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NEWS AND COMMENTARY

Sixtieth Anniversary of URSI

IN JULY 1919, the International Research Council was created on the initiative of the Academies of Science. The principal object of the IRC, renamed the International Council of Scientific Unions (ICSU) in 1931, was to bring together the many small and independent groups of scientists, more or less international in character, that had been formed during the 19th century. Some of these groups tended to overlap with each other and the intention was that they should be replaced by, or absorbed into, a few newly-created Scientific Unions, each covering one of the main branches of science.

In 1919, URSI (The International Union for Scientific Radio) and three other Unions (Astronomical, Biological Sciences, and Geodesy and Geophysics) were constituted and by 1923 three additional Unions had been created (Geographical, Pure and Applied Chemistry, and Pure and Applied Physics). There were no further additions until 1947 and there are at present seventeen unions covering virtually all branches of science. ICSU is now supported by UNESCO.

URSI had been preceded by the International Commission on Scientific Wireless Telegraphy which had been formed in Brussels in 1913, mainly on the initiative of Dr Robert Goldschmidt, a Belgian scientist who took a practical interest in radiocommunications from the early years of the century. The formal proposal to create URSI was submitted by the Belgian Delegation to the IRC Assembly in 1919 and Dr. Goldschmidt, who had been Secretary of the 1913 Commission, became the first Secretary General of URSI with General Ferrié as the first President. (The President now is Professor W. N. Christiansen of the University of Sydney.)

To mark the 60th anniversary of URSI, a special two-day Colloquium is being organised on 17th and 18th September under the patronage of the present King of the Belgians, and a number of distinguished scientists, including several Nobel Prizewinners, will present their views on various aspects of telecommunications and the impact of this branch of science on mankind. The Colloquium will be held in the Palace of the Academies in Brussels, the actual building in which URSI was constituted 60 years ago.

Maxwellian Anniversaries

CONTINUING the reminiscent vein above, we would like to remind members that it will be 100 years, on 5th November 1979, since Professor James Clerk Maxwell died. The anniversary of a death is, however, seldom cause for celebration, particularly when the person concerned died at the relatively early age of 48 years and was, moreover, a giant of Victorian science whose further contributions to learning had he achieved the Biblical three score years and ten must surely have been momentous.

So we note the anniversary and look forward rather to 13th June 1981, which will be the 150th anniversary of Maxwell's birth in Edinburgh in 1831. Maxwell's Centenary celebrations took place in London in 1931, a memorial tablet being unveiled in Westminster Abbey by J. J. Thomson, Master of Trinity College, Cambridge, but it is fair to say that most of the running seems to have have come from the physicists, electrical engineers in that year being absorbed in commemorations concerned with the discovery of electromagnetic induction in 1831 by that other genius, Michael Faraday. The relevance of Maxwell's work to radio and electronic engineering is now more widely understood and appreciated and his memory is kept green in this Institution through the Clerk Maxwell Lectures and the annual Premium for the most outstanding paper in the Journal. The 150th anniversary of Maxwell's birth will therefore be fittingly marked in two years' time.

Announcements

Exploitation of Frequencies between 100–1000 GHz

The July and August issues of the Journal will be combined as a Special Issue with the above title and containing nine papers on millimetric and sub-millimetric applications. The issue, which has been assembled for the Papers Committee by Professor D. J. Harris as Guest Editor, will include the following papers:

'Millimetre wavelength impatt sources' J. J. Purcell (Allen Clark Research Centre)

'Waveguides for the 100–1000 GHz region' Prof. D. J. Harris (UWIST)

'Schottky diode receivers for operation in the 100–1000 GHz region' B. J. Clifton (Lincoln Laboratory, MIT)

'Atmospheric propagation in the frequency range 100–1000 GHz range' R. J. Emery and A. M. Zavody (Appleton Laboratory)

'Radar systems for operation at short millimetric wavelengths' S. L. Johnston (US Army Missile Research & Development Command).

'Commercial and scientific applications of mm and sub-mm waves' H. Meinel and B. Rembold (AEG-Telefunken)

'Physical measurements in the 100–1000 GHz frequency range' T. G. Blaney, J. R. Birch, A. E. Costley, R. G. Jones, M. J. Bangham, J. E. Harries and N. W. B. Stone (National Physical Laboratory)

'Space applications and technology in the 100–1000 GHz frequency range' P. F. Clancy (ESTEC)

'Development of model radar systems between 30 and 900 GHz' L. A. Cram and S. C. Woolcock (EMI Electronics).

The July/August issue is scheduled to be published in mid-August. Copies may be purchased from the Institution's Publications Department at a cost of £5 or \$9·50.

Meteosat Up-date

In the short article 'First in-orbit anniversary for *Meteosat*' on page 221 of the May 1979 Journal reference was made to *GEOS East* and *GEOS 1* (2nd column, line 8). These acronyms were in fact given incorrectly: in each case they should have read '*GOES*'. *GOES* stands for Geostationary Operational Environmental Satellite, *GEOS* for Geostationary Earth Orbiting Satellite.

Engineering details of *Meteosat 1* were given in the Journal last year (April 1978, pp 202–3) and included information on the COSMOS Consortium who are building the five satellites scheduled for this important space application that is already helping meteorologists in their difficult task of forecasting World weather.

Index to Volume 48

The Index for the 1978 Volume of *The Radio and Electronic Engineer* (comprising title page and principal contents list, subject index and index of persons) will be sent automatically to all subscribers to the Journal with the July/August issue.

Members who wish to obtain a copy of the Index for binding with their Journals or to keep separately for reference purposes may obtain a copy free of charge on application to the Publications Sales Department, IERE, 99 Gower Street, London WC1E 6AZ, by letter, or by telephoning 01-388 3071.

Indexes are included in all bound volumes of the Journal and members who propose to send their 1978 issues for binding need not apply for a copy of the Index beforehand.

CEI Convention on the EEC

The Council of Engineering Institutions, in conjunction with the Institution of Mechanical Engineers, is convening a meeting of Officers and Officials of major professional engineering organizations within the Member States of the EEC. The expected theme of the convention will be in the context of 'The EEC and Some Implications for the Professional Engineer'.

It is planned that the convention will be addressed by a senior official from the EEC and held on Tuesday and Wednesday, 16th and 17th October 1979, at the Institution of Mechanical Engineers. It is hoped that one of the most important sessions will be open to CEI members and as soon as the programme is agreed a further announcement will be made.

SERT Microtest Symposium Report

Organized by the Society of Electronic and Radio Technicians in association with the Microprocessor Application Group of the Institution of Electrical Engineers and the IERE, the three-day residential Symposium 'Microtest' was held at the University of Sussex from 2nd to 5th April. There were 310 delegates, authors and chairmen present and ten days before the event took place it was necessary to turn away many more who wished to attend.

The Symposium covered all aspects of testing maintenance and reliability of microprocessor-based equipment.

A total of 21 papers was presented in the technical programme and the overall standard was extremely high. Associated with the programme was an exhibition of microprocessors and microprocessor-based equipment.

The Keynote Address was given by Mr Colin Crook, Managing Director of Rank Precision Industries and previously Group Operations Manager for micro-products with Motorola in Austin, Texas. Mr Crook's address ranged widely over many issues connected with microprocessors and set the scene for the papers which were to follow during the ensuing three days.

Copies of all the papers given at the Symposium are available in one bound volume which is available to members of SERT, IEE and IERE for £8 including post and packing; the price to non-members is £10. Orders should be sent to the Publications Dept, SERT, Faraday House, 8-10 Charing Cross Road, London WC2.

Communications Equipment and Systems

Following the successful Communications 78, the Institution of Electrical Engineers is again organizing a major international conference as a part of the Communications 80 Exhibition at the NEC, Birmingham, from 15th-18th April 1980. As in 1978, the IERE is among the supporting institutions, trade associations and government organizations.

The Conference will be held at the Metropole Hotel which is on the NEC site and will cover three distinct themes, public telecommunications, business communications systems and civil radio and emergency communications. Papers will cover engineering, user and operating interests and factors likely to affect overall strategy in each theme area.

Further information can be obtained from: the IEE Conference Department at Savoy Place, London WC2R 0BL.

Members' Appointments

CORPORATE MEMBERS

Professor J. D. E. Beynon, Ph.D. (Fellow 1977) has been appointed Head of the Department of Electrical and Electronic Engineering at the University of Surrey, Guildford, in succession to the late Professor D. R. Chick (Fellow). For the past two years Professor Beynon has occupied the Chair of Electronics in the Department of Physics, Electronics and Electrical Engineering at the University of Wales Institute of Science and Technology, Cardiff.

From 1964 to 1977 he was in the Department of Electronics at Southampton University, latterly as a Reader. In addition to serving as Assistant Dean of the Faculty of Engineering and Applied Science at Southampton, Professor Beynon's academic activities have included the setting up of Micro-electronics Laboratories at the University of Cairo and the Indian Institute of Technology, Delhi.

Brigadier R. H. Borthwick (Fellow 1977) has been appointed to the Board of Bunzl Telecommunication Services; he is currently Managing Director of Automatic Switching, the minority shareholder with Bunzl Pulp & Paper in this information processing company. After retiring from the Army in 1973 as Director of Electronics and Telecommunications, Mobile Area Systems at the Ministry of Defence, Brigadier Borthwick was Head of the Telecommunications Division in the Central Computer Agency of the Civil Service Department.

K. A. Russell (Fellow 1966), a member of the Board of British Relay, has been appointed Technical Director of the Group's newly formed autonomous company, British Relay (Electronics).

G. A. Jacob (Member 1978, Graduate 1966) who has been with the Plessey Company in Liverpool since 1965, latterly as a Design Assurance Engineer I has joined the Mostek Corporation, Carolltown, Texas, as a Standards Engineer in the Memory Systems Division.

N. M. Maslin, Ph.D. (Member 1978) has joined Softwares Sciences, Farnborough, Hants, as a Senior Analyst and Designer. Since leaving Cambridge, where he undertook research in the Cavendish Laboratory for three years after graduating, Dr. Maslin has been a Higher Scientific Officer at the Royal Aircraft Establishment. A paper describing some of his work on h.f. propagation at RAE was published in the October 1978 Journal.

Lt. Cdr. B. R. O'Carroll, RN (Ret.) (Member 1971, Associate 1968) has been appointed Operations Manager based in Riyadh for PCR-Swedtel who are Consultants to the Ministry of Posts, Telegraphs and Telephones of Saudi Arabia on the S.A.I.K. Project (Saudi Arabia Inter-Kingdom). Prior to his retirement from the Royal Navy in 1976 he was in HMS *Collingwood* as a Project Officer.

Lt. Col. B. Reavill, REME (Member 1965), Senior Quality Officer, Army Ground Radar Group, EQD, since 1976, has now been appointed on promotion to be Officer Commanding 33 Central Workshop, REME Newark, Nottinghamshire.

R. G. White (Member 1975), who has been with the Communications Department of the Wales Gas Board since 1971, has joined International Harvester of Great Britain, Doncaster, as Communications Engineer.

NON-CORPORATE MEMBERS

G. V. Tierney, B.Sc. (Associate Member 1977) has taken up an appointment as an Electronics Design Engineer with Shields Instruments, Osbaldwick, York. Following five years as a Research Technician in the Department of Biochemistry at the University of Bristol, he has been for the past year an Academic Specialist in the Department of Physiology and Biophysics at the University of Illinois, Urbana.

P. J. Downs, B.Sc. (Graduate 1972) has joined Thalassa Offshore (Scotland) as an Electronics Engineer. For the past three years Mr. Downs has been studying for his Bachelor's degree in Engineering Science as a mature student at Aberdeen University. His previous appointment was a Technical Officer at the Ministry of Defence.

Lt. Cdr. M. F. Phillips, R.N. (Graduate 1967) has been posted to the Ministry of Defence Procurement Executive) in the Directorate of Airborne Radio where he is concerned with avionics development for the *Sea King* Replacement.

Obituary

Captain Francis James Wylie, O.B.E., Royal Navy (Ret.) (Fellow 1957) died on 5th April 1979 at the age of 82 years following an operation a week earlier. He leaves a widow and two daughters.

'Bill' Wylie retired from a long and active career in the Royal Navy in 1947. He had held several important appointments in Wireless Signalling and had commanded three destroyers and two cruisers. His task as Director of Radio Equipment at the Admiralty from 1944 to 1946 had included the provision of the wartime 3 cm radar Type 268 to numbers of Merchant vessels and he quickly recognized the peacetime value of such devices as aids to collision warning and navigation.

On his retirement from the Royal Navy in 1947, he became Director of the Radio Advisory Service, a small unit established jointly by the Chamber of Shipping of the United Kingdom and the Liverpool Steamship Owners Association, to provide Shipowners collectively and individually with information and advice about the fitting, applications and limitations of marine radar, the newly-available radio-navigation systems and the more recent developments in marine radio and electronics generally. His keenly penetrating mind, clarity of thought and expression and his deep concern that all mariners should be able to derive the maximum benefit from modern electronics resulted in the regular issues from his office in Fenchurch Street of the famous 'RAS Circulars' to ships, which laid the foundations for a great deal of more formal work by his own Organization and others on Marine Radio and Radar matters. In 1952, he edited for the UK Institute of Navigation a book, 'The Use of Radar at Sea', which has become the standard work in the field and which has been revised, updated and reprinted many times—and into several languages. The latest revision was completed only during 1978. He represented both British and International Shipowners at many meetings and conferences, including the International Conference on Radio Aids to Marine Navigation, the (UK) Radio Aids to Marine Navigation Advisory Committee, the Ships Wireless Working Party, and five CCIR meetings between 1949 and 1959.

He was active in promoting the work of Institutes of Navigation not only in the United Kingdom but in several other countries and he was honoured by the Royal Institute of Navigation at home and by sister Institutes in the United States, Holland and France. He was appointed O.B.E. in 1966.

In our Institution he was a founder member of the Radar and Navigational Aids Group Committee, serving from 1960 to 1967, and in 1967 he contributed a paper to the Journal surveying marine radar reliability.

In addition to his enormous efforts in promoting the safe and efficient use of information derived from radar, which continued right up to the time of his death, he was a major contributor to other efforts to improve and develop radio services for the mariner. He was the originator of the re-organization of m.f. radio-beacons in Europe, played a major part in preparing specifications for marine radio and survival craft equipment and in organizing the marine v.h.f. services on a basis of worldwide compatibility of systems and channel useage.

He will be remembered with both affection and respect by all who worked for him and with him, not least for his extreme regard for the use of language—his own precision of thought was reflected in the care of his writing.

'Bill' Wylie himself will be greatly and sadly missed—his influence will be with us for many years to come.

T.W.W.

Applicants for Election and Transfer

THE MEMBERSHIP COMMITTEE at its meetings on 1st and 25th May 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.

May Meeting (Membership Approval List No. 258)

GREAT BRITAIN AND IRELAND

Transfer from Graduate to Member

GALLON, John Bernard. *Frodsham, Cheshire.*
HARDWICK, David Herbert. *Tonbridge, Kent.*
JUNEJA, Surinder Kumar. *Harlow, Essex.*
PATEL, Jay Prakash. *London.*
WEIR, Arthur Alexander. *Egremont, Cumbria.*

NON-CORPORATE MEMBERS

Direct Election to Graduate

FEUCHTWANGER, David Michael. *Cheltenham, Glos.*

Direct Election to Associate Member

KHAN, Safi Ullah. *London.*

Direct Election to Associate

HOLDEN, Francis Edwin John. *Rochester, Kent.*
MORGAN, Gerald. *Edgware, Middlesex.*

Direct Election to Student

CHURCH, Graham Paul. *Prestbury, Cheshire.*
ELLIOTT-POTTER, Steven Edwin. *Bedford.*
FINDLAY, Andrew James. *Andover, Hampshire.*
HILLIER, Michael William. *Norton, Glos.*
PRETTY, Alfred Robert. *Dursley, Glos.*
WEJI, Christopher Chidi. *Plymouth, Devon.*

OVERSEAS

CORPORATE MEMBERS

Direct Election to Fellow

RAKOVICH, Branko. *Belgrade, Yugoslavia.*

Transfer from Graduate to Member

PUA, Qui Thin. *Kuala Lumpur, Malaysia.*

Direct Election to Member

SANDIFORD, Simeon Louis. *St. Joseph, Trinidad.*

NON-CORPORATE MEMBERS

Transfer from Student to Graduate

YEO, Mang Song. *Singapore 5.*

Transfer from Student to Associate Member

CHAN, Cheong Loong. *Seberang Prai, Malaysia.*

Direct Election to Associate Member

LAI, Shui Man. *Seria, Brunei, Malaysia.*
LIM, Ping. *Sibu, Sarawak, Malaysia.*
MOHANKUMAR, Padmanabhan. *Abu Dhabi, U.A.E.*
NANJI, Nasrullah Madatali. *Stamford, U.S.A.*
SI, Yok Fong. *Singapore.*

Transfer from Student to Associate

BORKAR, Anil Vithal. *Bombay, India.*

Direct Election to Student

ACHILLEOS, Michalis. *Limassol, Cyprus.*
CHU, Man Fai. *Kowloon, Hong Kong.*
HO, Kam Wai Philip. *Taikoo Shing, Hong Kong.*
LO, Chi Ming. *Kowloon, Hong Kong.*
NG, Hen Tat. *Singapore.*
TAN, Eng Teck. *Singapore.*
TSE, Chi Ping. *Quarry Bay, Hong Kong.*
VASANTHAROOPAN, Balasingham. *Colombo, Sri Lanka.*

May Meeting (Membership Approval List No. 259)

GREAT BRITAIN AND IRELAND

CORPORATE MEMBERS

Transfer from Member to Fellow

CHAPPEL, George Alfred. *Chalfont-St-Peter, Bucks.*
RAMSHALL, John Denis. *Ottery-St-Mary, Devon.*

Transfer from Graduate to Member

THORN, Richard George. *Plymouth, Devon.*
WEST, Barry John. *Canvey Island, Essex.*
WILSON, Paul Frederick. *Lowestoft, Suffolk.*

Direct election to Member

ALLEN, John William. *Chiswick, London.*
BUBB, Gordon Marshall. *Borough Green, Kent.*
HELBROUGH, Keith. *Great Missenden, Bucks.*
McINDOE, Thomas. *Glasgow.*
NIMMO, George Ross. *Bexhill-on-Sea, East Sussex.*
ROLPH, Barry. *Cambridge.*

OVERSEAS

CORPORATE MEMBERS

Transfer from Graduate to Member

WEERASOORIYA, Don Rubert. *Matugama, Sri Lanka.*

St Fergus Radio Frequency Spark Report Published

There is no likelihood of fire or explosion at the natural gas terminals at St. Fergus, near Peterhead in north-east Scotland, arising from radio transmissions broadcast by the Royal Navy station at nearby Crimond.

