on to a wafer and ion implantation (for control of threshold voltages, for drain and source diffusions and for drain and source extensions), will make it possible.[6–12]

4 Potential Applications of Very-high-speed S.O.S. Integrated Circuits

Improvements in high-speed data processing communications systems are badly needed as communications links now in use are inadequate to handle the large volume of data needed by users.[2] Development of sensors is far out-distancing the capacity of communications links, while command and control centres are becoming potential bottlenecks. Such vast amounts of information can be collected by sensors such as synthetic aperture radars that centres for sorting and displaying of information will be overwhelmed.

Some of the requirements, as called for in NASA publications,[13] include high-speed memories and high-speed data processing units for both on-board and ground-based applications.

The need for very-high-speed on-board data processing is indicated by the growing scope of a current problem—existing satellite systems are currently generating data at the rate of 10^{15} bits/year and processing this data[14] at the rate of 10^{13} bits/year with a projection of a 1986 volume of 2×10^{16} bits/year.

The referenced requirements[13,14] represent only a small sample of the many documented needs for a very-high-speed data processing technology to meet the challenge of the large quantities of data generated by present and future spacecraft.

The converging disciplines of computer technology and telecommunications are linked by developments in high-speed microelectronic component technology. Complex signal processing is made technically and economically possible by high-speed integrated circuit devices. Continued improvement in the parameters of these devices such as speed of operation, price per bit, radiation hardness and the increasing number of components on a chip, is causing a revolution, not only in the aerospace and electronics industries but in most other industries as a whole. The range of applications, many of which will have a significant effect on our society, is extensive.[15] Aerospace applications include on-board signal processing for satellites and high-speed navigation computers for missiles. Commercial applications include high-speed business computers and word processing systems which are likely to lead ultimately to an electronic mailing service.

Many of these applications will be met by very-high-speed s.o.s. integrated circuits.

5 Conclusions

Very-high-speed s.o.s. n-channel integrated circuit m.o.s.f.e.t.s have been fabricated which demonstrated a maximum oscillating frequency of approximately

6·4 GHz. These m.o.s.f.e.t.s—test transistors on a larger integrated circuit—were fabricated with a process which can be used to fabricate high-speed integrated circuits. Very-high-speed performance was achieved using a short-channel device structure, i.e. a gate length of 1·5 μm.

The s.o.s. technology, when implemented using 1 μm n-channel m.o.s.f.e.t.s, is projected to achieve propagation delay times of less than 100 ps with power dissipation of less than 1 mW per gate.

Spacecraft data processing requirements are evolving to indicate the need for very-high-speed devices which are capable of being integrated into large-scale circuits. Improvements in high-speed data processing communications systems are badly needed. Many of these applications will be met in the 1980s by very-high-speed s.o.s. integrated circuits.

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Indexing Terms: Electric filters, active

Some partially active R filter circuits

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SUMMARY

A general circuit configuration is presented from which two bandpass and two lowpass filters can be derived as special cases. These are partially active R filters since each filter has one capacitor and realizes a second-order function by making use of an amplifier pole.

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1 Introduction

The operational amplifier pole was first utilized by Rao and Srinivasan[1] to realize a bandpass filter incorporating one operational amplifier, a single capacitor and a few resistors. Since then the idea of utilizing the gain roll-off of the amplifier as a capacitor has been extended to completely eliminate capacitors in active filter design. The resulting circuits are fully active R filters.

The technique is useful for filters intended for use at relatively high frequencies at which the amplifier pole represents an extension of possibilities. Also, the dependence of filter design on the intrinsic pole of the amplifier, which is due to its imperfection, avoids many accuracy, sensitivity and stability problems[3] which result from neglecting the amplifier gain frequency dependence in active RC filter design.

It is the intention of this paper to fill the gap between the literature on partially active R filters[1,2] and that on fully active R filters. A general circuit is therefore proposed which leads to the circuit of Rao and Srinivasan,[1] and that of Soliman and Fawzy[2] as special cases. Consequently two lowpass filters are also generated as special cases. The performance of all circuits is investigated and different comparisons among them are illustrated.