A report published by the Health and Safety Executive (HSE) says that radio signals are unlikely to induce levels of power in operational fixed plant structures installed at the British Gas Corporation (BGC) and Total Oil Marine (TOM) sites which could reach even experimentally achieved threshold values at which gas might be ignited by a spark. In recognized 'worst cases' the levels will be sufficiently low to allow a large safety margin.

In addition the report says that insufficient power to produce ignition would be induced by r.f. transmissions in cranes tested at St Fergus although voltage levels could rise above the experimental threshold. It is therefore recommended that the currents and voltages induced in cranes operating on the sites at St Fergus should be investigated when Crimond is transmitting at full capacity.

The report says that tests should be made at the incomplete Shell site as structures are completed, and that Shell have agreed to apply any safeguards that might prove necessary. If any such safeguards are used the tests should be repeated to ensure that they are fully effective.

Separate consideration of possible r.f. ignition hazards will be necessary if further major developments are planned at the BCG and TOM sites. A theoretical assessment, supported if necessary by practical on-site measurements may be required at that time to confirm that no unacceptable radio frequency ignition hazard has arisen.

The report has been prepared by a Steering Committee chaired by HSE and comprising independent experts (Professor D. P. Howson (Fellow) of the University of Bradford, and Mr. G. A. Jackson of ERA), representatives from the Ministry of Defence and the operators at the three gas processing plants. It follows an earlier report made in August 1978 which recommended a programme of further research.

The committee implemented a series of extensive on-site tests, the report says, and this has led to the collection of a wealth of data about the way typical processing plant structures behave electrically when irradiated with radio waves. It seems likely, says the report, that these data will prove to be of general importance in the study of radio frequency ignition hazards. In addition it should lead to refinements being made to the underlying theory of the phenomenon and provide sound guidance about the criteria which should be assumed for how well structures pick up radio waves at various frequencies.

Research commissioned by the committee has also yielded important new results about the powers and voltages required to ignite various common gases when subjected to r.f. electromagnetic waves. It is expected that this work will open the way to a fuller understanding of the phenomenon of r.f. arc ignition of flammable gases.

Study of U.H.F. Land Mobile Radio Systems

The main advantages of quasi-synchronous u.h.f. radio schemes, in which several transmitters at different sites with overlapping service areas are operated on the same channel, are the increase in effective use of the spectrum (a single channel may be used to cover a larger area) and greater convenience in operation because the necessity for channel searching is reduced.

Such schemes have been in use for some years at v.h.f. using amplitude modulation but initial results from u.h.f. f.m. schemes, tried in the USA and elsewhere, have sometimes been disappointing. Despite the increase in demand for mobile radio facilities, the Home Office Directorate of Telecommunications, the body responsible for the specification and purchase of many public service radio facilities, has been reluctant to install such systems until limitations on performance had been determined.

Cambridge Consultants Limited have recently completed a study for the Home Office to assess the performance of quasi-synchronous u.h.f. radio schemes, the objectives of the work done so far being to establish criteria for acceptable reception and to discover what circumstances most affected system performance.

The first phase of the project involved setting up a laboratory simulation of a quasi-synchronous scheme. All parts of the communications link from the operator's console via land lines or fixed point-to-point radio links to the base station transmitters and ultimately to the receiver loudspeaker were subjected to objective and subjective measurements. Limitations on system performance have been identified and satisfactory solutions to technical problems have been recommended.

Home Office trials are now .being run to validate the laboratory results. The next phase of the work will attempt to find the optimum methods of measurement and equalization of base band links in a practical quasi-synchronous scheme, to ensure satisfactory system performance.

Medical Application of Night Vision Aid

A clinical trial is being carried out in the United Kingdom of a device that may be used to help people suffering from night blindness caused by retinitis pigmentosa. This project has arisen from the demonstration in a BBC Television programme 'Tomorrow's World' of an image intensifier night vision aid developed by Standard Telecommunication Laboratories. STL was approached by the British Retinitis Pigmentosa Society, which wanted to know more about the aid and as a result, an ITT night vision aid has been presented to the Society on indefinite loan. The Society has passed the device—which resembles a monocular opera glass and works by amplifying existing light—to the low vision clinic of a London eye hospital. The device neither cures retinitis pigmentosa nor arrests its progress. However, it does permit sufferers to use their best daylight vision in darkness and can help a significant number of the approximately three million people worldwide who have this genetic disease.

This type of night vision aid is, in effect, an electronic analogue to the eye. Existing light falling on a photocathode generates an electronic signal which the device then amplifies in a micro-channel plate, a glass disk containing about 1 million channels, each 4 microns in diameter. The electronic multiplication that occurs in the channels is analogous to what occurs in the optic nerve. The amplified electronic signal is converted to an inverted light image by a phosphor screen and this image is reinverted by a fibre optic mechanism and focused onto the user's eye by a lens. He sees a green image because of the colour of the phosphor screen. The device also compensates for bright flashes of light, just as the human eye does, by immediately cutting light intensity to protect the eye.

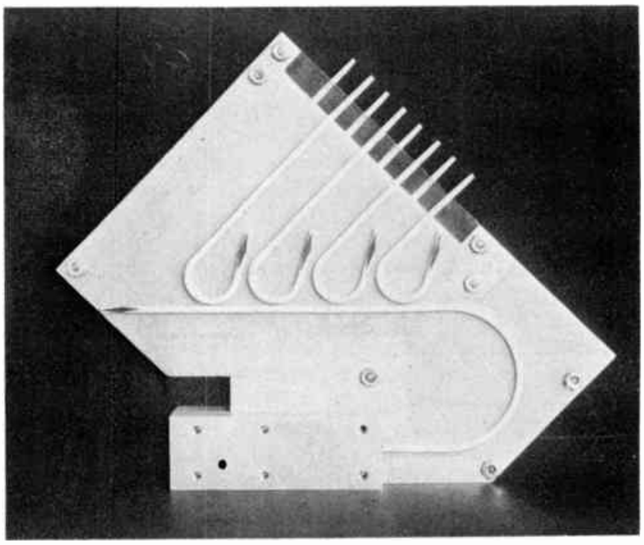
In an entirely different application ITT's night vision aid, originally developed for military applications, has recently completed a successful two-year trial with the Swiss Helicopter Alpine Rescue Service. Using the device, the Service's helicopter pilots have made night landings on glacier's in very bad weather conditions without even starlight illumination.

ERA to Collaborate with IITRI

The Electrical Research Association Ltd. has signed a new four-year agreement with the Illinois Institute of Technology Research Institute (IITRI), which includes the formation of a Joint Ventures Advisory Committee with a mandate to identify and actively to pursue new technical and business opportunities for the two organizations. This agreement follows the successful growth of the Radio Frequency Technology Centre at ERA from its establishment in 1974 as part of the international collaboration between ERA and IITRI. Since that time the RFTC business has grown rapidly at an annual rate of 90%.

The rapid growth of the RFTC reflects the increasing need for more sophisticated antenna and radio frequency component design in a wide range of applications due to the increasing demands for higher performance in an overcrowded spectrum. Current work includes the research, design and development of high performance antennas and components for communications, radar and space applications; the development and application of accurate measurement techniques at frequencies ranging from below 1 GHz to greater than 60 GHz; the production of a complete prototype 60 GHz man-portable communication-link system; the application of modern mathematical methods to the analysis of radio frequency devices; and the development of software for computer-aided design applications.

RFTC's specialization in research into new concepts of antenna design is actively supported by the European Space Agency and computers are used extensively in all stages of the work.



Novel form of beam-steered antenna. The 30 GHz frequency-scanned dieletric image-line array developed at the Technology Centre of ERA.

Automatic data processing for new Channel information centre

The Department of Trade has placed an order with Decca Radar for a very advanced Automatic Data Processing system. The first of its type in the UK, this is for incorporation in the new operations centre of the Channel Navigation Information Service (CNIS) now under construction at Langdon Bay.

The CNIS radar system currently consists of a Decca HR 25 (25 ft scanner) on the Coastguard station at St. Margaret's Bay and a similar but unmanned radar on the nuclear power station at Dungeness, both feeding into special, very advanced displays at the Coastguard Station. When the new site—some two miles to the west—is completed, the radar scanners will continue in their present position but other radar functions will be transferred to Langdon Bay, the radar data being transmitted by a combination of microwave link and landline.

A study carried out by the D.o.T. Operational Research Unit and Admiralty Surface Weapons Establishment confirmed the need for an ADP system if the new look CNIS was to operate efficiently, remembering H.M. Coastguard's multiple responsibilities. Tenders were invited from a number of manufacturers, and following demonstration of a suitable tracking system that from Decca was accepted.

In the operations centre there will be four 16 in. displays, a 23 in. Area Radar Plot display, six 12 in. rotating coil p.p.i.s for time lapsed recording purposes, and twelve alpha-numeric video display units with keyboards. Three of the 16 in. displays are for radar operators, the fourth for the Officer of the Watch, who will also have the Area Radar Plot 'conference' display for overall decision making purposes. The Automatic Data Processing system as a whole will accept information from many sources including radar data from the two remote stations. From this it will continuously track selected targets. Information may also be entered manually or from other automatic services. Facilities are provided for the operators to recall, amend or co-relate this data as required, displaying it either in alpha-numeric form on their v.d.u.s or in synthetic graphic form on the Area Radar Plot.

All these and other tasks will be performed by two integrated sub-systems: the autotrack system and the information system (each duplicated for reliability). The autotrack system will automatically track up to 250 targets, acquired either manually or, in a limited area, automatically. Following acquisition of each one, it is continually tracked; positions, courses and speed of selected targets are computed from immediate past history and either the vectors or the past history or both displayed if desired; and many subsidiary functions, such as the bearing and distance of one ship from another, are made immediately available.

The information system is a data storage and processing system, which, using two mini-computers, feeds the main radar displays (through the autotrack system), the Area Radar Plot display, the v.d.u.s, a paper plotter, a line printer and a magnetic tape unit. Examples of interplay between the two systems are the display of data identifying a 'rogue' ship (that has transgressed the traffic separation scheme rules) as received from the investigating aircraft; and the important procedural reporting by certain ships as they move through the scheme. When a ship arrives within radar coverage her details, stored in the information system, are related to her tracks as stored in the autotrack system, enabling her echo to be positively identified.

Among other inputs to the information system there are the important ones from v.h.f. d.f. stations at South Foreland and Fairlight, invaluable both in identifying echoes and in search and rescue operations. Both very advanced, the 16 in. displays and the autotrack system have been under evaluation at the St. Margaret's Bay centre since 1976, being latterly on loan to H.M. Coastguard and in extensive operational use there. The former are fixed coil displays that provide a very bright picture by means of recycling and restoring the video. The autotracking system is a microprocessor system of which the main features are the ability to deal with multiple echo situations and to carry out reliable tracking in clutter conditions; execution of the latter is based on Decca's highly successful Clearscan device. The contract calls for the new Automatic Data Processing system to be operational at Langdon Bay by mid-1980.

Standard Frequency Transmissions

(Communication from the National Physical Laboratory).

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

April 1979	MSF 60 kHz	GBR 16 kHz	Droitwich 200 kHz
1	0·3	3·6	12·8
2	0·1	3·4	13·0
3	0·1	2·8	13·3
4	—0·7	3·2	13·5
5	—0·3	2·9	13·8
6	—0·5	3·0	14·1
7	—0·6	2·9	14·3
8	—0·6	3·0	14·6
9	—0·6	2·7	14·8
10	—0·5	2·5	15·1
11	—0·6	5·0	15·4
12	—0·6	6·0	15·7
13	—0·6	5·0	15·9
14	—0·5	5·8	16·1
15	—0·1	4·7	16·3
16	—0·4	4·8	16·5
17	—0·4	4·6	16·6
18	—0·4	4·9	16·9
19	—0·4	4·9	17·1
20	—0·4	5·0	17·3
21	—0·5	4·9	17·5
22	—0·3	4·6	17·8
23	—0·3	4·9	18·1
24	—0·3	4·8	18·4
25	—0·3	4·6	18·7
26	—0·5	4·6	19·0
27	—0·6	4·7	19·2
28	—0·5	4·5	19·4
29	—0·4	4·5	19·6
30	—0·4	4·6	19·8

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.

General Interest Article

Instrument technology of the future

REGINALD S. MEDLOCK, C.Eng., F.I.Mech.E., F.I.E.E.

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Possible areas of advancement for instrument technology of the future are examined which are aimed specifically at improving the efficiency and safety of the process industries (including energy generation and utilities) through measurement and control. Such advances would revolve largely around the rapidly developing area of microelectronics and its expanding application.

The process industries have a tradition of conservatism towards the adoption of new products whose performance, reliability and back-up resources are still to be proved. Furthermore, through replacement, retrofit and user familiarity, the life of older products is extended, sometimes to obsolescence, thereby retarding the acceptance of newer developments. The effect of this is that the prophets of technology tend to overestimate progress in the first five years but underestimate in the following 20 years.

Microelectronics

There is, however, a unanimity of opinion amongst forecasters that the instrument industry will be dominated by microelectronics for the foreseeable future. In this article microelectronics covers microprocessors, memories, silicon chip devices, integrated circuits, l.s.i. (large scale integration) and v.l.s.i. (very large scale integration). (The news media frequently refer to microprocessors when they really mean microcomputers or even the broader field of microelectronics.)

The precise boundary between microprocessors and microcomputers is difficult to establish as it depends on what has been included on the microprocessor chip. Normally a microprocessor is a component and a microcomputer is an assembled kit on a printed circuit board comprising one or more microprocessor units and additional components such as memories.

The growth of computers in industrial control was slow to start but has begun to reach explosive proportions. It has been estimated that, since the first computer control applications in 1960, half a million central processors have been installed (not just on process control of course). The growth rate has averaged 67% per annum and it is believed that 240,000 processors were added in 1978 and 340,000 will be added in 1979.

Microcomputers look as if they will invade all industry as well as the domestic market. Retail shops in this country and the USA are now offering the public microcomputer hobby kits for as low as £40 and microprocessors are being incorporated in domestic equipment and motor cars. A chess set with a microcomputer able to play the game against a human opponent, is available at around £200 and a 'Speak and Spell' toy costing about $50 in the USA has been developed using a microprocessor chip to synthesize human speech. Although some of these examples are not directly relevant to instrumentation it can be inferred that huge domestic demand will bring cost reduction benefits to microcomputer hardware which will have

Table 1.
Applications of microelectronics in measurement and control

Transmission:	Multiplex signals from transducers to simplify cabling. Control interrogation of transducers and data buses.
Data Presentation:	Data processing. Control of v.d.u.s and other displays.
Alarms:	Control of priorities. Provide alarm logic. Control alarm devices.
Control:	Incorporation in discrete digital controllers. Incorporation in distributed control systems. Provide automatic tuning of controllers. Flexible programmable controllers. Diagnostics. Signal conditioners.
Actuators:	Control speed, position and safety routines. Variable characterization and loop gain adjustment.
Transducers:	Self-calibration and fault diagnosis. Adjustment for drift. Linearization. Improvement in accuracy and repeatability by eliminating systematic errors. Data processing of data from two or more transducers to compute, for example, mass flow or flow measurement by cross-correlation. Analogue to digital transmission of data and vice versa. Control of analysers particularly mass spectrometers and gas chromatographs.

Reg Medlock has spent his whole working life to date with George Kent Ltd. In August 1977, when he retired from the post of Group Technical Director, he was appointed Consultant to the Company. His experience in the instrumentation and process control field has extended over a period of 45 years and he has witnessed the evolution of measurement and control technology since its surge forward in the early thirties. He has been closely associated with the Institute of Measurement and Control since its inception when it was known as the Society of Instrument Technology. He is presently serving on the Department of Industry's Advisory Council on Calibration and Measurement as Deputy Chairman and has recently retired as a Director of the SIRA Institute.

a tremendous influence on instrument technology. Even today a microprocessor chip can be bought for only £3. The UK Government, having realized the potential impact of microprocessors on industry, have offered large sums of taxpayers' money to encourage their manufacture and application.

Microelectronic applications

Microcircuitry including microprocessors and other microcircuit chips will be applied to instruments in many ways. Table 1 shows a number of examples.

In the excitement of microprocessor application it is sometimes forgotten that usage will be limited by problems associated with interfacing, software production and mechanisms in the same way that the human capability can be less effective when the brain is not coupled to five good senses and healthy muscular action.

At the present stage of development the applications likely to progress fastest are those which are simple enough to have proprietary application programs built into the chips via the final masking in their manufacture. On economic grounds, this implies at least 500 identical chips being required.

The software problems, that is the cost, the manpower shortage, and the lack of standardization, will dominate the microprocessor scene for many years and will control growth. Some relief will be obtained by the availability of more powerful computing facilities, bigger and cheaper memories and the adoption of hardware in place of software. Magnetic bubble memories which are non-volatile (suffer no change with power loss) are becoming available. It has been forecast that by 1982 a bubble memory capable of storing one million bits and having a one microsecond response time will be available on a single chip 25 mm square and will cost about £50. Although the greatest excitement has been generated by the advent of the microprocessor, nevertheless the technical achievement of cheap large-scale memories deserves equal attention.

It is fairly certain that microcomputers will be oversold and misapplied and give rise to some disillusionment in common with every new technical tide. This disenchantment will not be due to lack of reliability or performance *per se*, but to misapplication, overselling, inadequate memory, inadequate word length, rapid obsolescence and shoddy software, all born from hasty optimism and the desire to make a quick return. It has to be remembered that a microcomputer is relatively more difficult to program than a minicomputer; also a microcomputer is only as capable as an ordinary main frame computer in performing tasks. The revolution brought about by the microprocessor is that it can perform these tasks more cheaply and flexibly.

In a recent Frost and Sullivan report it was estimated that by 1981 the European process control market will require £20 million worth of microcomputers and this figure will rise to £100 million by 1986.

Controllers

Pneumatic controllers

Pneumatic controllers will continue for some time to surprise many electronics experts who cannot understand why they are not dead and buried. These experts overlook three factors:

A vast amount (ranging from 40% to 75%) of instrument companies' orders is made up of replacement, retrofit and user familiarity business, which has prolonged the life of pneumatic products.

Pneumatics have had some design boosts from third-generation instruments.

Pneumatics and electronics can interface fairly well to get the best of both worlds.

Further pneumatic developments should be possible allowing lower production costs and improved specification, but it is unlikely that any manufacturer will seize the opportunity to move into a non-competitive field. Most manufacturers will prefer to go with the scores of competitors into the more stimulating field of the microprocessor.

Programmable controllers

One of the largest growth areas in control is the programmable controller. World sales of these controllers are estimated to be currently £100 million. They are applied 60% to manufacturing and 40% to process industries and for the last two years have shown a growth rate of 100% per annum. The basics of a programmable controller can now be purchased very cheaply as a single silicon chip, so a further escalation of development and sale can be confidently predicted.

Three-term analogue electronic controllers will be required in economic production quantities for about seven years. but it would be hazardous to forecast a longer life now that a microprocessor single loop controller could possibly be produced at a cost comparable with the analogue version. In any case the single loop controller will be restricted to small plants and for back-up purposes until the cost of alternatives can be reduced sufficiently for manufacturers to employ a high degree of redundancy. Indeed, a major advance expected in the future for all microcomputer-based equipment is the improvement in reliability through redundancy and self diagnosis. It seems probable that manufacturers will take advantage of the continuously reducing prices of electronic components to improve performance and reliability for the same product price rather than maintain previous standards at a lower price.

Primary Measurement

Flow

Flow measurement has enjoyed two decades of major development and there is now a wide choice of flowmeters for most applications. It is unlikely that a radically new technique of industrial flow measurement will appear in the next ten years. But improvement of existing types is badly needed including the application of electronics to improve accuracy standards. Table 2 gives some of the likely areas of improvement.

Pressure

Developments in pressure measurement seem destined for the next decade to be related to diffused or iron-implanted silicon diaphragms with or without integral amplifiers. Longer term, optical measurement of diaphragm movement by optical fibres could come about (see Transducers).

Temperature

No revolutionary changes can be expected in temperature sensing materials but some minor developments will occur involving the use of semiconductor materials as temperature sensors. Fibre optics pyrometers have recently been developed and will play a useful role in certain difficult applications of temperature measurement.

Level

Ultrasonic techniques have a good potential for this type of measurement for liquids and solids and will probably show a higher growth rate than other techniques.

Analytical

A great deal could happen in this field but it is unlikely that there will be any single large-scale development. Infrared, gas chromatography and mass spectrometry will benefit from development expenditure. New developments can be expected in specific ion measurement in liquids involving selective membranes; for gases, semiconductor sensors could make a useful contribution but their performance to date has not been very encouraging.

Transducers

In the last five years the market has swung away from force-balance designs and has welcomed micro-displacement types of which the two most important contenders are those of the silicon strain gauge and the variable capacitance system is well established and is likely to persist in the market for at least a decade. However, the silicon strain gauge approach is attractive and is linked to a rapidly advancing field of technology, at least as far as the silicon chip is concerned.

In the interest of plant safety, there will be a demand for 'passive' transducers, that is, transducers consuming no power *in situ*. Interest is turning to fibre optics to solve this problem. The basic idea is simple; to use light conducted down a fibre to measure the strain or displacement of a transducer diaphragm. The light would then be modulated and returned to the control room for interpretation. For the moment the ideal light modulator for this application does not exist but when a solution is found it will offer an exciting opportunity to the manufacturer.

Table 2. Areas of likely improvement in flowmeters

Pressure Difference Devices:

Research being undertaken in various parts of the world should refine available knowledge on coefficient data and on the effects of asymmetrical flow profiles. Pressure differences will be measured more accurately by transducers (q.v.).

Electromagnetic:

Direct current pulse excitation in various forms will continue to be developed and refined, but alternating current excitation could be improved to compete with the newer systems being offered. Improved electronic circuitry will permit the application of these meters to non-conducting fluids.

Vortex:

This still promises to be the meter of the future but both electronics and vortex detection must be improved. Other types of flowmeters based on other forms of hydrodynamic instability will appear on the market but, because they rely on similar principles to that of the vortex meter, will experience similar limitations in accuracy and performance.

Cross-correlation:

One commercial meter has appeared on the US market. Several more will appear in the next five years for application to some of the more difficult two-phase fluids. The correlator will almost certainly be microprocessor-based.

Turbine:

This type of flowmeter still has a long life ahead of it but development work will be aimed at refinement rather than at fundamental change. Improved pick-offs could reduce cost and improve performance.

Mass Flow:

Microprocessors will increasingly be used in the various systems of mass flow measurement both for calculating corrections for a range of variables and for controlling the input of data.

Ultrasonic:

This type of meter has not enjoyed a high reputation in the past although it has been offered in the market for 20 years or more. It is possible that it could emerge as a main competitor to most other types.

Telemetry

Economic, efficient and reliable telemetry will be one of the main contributors to modern developments in the integrated control of process plants.

In the past, measurement and control instrumentation was necessarily distributed around the plant. As improved short-distance transmission signals evolved, control became centralized in one control room. Recently the growth in size and complexity of plants and the necessity for a more hierarchical organization of information and control is again leading to a distributed control system, but this is not a complete reversal. Modern systems have a facility for transmission of data which could scarcely be envisaged thirty years ago enabling a wide distribution of equipment to be made without loss of essential information at any point. Transfer of large amounts of data can only be carried out economically in a digital mode. The transfer medium can be multicore cable, coaxial cable, microwave or optical fibre. Technological developments of the last two will have an increasing impact on process control.

Displays

The increasing amounts of data required by some modern processes have made the instrument panels of ten years ago quite impracticable. Various alternatives have been introduced to overcome the difficulty, the most successful being the v.d.u. in the form of a black and white or colour television-type screen. In general terms, one v.d.u. can replace 15 metres of conventional panel. The v.d.u. and its associated controls become the interface between the operator and the process. There will be constant pressure from the user to improve this interface still further, particularly as far as the display feature is concerned. New devices employing plasma discharge panels are coming on to the market which, though simpler, less bulky and probably cheaper than the television type of v.d.u., are no more pleasant to look at. A more attractive possibility could be projection type liquid crystal displays but these are a long way off. Nevertheless, for other instrument displays, liquid crystals could be the main challenger and the instrument industry could reap the benefit of an increasing domestic demand for cheap displays for watches and calculators.

Other as yet unidentified physical/electrical phenomena will almost certainly be introduced in the next ten years.

Ultrasonics

Now the ultrasonic transducers and associated electronics are well developed and available at economic prices, the application of ultrasonic techniques is likely to accelerate in the instrument industry. Apart from the known uses of these techniques in flow, chemical concentration, level and electrode cleaning, there could be many other applications such as the accurate measurement of displacement, consistency, gas analysis and temperature.

Robots

To many people, the term robots conjures up thoughts of science fiction, but there are over 3000 robots operating in the USA and probably as many in Japan. The UK is considered to be ten years behind Japan in 'robotics' and the Government has now started to become concerned and is despatching a study team to Japan. The subject is outside the scope of this article as the current use of robots is for physical jobs involving health hazards and repetitive tasks—welding being a typical example. Nevertheless, robots are mentioned because ultimately, their development will interact with the requirements of process control.

Conclusion

In general terms the development of process control instrumentation has worked towards the long-term goal of equalling or surpassing the capabilities of the human being in terms of the senses (hearing, sight, touch, taste, smell), calculation, memory and logic (the brain) and in physical actuation (the muscles). In spite of the tremendous advances of science the gaps between human capability and instrument capability are enormous in some areas, and it is these gaps which provide the pointers to the direction of future developments.

For example, pattern recognition by computer is very weak by human standards; chemical analysis is cumbersome and slow compared with the speed and sensitivity of taste and smell. Photo-electric and optical systems do not have the dynamic range of the eye, and microphones do not have the dynamic range and noise discrimination of the ear (10^{30} to 1).

As far as the brain is concerned computers have a long way to go; the brain normally has ten thousand million neurons for storage of information with an access time of one second (sometimes a little longer). Each neuron can communicate internally with five thousand other neurons providing facilities for an almost infinite variety of memory patterns. The bubble memory of 1982 will be able to store one million bits of information (only accessible externally) in a 25 mm square chip. For comparison, two thousand of these bubble memories would be needed to store the contents of the 'Encyclopedia Britannica'. The brain also has built into it a vast amount of redundancy in order to ensure that false signals and noise do not destroy the value of the information received. Thus, there is a long way to go to bring computers up to the memory capacity of the human brain.

Some time in the next fifty years there could be a breakthrough in reducing some of the large gaps between human capabilities and those of instruments but the probability is that it will involve the development engineer in a further science, such as biology or physiochemistry, for there is no doubt that biological sensors have a specificity and sensitivity greatly exceeding those made of metal, plastic and silicon.

Forthcoming Conferences

International Conference on Progress in Postal Engineering

The Chairman of the Post Office, Sir William Barlow, will open the three-day 'Progress in Postal Engineering' international conference which is being held in London during November. It takes place on 6th to 8th November 1979 at the Headquarters of the Institution of Mechanical Engineers, which is sponsoring the event with the support of the Institution of Electrical Engineers, the Institution of Electronic and Radio Engineers and the Institution of Post Office Electrical Engineers.

The last such conference was held in 1970 and there will be a review of postal engineering developments in recent years. Future prospects for mechanization and trends for the 1980s will be examined.

Topics being covered include the sorting offices in Oslo, Liverpool and Denmark. Other papers to be presented include: Examination of post offices and sorting procedures in Oslo, Copenhagen, and Helsinki; Parcel and packets soration in Canada; Voice data systems in Tokyo; Performance monitoring in mechanized letter offices and post code and sorting translation systems in the United Kingdom; and Chutes and chain conveyor systems. The electronic mail services of GEC Mechanical Handling, of Erith, Kent, will be featured. West Germany's coding of mail; use of video equipment, as well as an experimental closed circuit television system in the United Kingdom, are the subjects of three further reports.

On the day after the conference delegates will be guests of the Post Office for a trip to the North West of England so that a detailed study can be made of the Liverpool Head Sorting Office.

The Manipulative and Mechanical Handling Machinery Group of the Institution of Mechanical Engineers is organizing the conference and registration forms can be obtained from the Conference Department, The Institution of Mechanical Engineers, 1 Birdcage Walk, London SW1H 9JJ.

Video Rights 79 Conference

The explosion of activity in pre-recorded video cassettes and disks has left innumerable commercial and legal loose ends. Most existing films and broadcasts never carried clauses covering rights for video distribution—because video did not exist at the consumer level.

A conference to examine the solution to the rights tangle —Video Rights 79— is being held at the Cafe Royal, London, on 26th and 27th November 1979.

The UK rights situation will be examined—with emphasis on any changes the new British Government is likely to introduce under the pressure of the arrival of video, and following the 1977 Whitford Report. Speakers will also deliver papers on the rights situation in other European territories. Full post-conference publication of the papers delivered at Video Rights 79 is planned.

Further details on Video Rights 79 may be obtained from Miss Agneta Moe, Nord Media Ltd., 37 New Bond Street, London W1Y 9HB (Tel 01-629 9381).

General Interest Paper

Satellite Communications in the Royal Navy

R. G. STILL, C.Eng., M.I.E.R.E., F.I.T.E.

The United Kingdom has interests that are world wide, and in order to protect those interests it is necessary that the Royal Navy have reliable and secure communication coverage. There is also an increasing need for central control of strategic operations. It was seen that satellite communications would satisfy these demands.

Based on a contribution to the IEE Colloquium on Small Earth Terminals for Satellite Communications held in London on 6th March 1979.

Consideration of the types of satellite systems currently available indicated that a geostationary system of satellites would be more stable in operation than the smaller, less expensive and expendable Earth-circling satellites.

In order to provide the Navy with the operational large bandwidths its applications required, consistent with reliable propagation, the 7/8 GHz frequency region which had been allocated by international agreement for the exclusive use of the military was chosen. This frequency region is in a quiet area of galactic noise level. The Earth to satellite link is at the upper end of this band and the satellite to earth link at the low end of the band.

The development of the satellite communication system known as *Skynet* began in 1966 and 1967 the Admiralty Surface Weapons Establishment (ASWE) had designed and developed a naval experimental ship terminal which successfully completed trials on a Royal Navy vessel.

The first operational satellite terminals developed were fitted to HMS *Intrepid* and HMS *Fearless* in 1970 and 1971 and became known as *Skynet V* terminals. These terminals were described at an IEE conference on Earth Station Technology during 1970, but a brief resumé may be helpful. The system used frequency modulation with frequency division multiple access in the frequency band of 7/8 GHz with a maximum transmitting power of 5 kW c.w. In each access the baseband facilities for each ship of telegraph and speech were multiplexed together. The speech channel would be either digital (secure) or analogue (plain language). The aerial was 1·8 m in diameter and stabilized by an internal gyro unit. Panniers mounted on the sides carried receiver pre-amplifiers and the transmitting klystron. The small aerial size of the shipborne terminals demands that a larger portion of the radiated satellite power is allocated to the ship's use than to the larger landbased terminals, and a separate narrow frequency band within the 7/8 GHz of the satellite has been allocated exclusively to Royal Navy ships. This is true for the United Kingdom satellite *Skynet* but not for American or NATO military satellites.

One disadvantage of the *Skynet V* terminal design was the single aerial head—see Fig. 1. The project team responsible for design would like to regard the ship as a floating platform for the convenience of the terminal and site the aerial with an all-round, unobstructed view of the sky! The Navy having the need to deploy other systems cannot give to the satellite terminal the priority to permit its siting on a good clear site.

It is obvious then that the siting of the aerial cannot be in the best position for communications and there are parts of the ship that obstruct the line of sight between the aerial and satellite depending on the position and course of the ship.

Fig. 1. *Skynet V* **aerial head fitted on RN ship.**

Raymond Still (Member 1973) served an apprenticeship in marine electrical engineering from 1942 to 1948 and during this period obtained technical qualifications as a result of part-time studies at Southampton University College. After working for four years with Vosper Ltd., he entered the Royal Naval Scientific Service in 1952 and since then has been with the Admiralty Surface Weapons Establishment; he is now a Senior Scientific Officer. Mr. Still has been concerned with research and development in a variety of radar, electronic tube, propagation and communications projects and he is currently engaged on the development of shipborne satellite communications terminals.

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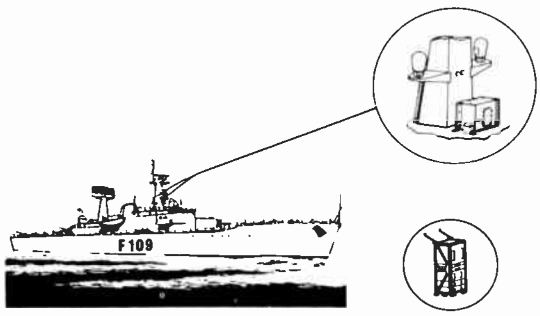

Fig. 2. SCOT terminal equipment.

This is known colloquially as 'wooding'. Another important consideration is to ensure for safety reasons for personnel that the field strength from the terminal should not exceed 100 W/m² at levels below 2 m above deck

The two original Royal Naval terminals operated within the *Skynet* networks terminating via the satellite at a Master Engineering Control Centre (MECC) in Hampshire. Land lines connect the MECC to a secure exchange in London where connections to the various speech subscribers are made. The telegraph transmissions are also conveyed by land line to the message handling communications centre in London. The MECC can be considered transparent to secure messages but is able to receive and transmit plain language speech and unencoded telegraphy. These facilities are useful for the engineering of the system, enabling power balancing disciplines etc. to take place.