2 The Network

Consider the circuit of Fig. 1.

$$\frac{V_0}{V_1} = \frac{Z_4}{Z_1+Z_4}\left[\frac{Z_2(Z_3+Z_5)-\dfrac{Z_1Z_3Z_5}{Z_4}}{Z_2Z_5+\dfrac{1}{A}[Z_2(Z_3+Z_5)+Z_3Z_5]}\right] \quad (1)$$

Let

$$A = \frac{A_0}{1+sT_0} = \frac{GB}{s+\dfrac{1}{T_0}}$$

$$T_0 = \frac{2\pi}{\omega_2},$$

where ω_2 is the 3 dB frequency.

$$GB = \frac{A_0}{T_0},$$

where GB is the gain bandwidth product or the unity gain cross-over radian frequency of the operational amplifier. (2)

With

$$Z_1 = Z_2 = Z_3 = Z_4 = R, \quad Z_5 = \frac{1}{sC},$$

as shown in Table 1 (circuit 1), the general circuit of Fig. 1 reduces to that of Rao and Srinivasan.[1] By taking $\frac{1}{2}CR \ll T_0$, the transfer function given by equation (1) reduces to that shown in Table 1.

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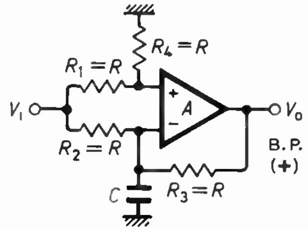
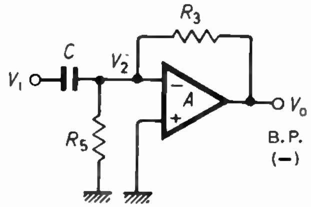
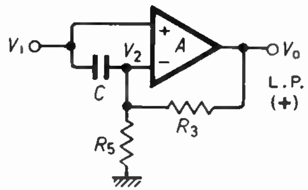
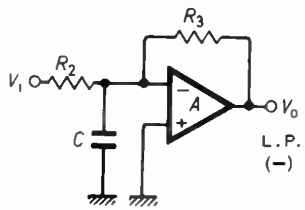
Table 1

Transfer function equations and circuits

Circuit no.	Transfer function	ω_0	Q	$\|G_0\|$ B.P.F.	$\|G_{dc}\|$ L.P.F.	Circuit diagram
1	$\frac{V_0}{V_1}=\frac{sCR.GB/2}{s^2CR+2s+GB}$	$\sqrt{\frac{GB}{CR}}$	$\frac{1}{2}\sqrt{GB.CR}$	$\frac{GB.CR}{4}$	—	
2	$\frac{V_0}{V_1}=-\frac{sCR_3R_5.GB}{s^2CR_3R_5+s(R_3+R_5)+GB.R_5}$	$\sqrt{\frac{GB}{CR_3}}$	$\frac{R_5}{R_3+R_5}\sqrt{GB.CR_3}$	$\frac{GB.CR_3R_5}{R_3+R_5}$	—	
3	$\frac{V_0}{V_1}=\frac{GB.(R_3+R_5)}{s^2CR_3R_5+s(R_3+R_5)+GB.R_5}$	$\sqrt{\frac{GB}{CR_3}}$	$\frac{R_5}{R_3+R_5}\sqrt{GB.CR_3}$	—	$1+\frac{R_3}{R_5}$	
4	$\frac{V_0}{V_1}=-\frac{GB.R_3}{s^2CR_2R_3+s(R_2+R_3)+GB.R_2}$	$\sqrt{\frac{GB}{CR_3}}$	$\frac{R_2}{R_2+R_3}\sqrt{GB.CR_3}$	—	$\frac{R_3}{R_2}$	