The introduction of satellite communications by means of these terminals into the Royal Navy facilitated reliable communications over 24 hours of each day, thus greatly increasing the message handling capacity of the fitted ships. Signals were transmitted and received without delays normally associated with long-haul h.f. communications; further there were no errors present in the messages so repeats were not required. The immediate role of these satellite communication fitted ships became that of a 'gateway'; all long-haul communications to an area served by many ships would go to the 'gateway' ship in that area where they could be distributed by v.h.f. or h.f. groundwave.

By means of these 'gateway' ships Britain was able to help in the flood relief of Pakistan in 1971 to an extent not previously realized. Also as the role of our forces in the Far and Middle East was reduced during 1972/73 the 'gateway' ships became the base headquarters of the Brigadier commanding the Land Forces in that area. Although *Skynet V* terminals were development models they are still operating but will become obsolete by 1980 after 10 years of continuous service with very little down-time. However, to achieve this result, dedicated efforts were required of the ship's staff in maintaining the equipment, sometimes under very difficult conditions.

The *Skynet V* terminals are heavy and expensive equipments not suitable for fitting in small ships. The requirements for a small lightweight system and the alleviation of the 'wooding' problem is met in SCOT which is a transportable terminal comprising four units: a port and starboard aerial unit, an engineering cabin and a control console which is fitted in the ship's main communications office (Figs. 2 and 3).

The SCOT Naval Satellite Communications System is shown schematically in Fig. 4. A variable number of SCOT fitted ships are deployed in communication with the Master Engineering Control Centre (MECC) through the satellite transponder. Because of the two-aerial configurations of SCOT, the 'wooding' problem for reception is small. This means that all ships can receive a continuous Fleet Broadcast with a high degree of confidence in continuity. The Broadcast consists of two multiplexed components:

the operational broadcast, and

an engineering broadcast originated by the MECC.

In the normal run of events each ship will transmit a ship/shore reply only when it has traffic to clear. This it will do by acquiring one of the accesses through the satellite and holding it only as long as is necessary to establish confidence in the link and pass traffic.

Each ship/shore reply similarly has two multiplexed components:

operational traffic, and

engineering traffic terminating in the MECC.

The engineering broadcast is used by MECC to address the subscriber ships either individually or collectively. Apart from the setting to work of ship terminals as they enter the system and the resolution of crises, the load on this service should be light since the SCOT terminals are not required to power balance their ship/shore reply transmissions on a short term basis. It is an aim of the current system design to minimize operator involvement at MECC. The engineering broadcast can also be used to transmit satellite service information which is usually in tabular form such as satellite acquisition data. Obviously it would be uneconomic to provide a dedicated access through the system for the ship/shore replies from each SCOT subscriber; this applies both to the multiple access scheme through the satellite transponder and the provisioning of ground equipment. So, in a fully deployed SCOT system subscribers will greatly outnumber accesses. This poses a multiple access problem and it follows that a sub-system is needed to exercise central control.

A related problem is that of giving the ship's operator confidence that his ship/shore message has been successfully received in UK. This would imply some form of receipting system.

A short term solution to both these problems has been the installation at the MECC of a Channel Availability and Receipting Broadcast (CARB) equipment designed by ASWE. In each ship/shore reply channel is a receiver demodulator

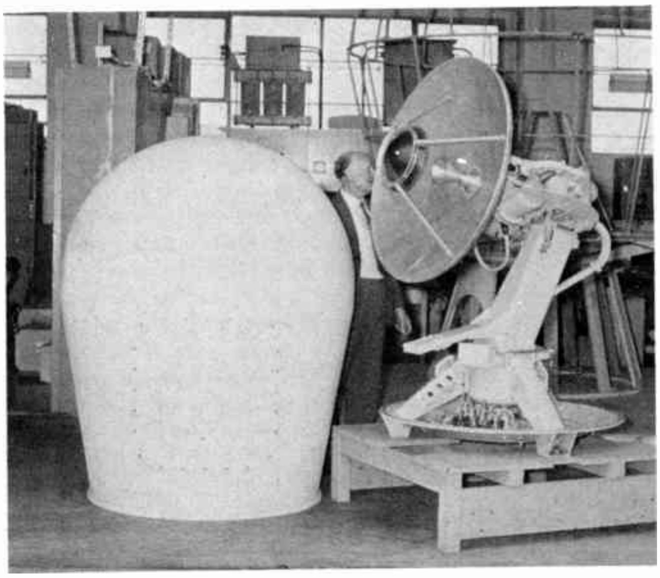

Fig. 3. SCOT 1 aerial head with protective cover removed.

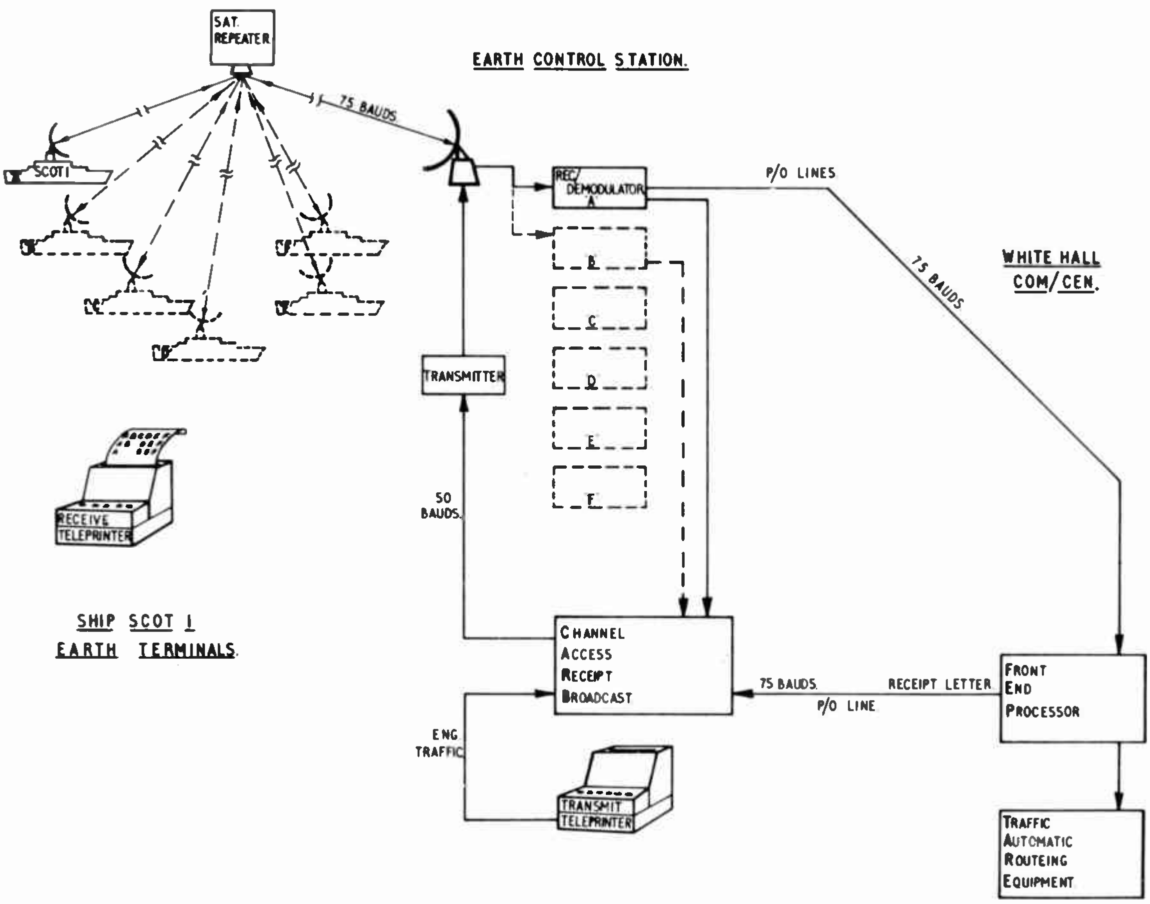

Fig. 4. Schematic of SCOT ships deployed with Master Engineering Control Centre.

incorporating a relay which gives remote indication of the state of a Phase Lock Loop. Relay contact closure signifies that the phase lock has not been acquired and so this demodulator condition represents a free access. Relay contact closures are sensed by circuits within the CARB equipment. The information on availability is then broadcast to all ships.

The availability broadcast can be transmitted in the following circumstances:

at regular intervals with periods selected by the user, and whenever there is a change of access indication, i.e. when any access is either acquired or relinquished;

when the output control key is in the 'CARB' continuous position;

at the end of a block of engineering traffic if an access change has occurred during the engineering busy state.

Receipt data are also entered into the availability broadcast. In addition if a channel availability broadcast is being transmitted when a receipt demand occurs, the equipment will allow this broadcast to finish and will then immediately transmit the receipting information. If a demand for a broadcast is received during a receipt transmission the equipment will transmit the broadcast after the receipt transmission has finished.

The standard SCOT 1 has now been in service several years, it has a narrow bandwidth, a limited tuning range and is restricted in the choice of communication and beacon frequencies. For inter-operability with more recent defence satellites the SCOT terminal is required to be more flexible, furthermore, future expansion of the system requires increased facilities. An Improved SCOT 1 is already being designed to provide inter-operability with the next generation of defence satellite systems by increasing the tunable bandwidth and providing an enhancement of the traffic handling capacity by introducing multiplex equipments and by the adoption of spread spectrum modulation techniques to ease the multiple access problem.

The new command cruisers shortly coming into service will be fitted with a variant of SCOT 1 known as SCOT 2, the difference being 1·8 m diameter aerials instead of 1·1 m and a secure speech baseband similar to the *Skynet V* terminal. Thus the SCOT 2 can be considered an interim measure between SCOT 1 and the proposed Improved SCOT.

The Royal Navy has now considerable experience in the operation of s.h.f. satellite communications as a system, perhaps more than any other military force to-day.

Contributors to this issue*

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Brian Wyndham joined the Radar Research and Development Establishment, now the Royal Signals and Radar Establishment, Malvern, in 1953, where he is currently a Principal Scientific Officer. He gained an HNC in Electrical Electronic Engineering at the RRE College of Electronics in 1953. His work has been principally concerned with ground radar and in recent years with developments in secondary surveillance radar for civil air traffic control.

David Bull completed an HNC course following his National Service with the RAF. After four years with Vickers Armstrong working on guided weapons development he joined ERA in 1959 to carry out research on discharges in gases. He moved to the Electromagnetic Interference Department in 1968 and is now a Principal Engineer in charge of aircraft investigations and flight programmes.

Brian Smithers graduated in physics at the University of London in 1955 and was subsequently awarded a Ph.D. degree for post-graduate work on electrical discharges in gases. His early training was with A. C. Cossor and the M-O Valve Company and he then joined the Electrical Research Association in 1958 to work on arc phenomena in relation to vacuum circuit-breakers. Since 1969 Dr Smithers has been with the Electronic Interference Department where he is at present a Principal Physicist.

*See also page 287

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The detection of ice at sea by radar

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SUMMARY

The paper begins with an historical review of radar for ice detection. The most up-to-date estimates of radar cross section of ice in various forms are presented and discussed. Comments are made on recent sea clutter measurements and the influence of scanner height on marine radar.

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1 Introduction

The threat of collision into ice has been present as long as man began sailing in arctic waters. The loss of the *Titanic* gave an enormous drive to obstacle detection in dark and foggy weather and the fitting of 2 cm microwave radar aboard the French liner *Normandie* in 1935 heralded the beginning of civil marine radar.

The satisfactory detection of icebergs has thus long been regarded as a problem area. But radar, which was originally regarded as the panacea for obstacle detection in bad weather, has had a chequered history of success and failure against floating ice.

Let us examine a slightly similar case of radar ice detection by the meteorologist. Here the spatial extent of a hail-bearing cloud presents a target so large that in most cases it fills the beam, and in many cases the complete resolution cell, of a search radar probing the sky to map out a hailstorm as opposed to the more usual rainstorm. The meteorologists have therefore been using radar as a research tool for many years, and back-scatter coefficients for falling rain, snow and hail are well known as a function of equipment parameters such as wavelength, polarization, etc., and drop size and density, temperature etc.[1] The whole subject was presented by Battan[2] as far back as 1954 and since then the literature abounds with ice observations.

The situation for floating ice is much more complicated due to many factors. These include variations of size, shape, aspect presented, surface conditions, height above water of each individual piece of ice, all of which will vary in some degree even whilst observations are being made. In addition, the target echoing area (r.c.s.) will be apparently modified by the lobe pattern of the aerial due to interference from the sea, and this will itself be some function of sea state.[3]

For the larger ice masses, i.e. bergs, the r.c.s. and height above water will give many miles of detection range on quite a modest radar, but the smaller growlers are regarded as a problem and this paper attempts to quantify expected radar performance for any given set of conditions by drawing on available results allied to a simple model.

2 The Nature of the Problem

2.1 Definition of Terms

To avoid ambiguity in the use of ice terminology, the definitions laid out in the illustrated Glossary of Snow and Ice[4] will be used in this paper. For sea ice, the ones of note are as follows:

ICEBERG

A discrete lump with more than 15 metres clear of the water.

BERGY BIT

Less than 5 m protruding above the water and less than 10 m across the exposed part. Generally with smooth surfaces due to the sea washing over and general weather erosion.

0033–7722/79/060275+13 $1.50

GROWLER

A piece of ice almost awash, smaller than the bergy bit above. The picture[4] shows a very smooth object, almost like a seal humped out of the water.

GREASE ICE

A thick, soupy layer floating on the water almost like an oil slick. Unlike the three classes above, this is the first type to be formed from scratch in its parent sea water, as opposed to the glacier parentage of the bergs, bits, growlers, etc. Its appearance is matt-like, having poor light reflecting properties.

SEA ICE

The results of fully formed grease ice, usually a fairly smooth profile.

SECOND-YEAR ICE

The consolidated result of the previous year's sea ice which has not melted during the summer, and is now reckoned to be 2 m or more thick with a more lumpy surface than the grease or first-year ice.

One final definition is that of ICE DENSITIES.

ICE DENSITIES

In the ice charts, fully open sea (99% clear of ice?) is denoted ice free.

1 okta denotes open water.
1–3 okta denotes very open pack.
3–6 okta denotes open pack.
6–7 okta denotes close pack.
7–8 okta denotes very close pack.
>8 okta denotes compact pack of ice.

Our main interest is, of course, in the detection of small discrete pieces of ice in areas the meteorological charts describe as ice free, i.e. 0 to 1 okta.

PIECE SIZE

The US Coastguards have issued an observer's handbook for arctic regions[5] with an even more down to earth arctic glossary which basically equates an iceberg to a large merchant ship, the bergy bit to a house and a growler to a grand piano, to help Naval personnel estimate sizes at sea.

2.2 Threat Probability

Prior warning of meeting a growler may be useful at the beginning and end of the winter. In 1971 the position of warning services for the spatial distribution in the North Atlantic during the year was as follows. Several sources of information have been available, which, in terms of up-dating of information, rank as follows:

Ice broadcasts transmitted by the US Ice Patrol Station (VIK) which are put out at 00.18 and 12.18 hours GMT in morse transmissions on the following frequencies daily:

0·427, 5·320, 8·502, 12·8805 MHz.

For those equipped with Mufax—duplex—radio facsimile recording gear, a map is put out daily at 13.30 GMT on the three higher frequencies of those used for the morse transmissions.

The British Meteorological Office puts out a similar service for the North Atlantic called Icefax at 14.00 GMT on the following frequencies:

4·782, 9·203, 14·436, 18·261 MHz.

Monthly dyeline maps[6] of the sea ice distribution and sea isotherms are issued by Bracknell within about a fortnight of the end of each month. These are prepared from some 15,000 weather reports sent in to Bracknell daily by stations and merchant shipping in the North Atlantic.[7] A more elaborate ice chart is also issued monthly,[8] and this includes satellite aerial photography of the area, but publishing times are longer than for the dyeline charts.

For those wishing to do their own forecasts based on someone else's analysis of the previous year's statistics, the UK Meteorological Office publish a very good small book entitled 'Marine Observer', which is a quarterly journal including maritime meteorology prepared by the Marine Division of the Meteorological Office.[9] This book includes a table of icebergs sighted by aircraft and merchant shipping within the following grid:

Latitude 40°N to 65°N. Longitude 40°W to 65°W.

There is also a fringe distribution just South of this area. For example, in one year the total number of significant large pieces of ice (bigger than bergy bits most likely) was 318 in July, 48 in August, and 40 in September, a total of 406 observed pieces, which gives some idea of the probable density in the so-called zero okta region

As well as these recent reports in the last year, the handbook issues a couple of long term up-dates, as shown in Table 1.

The arctic ice maps now have a very great input from the US Coastguards project 'Birds Eye' which has six objectives, one being the evaluation of radar airborne imagery to determine the ice edge and features within the ice pack. The preparatory experimental work for this part of the project is reported by W. K. Lewis and R. K. Moore.[10] Thus airborne sideways-looking radar is extremely useful for the *gross* analysis of the ice state in fields and over continents, but has not been used for the detection of isolated growlers floating in the sea clear of the main ice fields (to the author's knowledge).

Project 'Birds Eye' is reported on a sequential basis.[1] For example, the referenced report shows some very good airborne films taken of the ice-breaker *Manhattan* whilst she was in the Melville Sound during her arctic voyage.

From a very limited discussion, it would seem that a family tree type distribution—1 iceberg = 10 bergy bits = 100 growlers—is most unlikely as the iceberg may last for many years, only very slowly shedding its bergy bits and growlers, and the sea currents rapidly disperses these such that the distribution of growlers cannot be sensibly related around each major iceberg.

Table 1 Incidence of icebergs and pieces in the North Atlantic.[9]

	Month											
Year	Sept	Oct	Nov	Dec	Jan	Feb	Mar	Apr	May	June	July	Aug
1967/8	4	0	0	0	0	0	0	0	35	17	1	0
Average 1900 to 1969	3·6	1·6	1·6	1·2	1·9	7·2	38·4	97	122	63	20	5·7
Average 1946 to 1969	0·2	0·1	0·2	0·2	0·4	3·6	24·4	82	62·7	39·3	9·3	0·5

It therefore seems that merely keeping clear of actual large icebergs using radar is not a sufficient insurance policy to avoid growlers, and a margin of many hundreds of miles is really called for to raise the probability of missing ice to any acceptable safe value. Therefore the criterion for a radar set that it can spot icebergs only cannot be considered a safe method of navigation in arctic waters.

3 A Survey of the Literature

3.1 General Bibliography

Going back in time, it would seem that R. E. Perry's[12] log of his voyage on the S.S. *North Anglia* is one of the earliest useful fact-finding voyages with open publication.

Capt. F. J. Wylie[13] in his book 'The Use of Radar at Sea' makes three observations as follows:

SHEET ICE

> Little return except from the edges. Its presence can be seen as a black contrast against a white sea return, generally up to 3 nautical miles.

GROWLERS

> He merely notes that they are small, weak, irregular echoes which may be difficult to perceive in clutter.

BERGS

> Have a wide variation of echo (as one would expect) and can be seen between 3 and 15 miles, and normally have a slow movement.

Sonnenberg[14] does not clarify the problem very much further" other than to say that the radar display will give a clear indication of the situation once one is within the ice field, but that growlers can only be seen a few miles distant, and then only providing the sea clutter returns are of low magnitude.

One of the most recent books entitled 'Polar Operations' by a retired US Navy Captain, E. A. MacDonald,[15] makes very small reference to the use of radar, merely suggesting 'the gain be turned down to sort out conditions of either sea or ice clutter'.

3.2 Early Quantitative Measurements

The above situation appears to have been accepted by most marine users for 20 years, and the average short literature search stops at this point. However, on looking into the matter a little further, the US Coastguards ran a very good two-year investigation into the detection of discrete lumps of ice in 1945/46, which remained classified for many years, to be published in May 1959.[16]

For this exercise the US Coastguard ship *Mojane* spent a long time off the Grand Banks, Newfoundland, within the area contained by latitude 42°N to 52°N and longitude 45°W to 55°W. They carried a couple of American service radars which had been arranged for either p.p.i. presentation with rotating aerials, or could be stopped and hand trained on a chosen ice target whose amplitude was measured on an A-scope while the ship was making radial runs in and out.

The basic parameters of the sets are given in Table 2.

Table 2 Parameters of radars in US Coastguard trials.[16]

	Set 1	Set 2
Frequency	S-band	X-band
Aerial height	22·1 m	24·6 m
Aerial aperture	60 cm	60 cm
Horizontal beam	12°	3·8°
Vertical plane	Unstabilized	Stabilized
Transmitter pulse length	1 μs	1 μs
Transmitter p.r.f.	400 s^{-1}	600 s^{-1}
Transmitter power	+75 dBm, i.e. 32 kW	+76 dBm, i.e. 40 kW
Receiver sensitivity	−96 dBm	−89 dBm
Aerial gains	unknown	unknown
Feeder losses	unknown	unknown

In this report photographs of actual lumps of ice are given with the excess performance on the two radars, which were used at the same time to record signal strength compared to that of a signal generators (pulsed) to give a diagram similar to Fig. 1.

On these diagrams were superimposed sea clutter returns which had been taken at different times as the

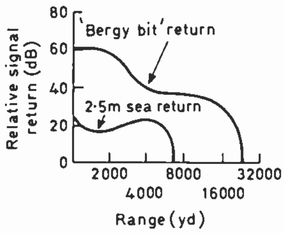

Fig. 1. Ice and sea returns on X-band.

opportunity arose, and thus they were able to forecast up to what sea state each particular size of ice 'lump' could have been picked up with some degree of certainty.

From these particular trials one gets the impression that against icebergs the S-band was more effective than the X-band set, but unfortunately in their result analysis the respective λ^2 terms became mixed up, this, coupled with a lack of accurate knowledge of aerial gains, spoilt an otherwise well-planned trial mounted with the best of intentions.

The final conclusions of Ref. 16 are a great deal better on the particular targets observed than the very general observations given in Section 3.1, and some specific details have been extracted.

3.3 Probable Visibility of Growlers Based on 1945/46 U.S. Coastguards' Work[16]

It is understood that visual sightings resemble radar ones in that both are degraded in rough seas (due to wave obscuration or general spray cover?). The signal/clutter ratios were taken from data at 5000 yards as reference range.

Sea State 1 (Smooth)

Visually the growler reflects any light available and is seen against a smooth sea background for a large 3 m high growler; there is a 17 dB differential in signal to sea clutter received power in favour of the growler at S-band, and a 6 dB differential at X-band.

Sea State 2 (Slight)

The growler is now recognized not so much in its own right as by the mist, spume and general area of disturbed sea around it. For a 3 m exposed growler the S-band growler to sea clutter is given as 10 dB, and at X-band increases to 11 dB.

Sea State 3 (Moderate)

The growler is now less easy to recognize visually, and on the radar the following ratios were measured:

$$\text{S-band: } \frac{\text{growler}}{\text{sea clutter}} = +5 \text{ dB}$$

$$\text{X-band: } \frac{\text{growler}}{\text{sea clutter}} = +8 \text{ dB}$$

i.e. it is just about possible to distinguish it on a conventional X-band radar.

Sea State 4 (Rough)

Very difficult to pick out visually at 2 to 3 miles; on the radar:

$$\text{S-band: } \frac{\text{growler}}{\text{sea clutter}} = +6 \text{ dB}$$

$$\text{X-band: } \frac{\text{growler}}{\text{sea clutter}} = +3 \text{ dB}$$

i.e. an apparent improvement for S-band, but very difficult ratios to operate on.

Sea State 5 (Very Rough)

The radar readings gave:

$$\text{S-band: } \frac{\text{growler}}{\text{sea clutter}} = +1{\cdot}5 \text{ dB}$$

$$\text{X-band: } \frac{\text{growler}}{\text{sea clutter}} = +0{\cdot}2 \text{ dB}$$

Neither of these ratios is worth considering as they are too small to be significant.

All of the above results are subject to at least two corrections:

(a) The growler values of signal strength are *peak*, not mean or median, which makes them optimistic.

(b) The clutter measurement was made with relatively broad beam aerials, i.e. 12° S-band and 3·8° X-band which increases the clutter patch illuminated beyond the optimum of a 2° beam as separately reported (Ref. 17, Chapter 31, page 28).

These two effects tend to cancel and, as far as can be ascertained in the open literature, the foregoing are the only real data available, apart from general observations, up until 1973 when Williams[18] published results of a carefully controlled trial by R. J. Holden carried out on M.S. *Thunfish* in 1971.

3.4 More Recent Work

In the last few years the bulk of observations on ice appears to have been carried out either from a satellite or aeroplane. The sensing of the environment by flights of one sort or another is attracting a great deal of attention, and in particular the work of Professor Richard Moore and his team at the University of Kansas is perhaps the most important.

It is convenient to review this work under four separate headings as follows:

(1) Sideways-looking airborne radar (with synthetic aperture radar as a special case).
(2) Radiometers. Passive techniques.
(3) Scatterometers.
(4) Conventional pulsed radars.

3.4.1 *Sideways-looking airborne radar*

Basically a large, fixed aperture aerial used with a high definition radar operating on X, J or Q-band is put aboard a plane and composite photographs built up of a radar picture after flying a strict pattern over the area of interest.

Such photographs as shown on page 25-45 of Ref. 17 indicating various types of ice were taken by the University of Kansas team, using a K-band equipment. This gross treatment enables one to identify leads in pack ice, but its sensitivity to the detection of growlers in an otherwise ice-free sea is not known.

Methods of improving the resolution of this type of radar can become very complicated and include the use of Doppler filter banks to yield superior azimuth resolution and holographic playback techniques on the ground after the trip. (The Doppler focusing or synthetic aperture radar technique.)

But basically, values of backscatter from the ice tend to be *average* ones from *fields* rather than those of *discrete targets* which do not fill the radar beam or effective resolution cell. (Cf. the meteorological case.)

3.4.2 *Radiometers*

A technique of possible detection is available using the differential radiation from earth, sea and ice at various temperatures. Although at very short wavelengths the techniques are those applicable to infra red, as the wavelength increases to a few millimetres the radiometer receivers bear a strong resemblance to the ones used in radar.

For example, on page 39-28 of Ref. 17 a data strip photograph is shown of pack ice and various sizes of icebergs as detected from an altitude of 5500 feet through heavy fog and cloud with a 15 GHz radiometer. The equipment used was the US AN/AAR-33 radiometer. No indication of scale is given, and one would not expect less than a medium-size berg to be detected on the evidence of this one photograph.

The essential difference between radiometer and the radar device is its purely passive or receiving role, i.e. there is no active transmitter as in the radar, only the 'heat' output from the various targets being viewed and, of course, there is no range information.

3.4.3 *Scatterometers*

The scatterometer is a general name given to any active radar device used to measure backscatter from terrain, and includes pulse radar—sideways-looking radar and synthetic aperture radar. The simplest scatterometer is a c.w. system which relies on a small beamwidth aerial to achieve the only resolution possible whilst looking down at the earth from near the nadir or vertical.

3.4.4 *Pulsed airborne radar flying at low altitudes*

By making use of the expression σ_0 for the backscatter coefficient given by:

$$\sigma_0 = \frac{P_R(4\pi)^3 R^3}{P_T G^2 \lambda^2 \frac{C\tau}{2} \theta} \text{ in m}^2 \text{ per m}^2$$

Table 3 Backscatter coefficients (area extensive)[19]

Depression	Aerial Polarization	X-band Ice	X-band Ice	S-band Ice	S-band Snow
Angle	H or V	‖	⊥	‖	⊥
2°	H	−40	−45	−45	−53
2°	V	—	—	−46	—
1°	H	−34	< −40	−37	−46
1°	V	−38	< −41	−45	—

‖ Indicates receiver aerial polarized as transmitter aerial.
⊥ Indicates receiver aerial cross-polarized relative to transmitter.

when using a pulse radar of known parameters, i.e.

P_R = received power (the variable being measured in a flight)
R = range to earth (which can be gated where desired)
G = aerial gain
θ = aerial beamwidth (azimuth plane)
τ = pulse length
C = velocity of e.m. waves

The values of σ_0 can be found when looking at continuous ground or ice clutter at more oblique angles than the scatterometer operating from or near the nadir.

Thus σ_0 is the backscatter coefficient in m² of r.c.s. per m² illuminated in a resolution cell determined in this case by aerial horizontal beam width and pulse length when the gate width is set to τ.

Ringwalt and MacDonald reported[19] the results of flights over the Thule area of Greenland in 1956 and some of these are shown in Table 3.

This illustrates gross coefficients of the order of 1000 to 100,000 times down on the physical area illuminated. The work was primarily carried out to enable the radar contrast of an airborne radar flying over ice to be obtained so that the moving target indicatior problem could be quantified. As such it does *not* give an immediate guide to the particular problem of the detection of small pieces of ice. However, the results will be used in Sections 5 and 6 on Elementary Models.

The ice and ground surface was sufficiently rough to cross-polarize the returns enough for the perpendicular (⊥) column to give results lying between 3 dB and 9 dB of those returns received in the same polarization as that transmitted parallel (‖).

Figure 2 shows the above results and others from Refs. 20 and 21 plotted together. It can be seen that the backscatter approaches a maximum value as the nadir is approached. One would expect that more data points for ice would keep it below snow around 20°.

3.5 Japanese Work

In 1969, Tabata *et al.* published[22] a lengthy report on the distribution of pack ice off Hokkaido, using three coastal stations operating on C-band. A translation from the original Japanese paper was made by E. R.

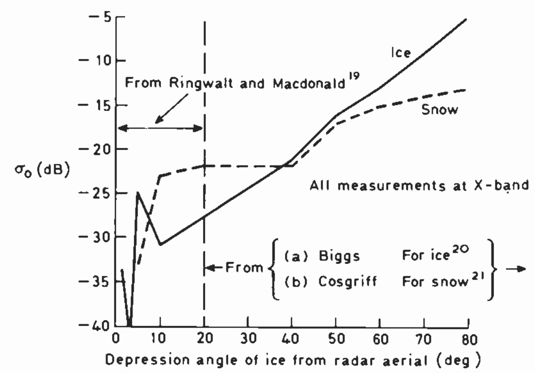

Fig. 2. Backscatter coefficient σ_0 for snow and ice as a function of depression angle for an X-band radar using horizontal transmissions and reception.
(Note σ_0 includes the depression angle correction.)

Hope of the Canadian Defence Research Board[23] and kindly made available to the author.

Tabata *et al.* quickly found that their observations were often confused by sea clutter, particularly when the ice was just forming and smooth compared to second year ice, or that from glacial sources. Their later reports[24] were again translated for the Canadian Defence Research Board by E. R. Hope.

Since this 1970 Japanese work the US ERT's satellite, followed by the unfortunate SEASAT-A's short useful life, has produced a wealth of remote sensing activity, for example by the 1978 publication on Remote Sensing.[25]

4 Ice Constitution

The question of the microwave reflectivity of a thick, plane ice surface is discussed in Section 6 and the present Section makes no more than a cursory examination of the likely make-up of sea water, ice and pieces of ice broken off from a glacier.

The main constituents of any piece of ice will be:

(a) pure water (in frozen form);
(b) occluded gases (mainly air);
(c) impurities, typically the liquid brine in young ice formed from the oceans.

R. C. Byrd, discussed the problem in some detail in his paper, 'Interpretation of radar returns from sea ice' at the 1971 IEEE Conference on Engineering in the Ocean Environment.[26] However, it must be emphasized that he was addressing a problem closely associated with the meteorological hail–rain problem. Here the radar is looking at one filled beam patch of the sea ice surface and records a signal significantly different from one perhaps one mile away; all other parameters such as the radar ones, scanner height (be it airborne or shipborne) and most important, ice surface topography or roughness, are the same so that the backscattering coefficient σ_0 was only a function of ice impurity.

Now in this paper the problem of small growler detection may well have as its biggest variable, the *aspect and particular shape* as the *dominant* variable, and ice content and purity would seem to have a small effect on growler r.c.s. in this context.

Byrd naturally refers to the work of Weeks, Hoekstra, Tinga and Evans and others listed as Refs. 27 and 36 and concludes that the influence of liquid brine on the real part of the complex permittivity (ϵ') of sea ice generally found in nature will only be felt at temperatures around melting point. This, he remarks, is borne out by Weeks[28] who found that for all sea ice of salinity 9% or less, the brine represented less than 10% of the total volume. The older the ice, of course, the less brine is contained therein as it migrates out even at its minimum temperature.

Byrd also discusses how ice formed from snow is likely to contain a high percentage of air, but continuous melting and refreezing increases its density and leaves old ice virtually air free.

5 R.C.S. Dependency on Shape

Much of the ice radar reflectivity data discussed so far has been mainly associated with large pieces or sheefs of ice so that the radar beam is filled in either one, two or three dimensions of length, i.e. large in area or volume.

This means that some form of imaging is possible but for growler detection this is not possible with microwave radars stopping in frequency at 30 GHz, or even 80 GHz. Only at 3000 GHz ($\lambda = 10\ \mu$m) can a radar have sufficient true angular beamwidth to enable imaging of growlers to be carried out, otherwise the $2D^2/\lambda$ criteria means that a spatial resolution set by the physical aperture size is the limiting factor. (Aerials may, of course, be focused but to do this dynamically in real time to match the velocity of light is not a trivial engineering task.)

We therefore must consider the general case of small object r.c.s.:

(1) Mean r.c.s. is proportional to the optical silhouette.
(2) Actual or instantaneous values of r.c.s. are highly variable with many radar parameters; to list but a few:
 (i) Wavelength.
 (ii) Polarization on transmit and receive.
 (iii) Object aspect for all but a true sphere.
 (iv) Degree of wetness and following this, its temperature.

The statistics of r.c.s. variability were first studied by Swerling together with Marcum.[37]

The important point to remember is that all the fading models cross-over around the median or 50% probability point (or a little less at 30%) to diverge widely at the high probability end. Thus the median values of irregular objects correspond roughly with their actual projected silhouette on a plane orthogonal to the line joining target and radar, whilst peak values (associated with target glint) occur infrequently and cannot be relied on for consistent detection. Now these peak values can vary according to various wavelength dependencies, Nathanson in his 1969 book[38] listing the following relationships for the changing of r.c.s. at two wavelengths, where λ_L is the longer and λ_s the short one:

$(\lambda_L/\lambda_S)^{-2}$—for a flat plate
$(\lambda_L/\lambda_S)^{-1}$—for a cylinder
$(\lambda_L/\lambda_S)^{0}$ —for a sphere
$(\lambda_L/\lambda_S)^{+1}$—for a curved edge or disk
$(\lambda_L/\lambda_S)^{+2}$—for the sharp end of a cone.

(Note that the sphere has dimensions very much larger than either wavelength to take it out of the Rayleigh region into the optical region.)

When the flat plate is viewed from *other* than a normal axis, Snell's law tells us that most of the incident energy is reflected not at the radar but out into space; like an aerial, only a small fraction of the energy is sent *back* to the radar, perhaps -13 dB or -20 dB, in a similar manner to the sidelobe radiation of a parabolic mirror fed with a feed horn.