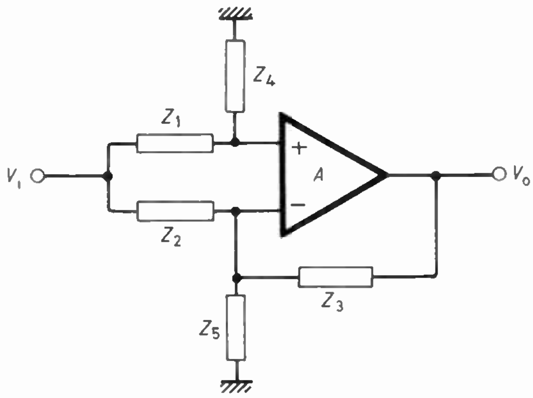


Fig. 1. The general circuit.

With $Z_1 = \frac{1}{sC}$, $\quad Z_2 = \infty$, $\quad Z_3 = R_3$,
$Z_4 = 0$, $\quad Z_5 = R_5$,

as shown in Table 1 (circuit 2), the general circuit of Fig. 1 reduces to that of Soliman and Fawzy.[2] By taking

$$C(R_3/\!/R_5) \ll T_0, \qquad A_0 \gg 1 + \frac{R_3}{R_5},$$

the transfer function given by equation (1) reduces to that in Table 1.

The other two circuits, namely 3 and 4, can also be derived by choosing the appropriate impedances for the general circuit configuration. One should mention that the expressions in Table 1 are valid provided that the following conditions are satisfied for circuits 3 and 4 respectively:

$$C(R_3/\!/R_5) \ll T_0, \qquad A_0 \gg 1 + \frac{R_3}{R_5}$$

$$C(R_2/\!/R_3) \ll T_0, \qquad A_0 \gg 1 + \frac{R_3}{R_2}. \tag{3}$$

It is to be noted that if the above inequalities are satisfied in filter design based on the single pole model described by equation (2), then the simplified transfer functions in Table 1 can alternatively be obtained by directly substituting in equation (1) the approximated single pole model

$$A = \frac{GB}{s}.$$

3 Discussion and Conclusions

Table 1 gives four partially active R filters obtained from the general circuit of Fig. 1.

Circuit 1 is a non-inverting bandpass filter[1] for which:

$$\omega_0 Q = \frac{GB}{2}, \qquad |G_0| = Q^2 \tag{4}$$

which means that for an op-amp with predetermined GB, only one performance factor can be separately specified, while the remaining two factors are dependent.

Circuit 2 is an inverting bandpass filter[2] for which:

$$\omega_0 Q = GB\,\frac{R_5}{R_3+R_5}, \qquad G_0 = \frac{CR_3R_5}{R_3+R_5}\,GB. \tag{5}$$

This circuit has the possibility of being designed with two separately specified performance factors rather than one.

Circuit 3 is a non-inverting lowpass filter, which has the same ω_0 and Q expressions as circuit 2. For a unity d.c. gain (i.e. $R_5 = \infty$) one obtains:

$$\omega_0 Q = GB. \tag{6}$$

Circuit 4 is an inverting lowpass filter having the same ω_0 and Q expressions as circuits 2 and 3 but with R_2 instead of R_5.

For a unity d.c. gain (i.e. $R_2 = R_3$):

$$\omega_0 Q = \frac{GB}{2}. \tag{7}$$

The $\omega_0 Q$ product of circuit 3 is thus twice that of circuit 4 with unity d.c. gain. For a given Q, it is seen that circuit 3 can have a frequency range double that of circuit 4.

Circuit 4 was built in the laboratory using an internally compensated op-amp ($A \simeq GB/s$), type LM 741 (National Semiconductor Corporation), having 935 kHz unity gain cross over frequency at 15 V power supplies. The following 10% tolerance components were used:

$$R_2 = 2{\cdot}4\,\mathrm{k}\Omega, \qquad R_3 = 2{\cdot}4\,\mathrm{k}\Omega, \qquad C = 47\,\mathrm{nF}.$$

In Fig. 2 the experimental plot excellently agrees with the theoretical computer plot. (Calculated from the transfer function equation given in Table 1, with $R_2 = R_3 = 2{\cdot}4\,\mathrm{k}\Omega$, $C = 47\,\mathrm{nF}$ and $GB = 2\pi \times 935$ krad/s.