Thus a growler or larger piece of ice with a gently sloping face will act as a mirror directing energy away from the radar. On the other hand, pockets or crevices in the ice surface may act as corner reflectors[39] to increase the backscatter in direction of the radar.

Let us examine three cases of target whose optical silhouette presents a disk of diameter 2 m, and the radar operates on 3 cm.

(i) A true flat plate placed normal to the line joining itself and the radar:

$$\text{peak r.c.s.} = \frac{4\pi A^2}{\lambda^2}$$

where A is the true area in square metres and is equal to

$$\pi \text{ m}^2.$$

Thus

$$\text{peak r.c.s.} = \frac{4\pi^3}{(0{\cdot}03)^2}\text{m}^2 = 137\,805 \text{ m}^2.$$

Note an enormous value compared to the optical area of π m^2.

(ii) A sphere of diameter 2 m^2: As this dimension is very large compared to the wavelength the r.c.s. is in fact the projected optical area of $\pi D^2/4 = \pi$ m^2.

(iii) The flat plate inclined at some angle, unlike (i): The usual treatment is to assume that the energy is reflected away and the effective r.c.s. is negligible, but in fact it is the same situation as with the side lobes of an aerial and the distribution is of the form $\sin x/x$, as shown in Fig. 3.

For this case with its uniform illumination, the first pair of side lobes are no more than 13 dB down, i.e. still 6000 m^2, and the first null in practice may be blurred

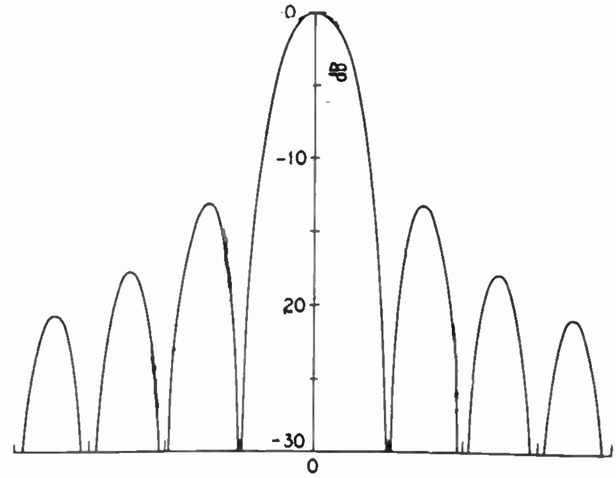

Fig. 3. $\dfrac{\sin x}{x}$ distribution.

and no more than 30 dB down, so that even if the plate is orientated to show a null the r.c.s. is still likely to be 137 m^2. A return of only π m^2, comparable to the sphere, is most unlikely in practice, even for a short time, as such orientation will not be held for long, even in a smooth sea, due to the ship's movement.

It is therefore suggested that the sphere model is perhaps as pessimistic as one can get as regards shape.

6 R.C.S. Dependency on Constitution or Surface Reflection Coefficient

The question to be answered is: Is the growler to be regarded as

(a) a low loss dielectric sphere;
(b) a lossy dielectric sphere;
(c) a conducting sphere.

Pure ice in a completely dry and clean form is relatively loss free, although dirt and saline, particularly in liquid form, can influence this. In Table 4 a few extracts from Von Hippel[40] are given.

Table 4

Dielectric constants and loss factor ($\times 10^4$) for water and ice.

Substance	Wavelength S-band (10 cm)		X-band (3·17 cm)		J-band (2 cm)	
	ϵ'/ϵ_0	$\tan\delta$	ϵ'/ϵ_0	$\tan\delta$	ϵ'/ϵ_0	$\tan\delta$
Water at +15°C	78	2050	49	7000	25	330
Water at +5°C	80	2750	41	9500	17·5	3950
Water at +1·5°C	90	3100	38	10300	15	4250
Pure ice at −12°C	3·2	9	3·17	7	3	?

The striking difference of both ϵ'/ϵ_0 and $\tan\delta$ for water in its liquid and solid states are well known, especially in the microwave reheating field of frozen food where it is very difficult to couple energy into low-loss frozen food.

If the reflection coefficient ρ of metal is taken as unity, what is the likely value for pure ice? Let us look at the usual equation

$$\rho=\frac{\sqrt{\epsilon}-1}{\sqrt{\epsilon}+1}$$

Using values of $\epsilon=3{\cdot}2$ (as ϵ is substantially constant for ice at all microwave frequencies), then

$\rho\approx 0{\cdot}28$ (An r.c.s. loss is $\rho^2=0{\cdot}07$).

This would be the voltage reflection coefficient for a single interface; for a loss-free dielectric there will be a similar reflection as the wave tries to leave the ice. This will be dependent on the ratio of dimensions of wavelengths/ice in determining if it is going to be constructive or destructive interference, and hence the total reflection varies from two interfaces each of $\rho=0{\cdot}28$ for a complete in-phase reflection to a nominal zero reflection when the two sets of waves cancel (cf. the Mie region for raindrops).

Thus at resonance the total reflection coefficient ρ has a value of 0·38, giving a loss of 4·2 dB compared to a metal sphere.

Lossy material in a growler will attenuate the internal wave quite quickly so that one might expect three 'boundary' conditions:

Metal sphere $\rho=$unity, i.e. 0 dB as a reference.

Loss-free sphere $\begin{cases}\rho=0{\cdot}38 \text{ or a 4 dB loss,}\\ \rho=\text{say } 0{\cdot}01 \text{ or a 20 dB loss.}\end{cases}$

where the 4 and 20 dB represent constructive and destructive interference conditions.

Lossy sphere $\rho=0{\cdot}08$ or an 11 dB loss.

Whilst these seem a wide variety of values a small crevice on a smooth sphere may well restore the r.c.s. None of these cases is as bad as the 'gross area' reflectivity coefficients discussed in section 3.4.

7 Best and Worst Models

This paper has shown that an ice growler of part spherical shape above water is a fairly pessimistic model and when lossy might have an r.c.s. value down by a factor of −11 dB. As the shape of the growler no longer tends to a sphere (or ellipsoid) the r.c.s. may well rise by 10 dB quickly, off-setting the lossy negative contribution as a dielectric.

As soon as a growler is sea-washed the very high value of dielectric constant sheathing so provided may well increase its r.c.s. to that of a metal sphere.

We therefore have a model for a small piece of ice and if that is in the 'growler' category we may take the part that appears above the surface to be a hemisphere of radius 1 metre. As the bottom part is touching the water, the field there must be tending towards zero so that if we regard it as a complete sphere of diameter 1 metre, an echo centre half a metre above water is a fair approximation.

The real reasons why growlers are hard to detect are:

(a) They are usually smooth with no corners or flat surfaces such as small ships or periscopes have.

(b) They are extremely low lying.

(c) They are often obscured by quite modest waves between themselves and the radar, particularly for low scanners giving low grazing angles.

(d) And, perhaps most important, they are usually hidden by sea clutter.

The model so far described represents an r.c.s. peak value of πr^2 or $\pi/4$ m^2, say 1 m^2.

From the very recent work especially by Schleher,[41,42] it has been shown that marine targets tend to have a log normal distribution and the peak r.c.s. exceeds the mean by typically 10 dB and this mean value in turn exceeds the median value by 10 dB. Our own measurements[43] show peak to median values ranging from 5 to 25 dB for reputedly non-fading targets such as Luneberg lenses anchored at sea.

Thus it is likely that the appropriate value to use on a civil marine radar employing a p.p.i. is not 1 m^2 nor 0·1 m^2 but 0·01 m^2.

To put this in perspective, Table 5 has been prepared from an earlier work published by the author[43] and shows the likely value of a growler r.c.s. median value.

It should be noted that in calm seas a modern marine radar is capable of detecting 0·01 m^2 at 2 miles, i.e. a performance 30 dB superior to that required by the UK Board of Trade requirements for Civil Marine Radar. The real problem s detecting such a target in the presence of clutter.

8 Practical Measurements

A summary of R. J. Holden's work reported by Williams[18] is given in Table 6 when 6 sets of ice targets were searched for using a pair of radars both of which operated at 3·2 cm but whose other parameters varied significantly regarding aerial beamwidth, transmitter pulse length and aerial height h_r as given in Table 7. (Photographs of the targets are given as Figs. 4 to 9.)

For the small discrete targets offered as growlers, the low resolution set gave the best results in sea clutter but for a more extensive type of target filling the radar beam and pulse length, i.e. whose area exceeded the radar resolution footprint, the set with the better spatial resolution and greater aerial height gave significantly superior results.

Thus it is shown that to depend on one radar, no matter how superior its performance may be, is less advantageous than using two sets with different scanners positioned at significantly differing aerial heights.

The diversity in spatial resolution and scanner height provided by two moderate cost civil marine radars is

Table 5

The growler in relation to measured values (σ_{05}) for other small marine targets.

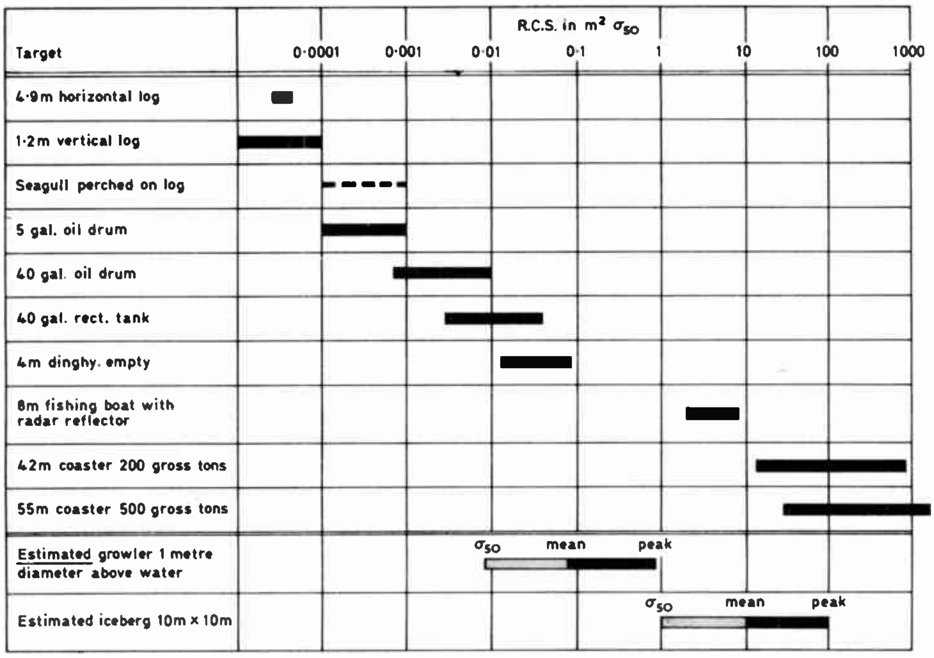

Table 6 Observations on various types of ice targets by a pair of marine radars aboard a ship. Note that the ice on Fig. 9 was unobserved by either radar.

Target	Sea state swell	Wind force	First range seen on approach		Range improvement ratio H.R./L.R.
			H.R. set	L.R. set	
Growler (or even bergy bit) (Fig. 4)	10 ft–12 ft	7	Never seen and recognized	0·6 n.m.	0
Growler (Fig. 5)	10 ft	5	0·5 n.m.	1·2 n.m.	$\frac{0.5}{1.2}=0.42$
Growler (Fig. 6)	Confused	—	Never seen and recognized	0·5 n.m.	0
Several Growlers in pack ice and mush (Fig. 7)	5 ft	3	3 n.m.	2·5 n.m.	1·2
Pack ice of above (Fig. 7)	5 ft	3	12 n.m.	2 n.m.	6
Ribbon ice (Fig. 8)	5 ft	2	12 n.m.	4 n.m.	3
Mush and brash ice (Fig. 9)	4 ft	3	Ice never spotted by radar		

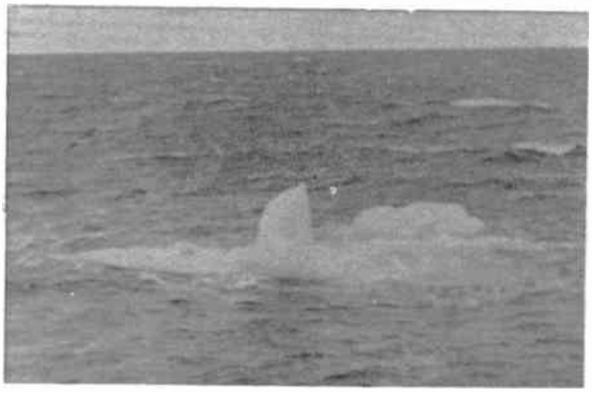

Fig. 4. Growler at 40 yards in 10 to 12 ft swell. Wind force 7. This was never detected on the high resolution radar, but picked up at 0·6 miles on the low resolution radars.

Fig. 7. Growler at 100 yards in 5 ft swell. Broken pack and mush ice up to 12 n.m. Pack ice from 3 n.m. Wind force 3. First detected 3 n.m. on high resolution 2·5 n.m. on low resolution radars. Pack ice detected out at 12 n.m. on the high resolution set.

Fig. 5. Growler at 200 yards in 10 ft swell. Wind force 5. First picked up at 0·5 n.m. and 1·2 n.m. on the high and low resolution sets respectively.

Fig. 8. Swell 5 ft. Sea crystalline (−5°C). Ribbon ice at 1·5 n.m. Wind force 2. First detected at 12 n.m. and 4 n.m. on high and low resolution sets respectively.

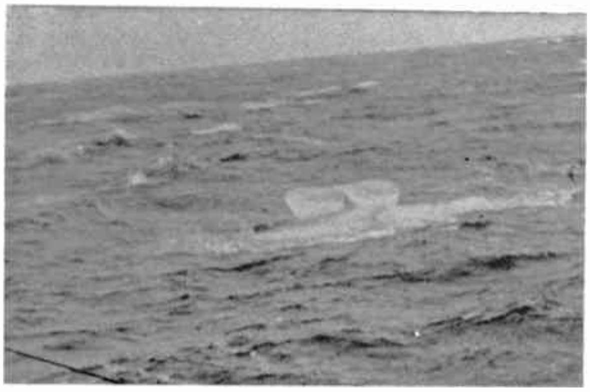

Fig. 6. Growler at 50 yards in confused swell. This one only detected at 0·5 n.m. on the low resolution set.

Fig. 9. Mush and brash ice in 4 ft swell. Wind force 3. No detection on either set.

clearly more cost effective than one sophisticated radar with its scanner positioned at a fixed height above sea level.

To conclude this Section on practical target measurements, Milwright and Le Page[44] in their early paper compared growlers to the *theoretical* performance of a metal sphere in *free space* and in those days the knowledge and understanding of radar system losses were not as complete as today. Therefore, an error of 10 dB could easily occur and that coupled to a multipath loss of, say, 7 dB easily allows for the 17 dB discrepancy as reported back in 1953,[44] even without considering the difference between σ_{peak}, σ_{mean}, and σ_{50}.

Table 7

Essential differences of marine radars used in Table 6.

Parameter	High resolution set	Low resolution set
Azimuth beamwidth	0·8°	2°
Aerial height	50 ft	30 ft
Long transmitter pulse	0·5 or 1 μs	0·6 μs
Short transmitter pulse	0·05 μs	0·08 μs
Transmitter peak power	25 kW	9 kW
Receiver noise factor	11·5 dB	10 dB

9 Radar Sea Clutter, Sea State or Local Wind Dependent?

Until a few years ago most working engineers considered that the radar backscatter from the sea surface depended only on sea state[38] but following the work of Wright[45] and Valenzuela[46] it has become accepted that the sea clutter backscatter coefficient σ_0 is more wind dependent than on sea state, particularly at X-band and shorter wavelengths.

To support this, a large amount of data taken by Small and later published by Williams *et al.*[47] has now been freshly analysed and the regression lines associating the maximum range of sea clutter seen on a conventional marine radar (without swept or adaptive gain control) to local wind speed are shown for separate periods throughout 1976–77 in Fig. 10.

From this it may be noted with some confidence that the wind speed over the sea is a better indication of sea clutter[48] than the sea state description long believed to be true and earlier reviewed by Nathanson.[38] The relationship between direction of maximum clutter and wind vector is also fairly evident in open water with the maximum sea return coming from the 'up wind' direction. In confined waters the situation is less evident but over a 12-month period a fair relationship is shown in the English Channel as given in Fig. 11 previously published by Williams *et al.*[47]

Finally the saturation effect at even modest winds has been noted by many workers indicating that the sea clutter problem is serious at quite modest winds after which a confused sea in a heavy storm is not much worse and in such conditions most ships have to reduce speed anyway to avoid storm damage from the high waves themselves.

10 The Effect of Scanner Height

We are now in a position to list the factors associated with using either a high or low scanner on a ship.