Concerning the ω_0 and Q classical sensitivities it can be proved that:

$$|S_x^{\omega_0}| \leqslant 0{\cdot}5, \qquad |S_x^Q| \leqslant 1$$

for each circuit, where x stands for any passive parameter (R or C) or any active one (A_0, T_0 or GB).

The op-amp gain bandwidth product (GB) can vary considerably from unit to unit, and can also change considerably with supply voltage variations and environmental conditions. The GB versus supply voltage curve has been studied in the literature.[1] Also, it was experimentally observed that its rate of change has an order of magnitude of 2·5% per volt. The GB versus temperature curve has also been studied in the literature[4] for a sample op-amp, where the rate of change has an order of magnitude of −0·2% per degC.

The resulting variations in ω_0 and Q can thus be large in spite of the low values of sensitivity. For a temperature-compensated filter incorporating an op-amp operating with stabilized supply voltages, for which GB is adjusted for each unit, these sensitivities can be considered to be acceptable.

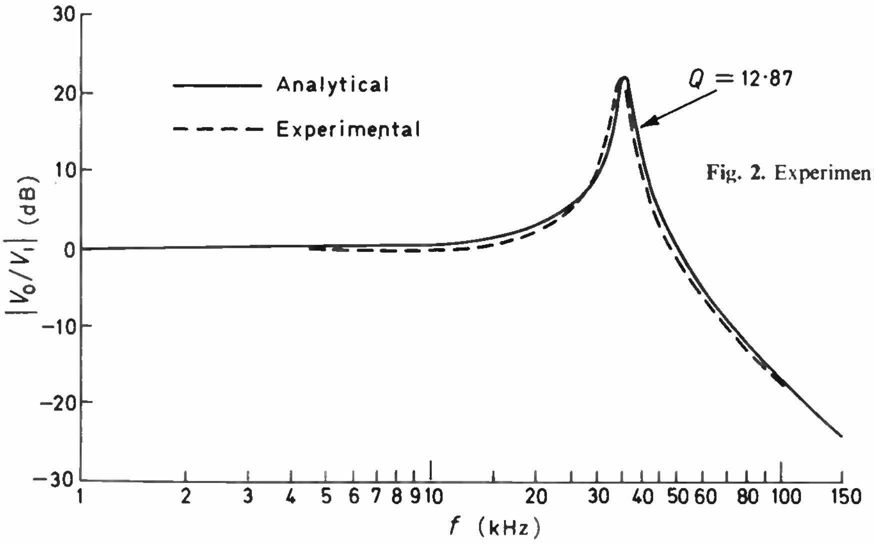


Fig. 2. Experimental and analytical plots of the inverting lowpass filter.

By taking V_2 instead of V_0 as the output, highpass and notch transfer functions can be respectively obtained from circuits 2 and 3. The biquadratic transfer characteristics are thus produced by circuits whose output impedance is not zero. A buffer amplifier designed to have a flat response in the frequency range of interest can be added at the biquad output terminal.

A quick survey of the four circuits derived in this paper reveals that there are two circuits (1, 4) having the capacitor grounded, and two circuits (2, 3) having the capacitor non-grounded. Although some people prefer to use circuits with grounded capacitors, it is possible to construct floating-capacitor circuits having satisfactory performance.

It is also interesting to note that introducing a resistor in series with the capacitor in the previous circuits results in a reduced f_0 and Q. Also, new terms may appear in the transfer function numerator, thus the filter transfer characteristics are no more valid. On the other hand, introducing a resistor in parallel with the capacitor decreases Q while leaving f_0 unchanged. The capacitor can be lossy in this case and hence the circuit is more suitable for integration. Also, by proper choice of the resistors independent control of Q and G_0 is possible.

4 References

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