Traditionally a low scanner has meant:
low sea clutter return, but
obscuration due to funnels, samson posts, derricks, etc.
A high scanner has meant:
a good horizon range
a large value of transition range,[49] but

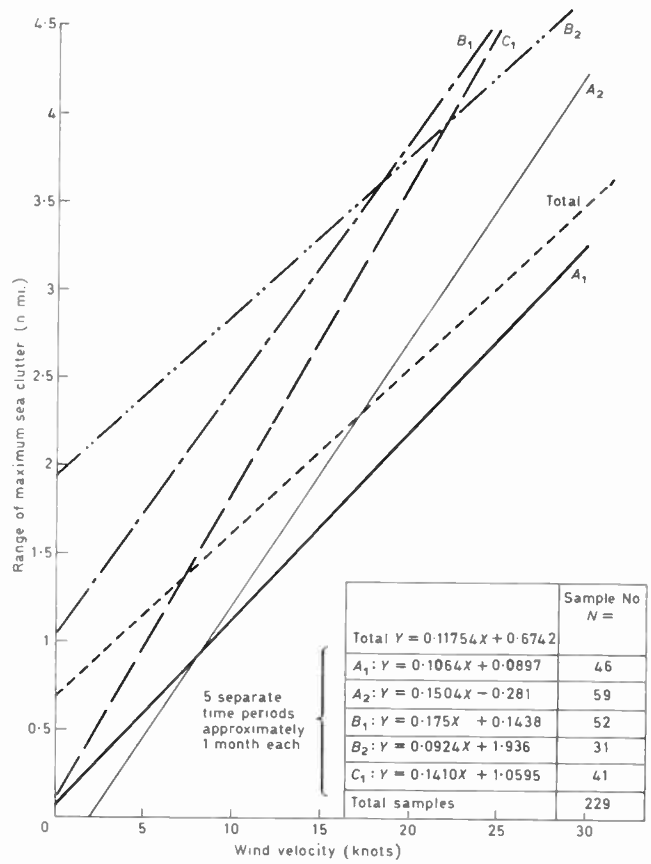

Fig. 10. Relationship between wind speed and sea clutter.

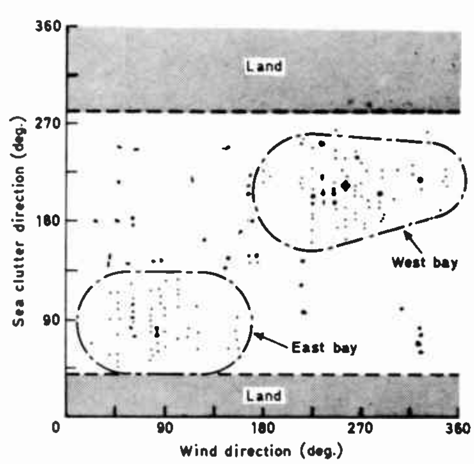

Fig. 11. Wind direction v. direction of maximum sea clutter in the English Channel.

extensive sea clutter return,
poor access for servicing,
excessive waveguide loss from a transceiver on the bridge to a scanner perhaps 60 m distant.

We now know a little more about the vagaries of the behaviour of the evaporative surface duct, although the excellent work by Doherty and Hood[50] has been hidden in the archives since it was published in 1957. A more modern review by Hay and Ting[51] includes a great deal of theoretical work but little in the way of trials results. In addition, the obvious fact that any small surface target will rarely be detected if it spends most of its time being obscured from the observing radar scanner by intervening waves has often been conveniently forgotten in the interest of not showing sea clutter on a p.p.i.

Thus it is mandatory that a ship seriously wishing to detect growlers, or indeed any really small targets, *must* first display sea clutter out to the desired detection range if there is to be any chance at all of making a detection. To achieve this two radar scanners are obligatory, and preferably more so as to be able always to launch beneath the duct and at the same time see over intervening waves as far as possible. Spatial diversity is therefore a far more cost-effective strategy than the use of a single radar, no matter how sophisticated it may be.

The satisfactory recognition and subsequent removal of sea clutter, particularly the 'spikes' associated with horizontal polarization at low grazing angles, must be regarded as the essential next step in growler detection in moderate to high winds. High speed scanning[52] has been shown to be effective in moderate seas but less effective when rough, and the practical realization of median detector[53] is perhaps the next step forward.

11 Conclusions

The radar detection of growlers at a worthwhile range of, say, 1 to 5 n.m.i. can only be accomplished with several radars and powerful anti-sea-clutter processing because of:

(i) the very small value of σ_{50} for growlers,
(ii) wave obscuration,
(iii) variability of propagation over the sea due to the evaporation surface duct,
(iv) competing radar clutter, especially that from the sea.

12 Acknowledgments

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Philip Williams studied electrical engineering at University College, Nottingham and joined the Electronic Switching Laboratories of Ericson Telephones, Beeston, in 1950. Two years later he went to the Development Laboratories of Decca Radar and since then he has had varied experience within the Company on marine radar, meteorological radar, low cost and p.a.r. equipment and pulse Doppler automatic search radar and on microwave heating. In 1969 he became Head of Future Projects, where his major interest is the relationship of theory to real results on working equipment. He has recently been acting as a consultant to Ice Engineering of Newfoundland who are concerned with investigating the safe transit of large tankers through the North West Passage.

In 1977 Mr Williams was awarded the Master of Philosophy degree of the University of London as a result of research work at University College which was described in his thesis on 'Dectection of small targets at sea by radar'.

Mr Williams received the Institution's Brabazon Premium in 1975 for his paper entitled 'Limitations of radar techniques for the detection of small surface targets in clutter' and the Clerk Maxwell Premium (jointly with Mr A. B. Schneider) in 1977 for the paper 'Circular polarization in radars'.

Surface-Skimming Bulk Wave Devices*

SSBW devices are a significant development which has come from the RSRE continuing basic research programme into SAW devices, and which is expected to extend considerably the capability of acoustic wave devices.

Over the past decade many groups in the UK and USA have investigated surface acoustic wave (s.a.w.) devices which are a family of electronic components operating in the range 20 MHz to 1 GHz and having the following attractive properties:

(1) They are small, cheap, rugged, planar components made from conventional materials like quartz of $LiNbO_3$ with a superimposed transducer pattern formed from an evaporated aluminium film by conventional photolithography.

(2) They are readily integrated with microelectronics, and are 'natural' components for adding selectivity to modern broadband microelectronic circuitry. For example: they are used as i.f. filters in television sets and radar receivers.

(3) Because of the *planar* construction the transducer patterns are infinitely variable, as are the electronic functions that can be performed by s.a.w. components. For example, they have been used in delay lines, tapped delay lines, bandpass filters, chirpfilters, resonators, oscillators, synthesizers, encoders and decoders, discriminators, convolvers and correlators.

A new family of acoustic wave devices which has been studied at RSRE employ surface-skimming bulk waves (s.s.b.w.) in which the crystallographic orientation of the substrate (e.g. quartz) has been so chosen that the transducers launch bulk waves rather than surface acoustic waves.

Provided a horizontally-polarized shear wave is chosen to satisfy the boundary conditions the transducers act like acoustic versions of endfire array antennae (Fig. 1). It has been shown that the *antenna gains* can be sufficient to offset the spreading losses of the acoustic beam. Thus these planar acoustic devices have insertion losses and electronic functions closely resembling those of s.a.w. devices but they offer several advantages for specific applications:

Higher propagation velocity which enables the limiting frequency of such components to be increased to 3 GHz.

Lower propagation losses which is important at microwave frequencies.

Superior temperature coefficients which makes them useful in stable oscillators and narrow-band filters. It is hoped that further work on s.s.b.w. will produce devices as stable as the conventional AT-cut bulk-wave quartz crystal oscillator.

Lower sensitivity to surface contamination because the s.s.b.w. energy is concentrated below the surface of the substrate. This should lead to reduced long-term ageing effects and simpler packaging of devices.

Relative freedom from interference from other unwanted acoustic waves.

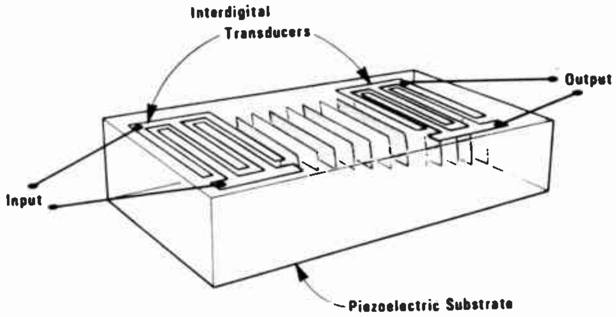

Fig. 1. Schematic diagram of the acoustic radiation pattern obtaining in an s.s.b.w. device. This pattern is similar to (one half of) the radiation pattern of the corresponding e.m. endfire array antenna operating in free space.

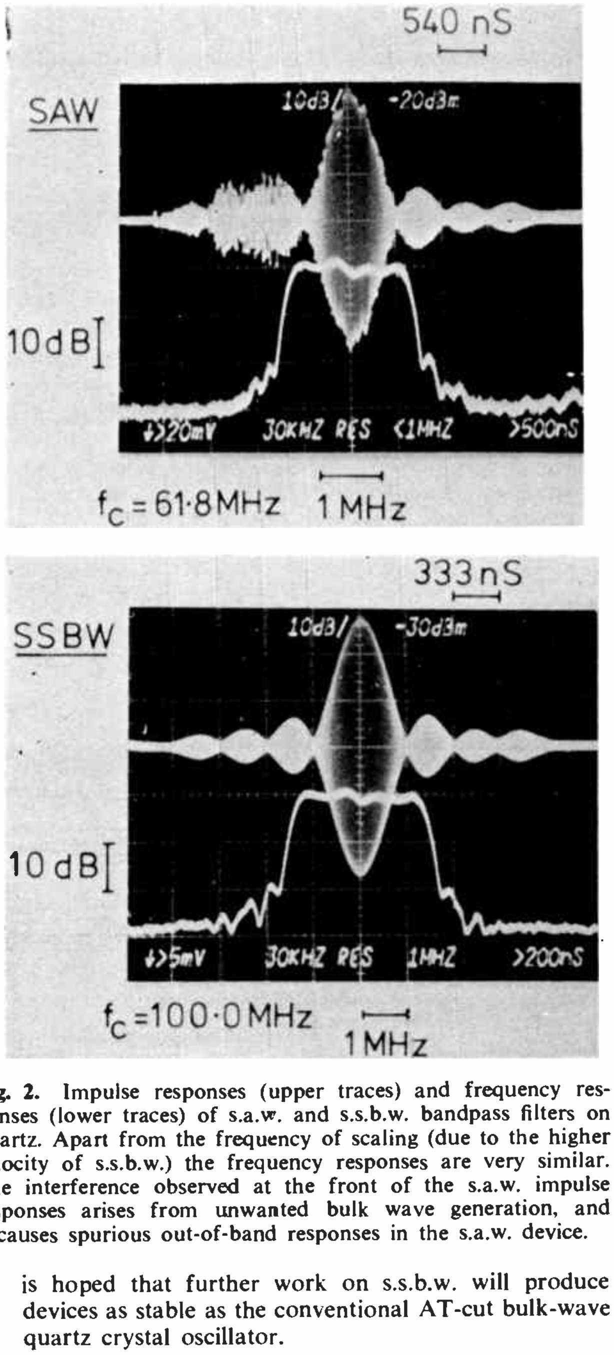

Fig. 2. Impulse responses (upper traces) and frequency responses (lower traces) of s.a.w. and s.s.b.w. bandpass filters on quartz. Apart from the frequency of scaling (due to the higher velocity of s.s.b.w.) the frequency responses are very similar. The interference observed at the front of the s.a.w. impulse responses arises from unwanted bulk wave generation, and causes spurious out-of-band responses in the s.a.w. device.

So far, s.s.b.w. oscillators and bandpass filters (Fig. 2) have been demonstrated on quartz. The investigation is being extended to other devices and new substrate materials.

Further information can be obtained from:

Dr. M. F. Lewis, RSRE, Malvern, Worcs, WR14 3PS (Tel. 06845 2733).

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Indexing Terms: Antennas, aircraft, Interference

Aerial isolation : a study of the interaction between co-sited aerials

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B.Sc., Ph.D.

Based on a paper presented at the IERE Conference on Electromagnetic Compatibility held at Guildford on 4th to 7th April 1978.

SUMMARY

The increased complexity of modern aircraft has resulted in a proliferation of aerials sited in close proximity to one another, a trend more marked in military types, which makes it even more necessary to ensure compatibility of working in a crowded r.f. spectrum.

The paper presents the investigations that have been made to determine the degree of isolation between transmitting and receiving aerials in aircraft within the frequency range 30 to 1250 MHz.

From measurements using cylinders, ground planes, helicopters and fixed wing aircraft, empirical formulae for the calculation of aerial-to-aerial isolation have been derived.

Investigations have also been carried out to determine the amplitude of the harmonics generated by airborne transmitters. The results obtained from both these aspects of the work are discussed.

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1 Introduction

Each new generation of aircraft requires an increased number of aerials for use with the many control, guidance and communications equipments that have been developed in recent years. This need is most marked in military aircraft, where systems have to operate over a very wide frequency range (say 10 kHz to many GHz) sometimes on very small airframes and sometimes on a very large airframe such as the E-4A, a Boeing 747 derivative[1] (Fig. 1).

ERA was asked to examine a number of aspects of the interactions between such co-sited aerials, with particular emphasis on the communications bands, to quantify the in-band and out-of-band isolation between systems on particular aircraft and to develop a procedure for the prediction of isolation between aerial systems on any aircraft. It has been uncommon, in most British military aircraft, to transmit and receive on two systems working in the same frequency band. For this reason emphasis has been placed on determining the degree of isolation between aerials where radiation from one aerial of out-of-band spurious transmissions, such as harmonics of a fundamental frequency f_1, affects another aerial and receiving system whose acceptance band embraces such spurious transmissions.

This paper describes some of the investigations that have been made to determine the degree of isolation between transmitting and receiving aerials within the frequency range 30 to 1250 MHz and has, for convenience, been divided into three Sections which are:

Investigations concerned with aircraft installations.

Laboratory investigations on aerials mounted on ground planes and on a cylinder, the behaviour of filters and the performance of in-service transceivers.

The development of empirical formulae.

2 Aircraft Installations—The Measurement of Isolation

In addition to the work carried out within a controlled environment in the laboratory, described in Section 3, many measurements of aerial-to-aerial isolation† were made between selected aerial systems in two transport aircraft and two rotating wing aircraft in order to establish some base-line information.

Values of isolation were obtained by injecting a sinusoidal signal into one aerial and measuring the resultant signal at the second aerial with a suitable measuring set. In order to avoid interfering with other users in the various communications and navigation bands the majority of the measurements were made with low levels of radiated

† Aerial-to-aerial isolation is the inverse of coupling where coupling C is defined as:

$$C = P_2/P_1,$$

where

P_1 = power which signal generator can supply to a matched load directly connected to the generator,

P_2 = power developed in load of measuring receiver.

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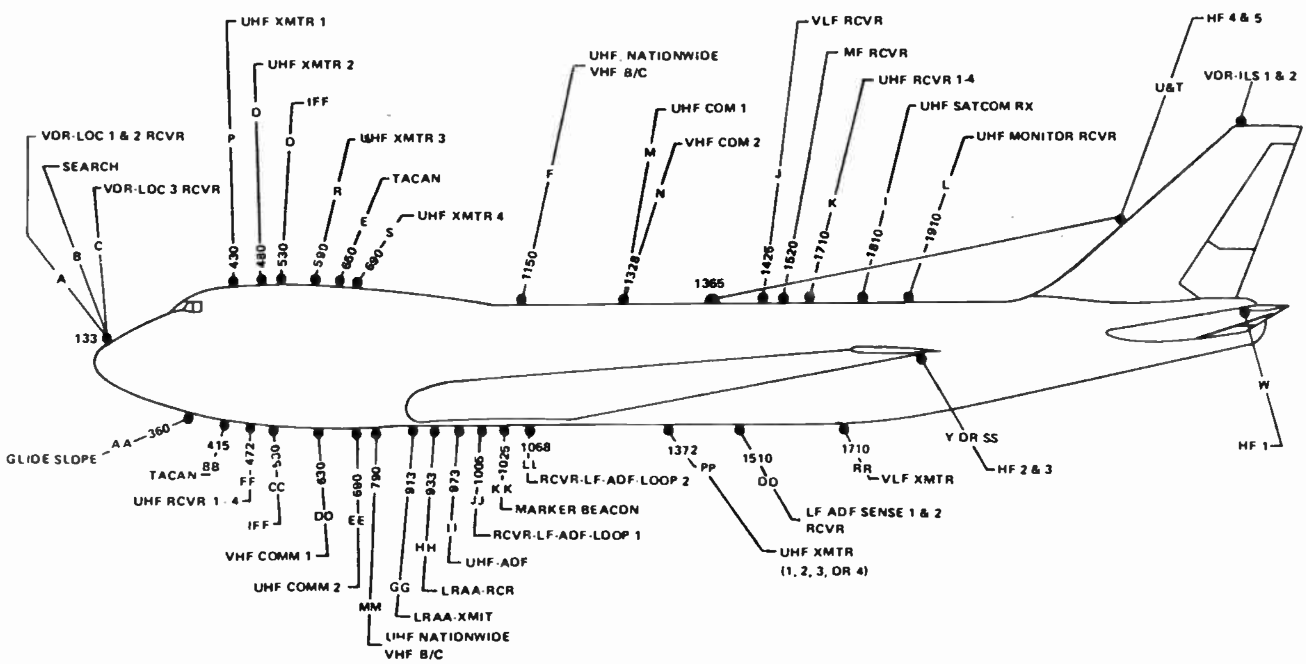


Fig. 1. E-4A aerial locations (after Dillion[1]).

power. However at certain frequencies comparative measurements were made at the nominal rated power for the particular equipment, to determine linearity.

Initial investigations were concentrated on those communications aerials designed to operate in the band 225–400 MHz and on the effect of transmissions in that band on the installed equipments working in L-band. Figures 2(a) to (d) show examples of isolation that were obtained from these measurements. Measurements were also made on communications equipment in the v.h.f. band and on certain navigational equipment within the frequency range 75 to 1250 MHz.

In the first series of tests, the fixed wing aircraft were positioned, on reinforced concrete, sufficiently far from any reflecting objects to minimize spurious responses. Similar measurements were then made during a series of flights and the results compared.

Both helicopters in the second series of tests were positioned on grassland for the major portion of the investigation. In addition, a model of one of the helicopters, scaled one-to-one, was equipped with the appropriate aerials and could be raised some 10 metres above ground level on a fibre-glass pole. Measurements made using this model were useful in assessing the effects of the ground, especially on those aerials mounted on the lower surfaces of the fuselage.

In total some 120 records were obtained of the isolation between selected pairs of aerials for both the in-band and out-of-band conditions.

3 Laboratory Investigations

Laboratory measurements were made using monopole aerials and broad-band commercially available blade aerials mounted both on ground planes constructed from rivetted aluminium sheets, and on a cylinder 1·12 m in diameter by 2·44 m long (Fig. 3). Isolation was measured between aerials both in-band and out-of-band. Unwanted coupling effects were avoided by operating the associated instrumentation under the ground plane or inside the cylinder, as appropriate.

3.1 Impedance Mismatch Effects

The designed insertion loss of an attenuator pad of defined impedance, say 50Ω, is obtained when the pad is included in a 50Ω system. Only one port, source or load, needs to be 50Ω for this to apply, but where neither source nor load are 50Ω this is no longer true. (See Table 1.)

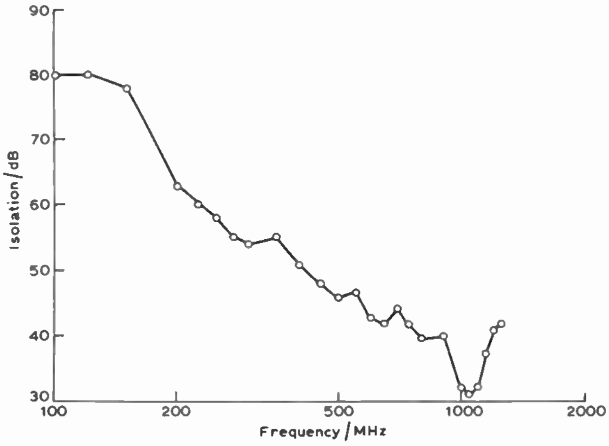
Table 1

Insertion loss of 6-dB 50-Ω pad

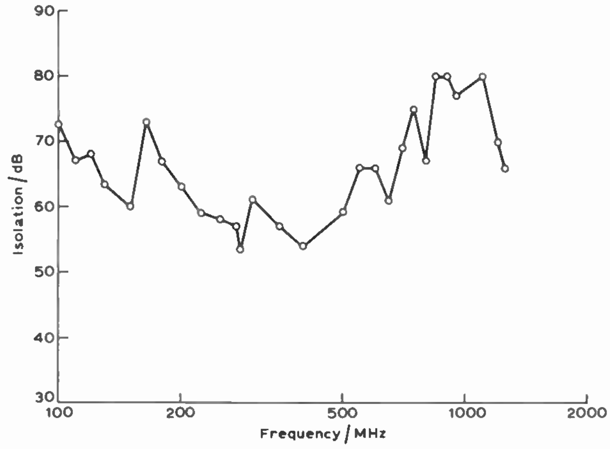
Source Z_1 ohms	Load Z_2 ohms	Insertion loss dB
50	50	6
50	1	6
1	50	6
1000	1	2·3
5	5	14
5+j20	5−j20	15

Values calculated assume direct connections, source-pad-load (i.e. no interposed transmission lines).

If a power oscillator or transmitter is used as a source then, irrespective of the impedance of either of these at

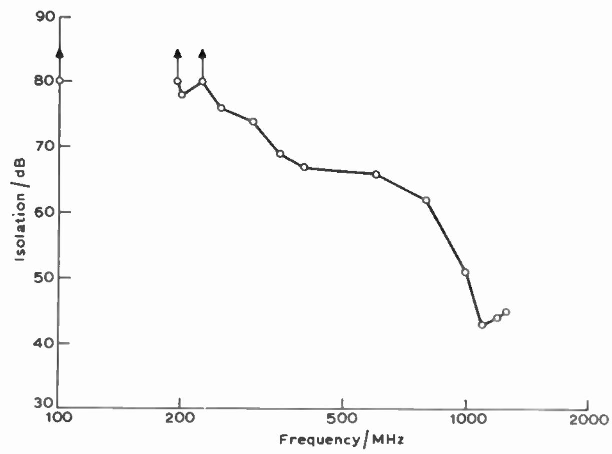


(a) Helicopter 1,



(b) Helicopter 2,

(c) Fixed wing aircraft 1,



(d) Fixed wing aircraft 2.

Fig. 2. Typical u.h.f.—L band aerial isolation.

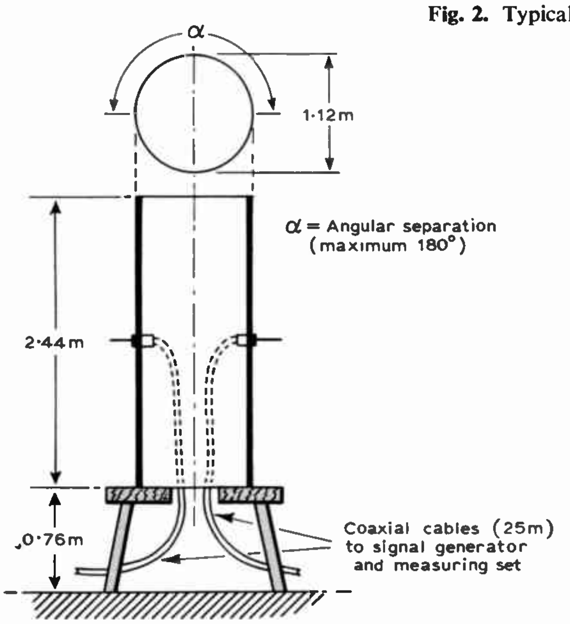


Fig. 3. Aluminium cylinder on wooden table.

fundamental frequencies f_1, the source impedance at any harmonic frequencies will not generally be 50Ω. In addition, that impedance will be transformed by the inclusion of cabling to some other (generally non-50Ω) impedance. The impedance of the transmitter aerial will itself not generally be 50Ω at harmonic frequencies, which are out-of-band for it designed operating range. The insertion of an attenuator between source and aerial will not, therefore, provide the designated attenuation. It follows that the insertion of an attenuator in the system can be used to indicate the non-equivalence of source and load impedance to the pad impedance.

Measurements of aerial isolation made at harmonic frequencies using a power oscillator as source have been found to exhibit variations in excess of 20 dB, depending on frequency and length of transmission line between source and transmitting aerial.

In the same test configuration the maximum variation was reduced to 5 dB by the inclusion of a 6 dB attenuator, thus illustrating the buffering action obtainable. From the foregoing it is apparent that correlation between signal generator isolation measurements and those in-band measurements using a transmitter/power oscillator source should be good. A bench measurement (into 50Ω), of the harmonic content of a transmitter output, if combined with a signal generator measurement of aerial-to-aerial isolation, can give rise to an underestimation of received signal level at unwanted harmonic frequencies which is indefinite in magnitude. A measurement in one system indicated a 19 dB error at a particular frequency. A filter interposed between transmitter and aerial capable of providing a substantially constant impedance at both fundamental and harmonic frequencies is therefore required. Under these conditions a bench measurement (at 50Ω) of harmonic amplitude at the filter output will permit reliable calculation of received signal levels using aerial-to-aerial isolation data obtained via signal generator measurements.

An absorption filter satisfying these requirements has been developed by a manufacturer for one particular frequency band. Constant *k* or *m*-derived filters are often ineffective because the impedance characteristics in the rejection band are not as required.

3.2 Harmonic Generation Effects†

In experiments with a power oscillator for which the third-harmonic content measured into 50Ω was 50 dB down on the fundamental level, an attempt was made to induce harmonic generation effects in a system consisting of two aircraft blade-aerials mounted on a ground plane made from rivetted aluminium sheets.

According to Landt,[2] third-harmonic signal levels due to non-linear junctions may be expected to increase as the third power of the transmitter output level. However, when operating the system referred to above at three frequencies in the v.h.f. band, and varying the power input to the transmitting aerial from 5 to 50 W (i.e. 10 dB ratio) at each fundamental frequency, the increase in received level at third harmonic frequencies did not exceed 10 dB.

The transmitting aerial base was then loosened and silicon carbide powder was introduced between base and ground-plane. Random variations of 1 to 2 dB in received signal level were obtained. It is considered that these small variations were caused by changes of aerial impedance affecting its matching to the source impedance at the harmonic frequency of the power oscillator.

It was evident that harmonic generation effects were not measurable under these conditions, as harmonics emanating from the power oscillator greatly exceeded in level those generated by non-linear junctions in the aerial-ground plane systems.

Nevertheless, it has been found possible to demonstrate the existence of non-linear junction effects at 50 Hz for near-zero force contacts. The circuit arrangement is shown in Fig. 4.

† This Section excludes the effects of intermodulation products from multiple signal sources.

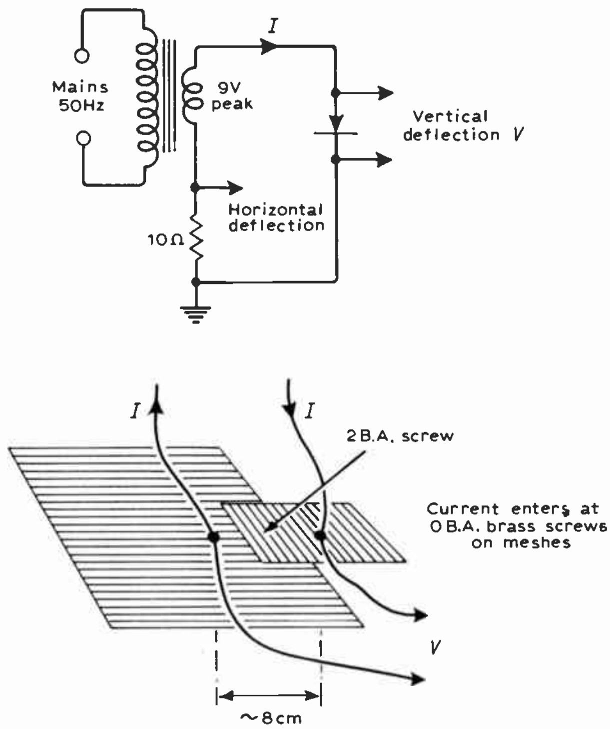


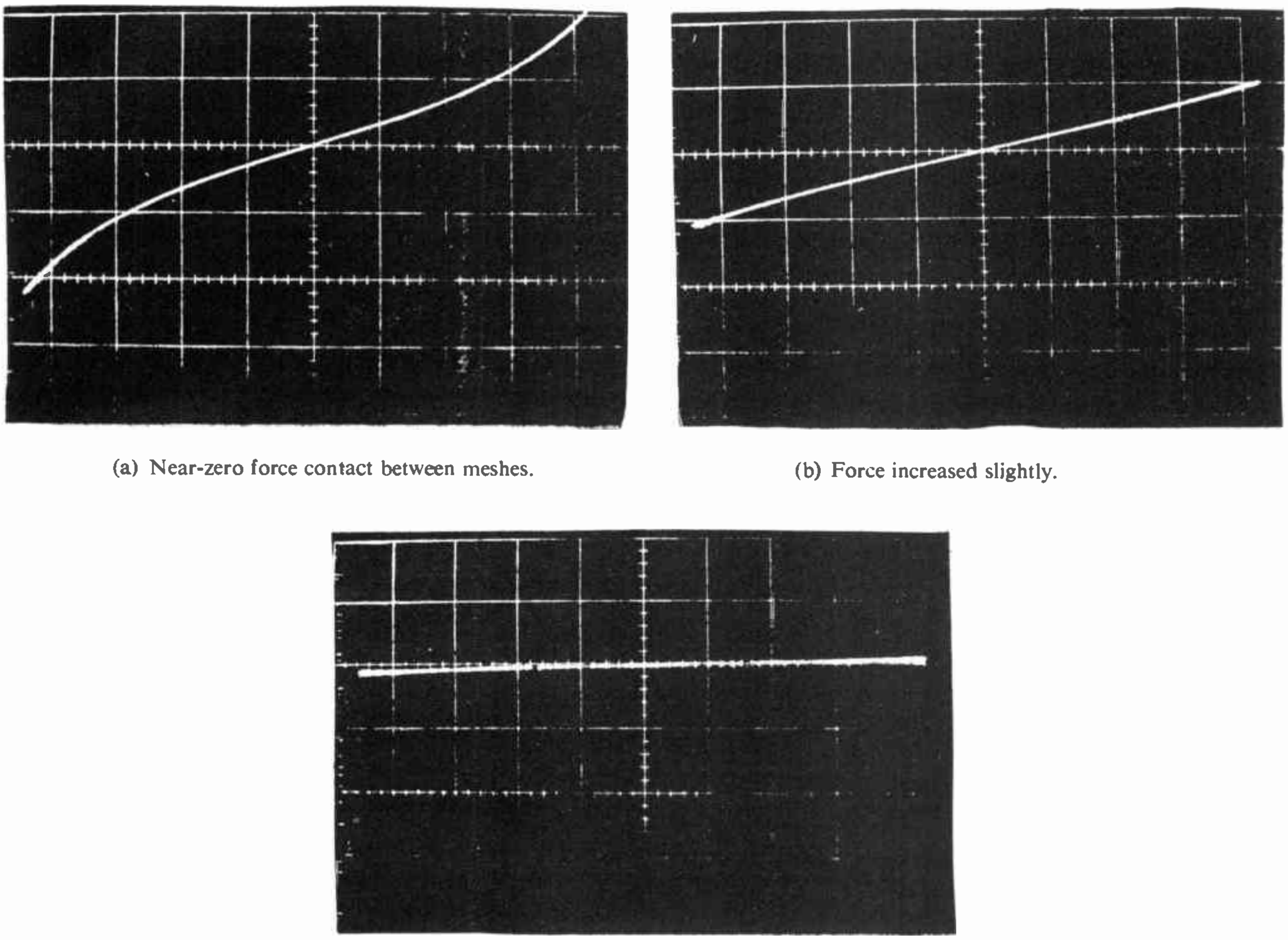
Fig. 4. Test arrangement for obtaining voltage-current characteristics of contacts.

V–I relationships for aluminium mesh sheets and the variation with contact pressure can be seen in Fig. 5. Effects noted when testing in an aluminium mesh cage, where harmonic generation in the cage walls was detected, are consistent with such *V–I* characteristics.

It should be noted, however, that components exhibiting harmonic generation effects at, for example, u.h.f., may not always display these effects at audio frequencies.

3.3 Transmitter Screening

The absorption filter previously referred to as providing 50Ω input and output impedances both in and out-of-band was found to have a rejection ratio exceeding 60 dB in the stop-band when tested in the laboratory. This filter, however, had been user-tested in the field with disappointing results. The apparent failure was found to be largely attributable to the effects of alternative paths of coupling in the system under test, mainly as a result of poor transmitter screening and inadequate attention to the r.f. emissions from the associated cabling. Screening improvements allowed the full performance of the filter to be demonstrated. The full performance was, however, not maintained, because of a non-linear effect which was subsequently identified as unsatisfactory contact in r.f. connector centre pins. These centre pins carry the full aerial current over relatively small contact areas and are, therefore, the most likely sources of harmonic generation.



(a) Near-zero force contact between meshes.

(b) Force increased slightly.

(c) Finger-tight 2BA brass screw and nut connects meshes.

Fig. 5. Voltage-current characteristics of contacts between aluminium meshes. Horizontal axis: current 0·2 A/cm; Vertical axis: Voltage 0·1 V/cm (except (c) 0·005 V/cm).

The application of an aerosol cleaning fluid incorporating a lubricant to all r.f. connector surfaces was found to eliminate this problem, but the permanency of this treatment has not been evaluated.

As an example of the effectiveness of these arrangements the received level at third harmonic frequency with a u.h.f. transmitter coupled to a co-sited aerial at third harmonic frequency may be expected to range from -18 to < -78 dBm without filter or lubricant, but could improve to -78 to < -138 dBm after treatment.

4 Empirical Formulae

A number of authors have produced schemes for the calculation of aerial-to-aerial isolation. These have generally been based on mathematical treatments such as those for diffraction of r.f. waves around a cylinder and at wing edges, but are not, in our experience, capable of being applied to the types of problems discussed in this paper. Such programs, e.g. ATACAP[3] (McDonnell-Douglas), ANTISO2[4] (Boeing) and AVPAK[5] (ECAC), are not readily applicable to out-of-band conditions and the methods of application to rotating wing aircraft are not described.

In view of these limitations an examination of the possibility of an empirical approach was initiated, based to a large extent on the laboratory and aircraft measurements.

A basic problem is the assessment of isolation between aerials sited around a curved surface where theory is somewhat restrictive in terms of physical dimensions and wavelength. In order to test the feasibility of deriving a formula for isolation between aerials sited on a curved surface, an aluminium cylinder was arranged as shown in Fig. 3. Sets of resonant $\lambda/4$ monopoles were installed as indicated, at angular separations of 180, 120 and 90°, for the range 100 to 1000 MHz (10 sets of aerials). Isolation was measured and the results are shown for aerials at an angular separation of 180° in Fig. 6.

Now Friis[7] has derived a simple transmission formula for a radio circuit made up of a transmitting aerial and a

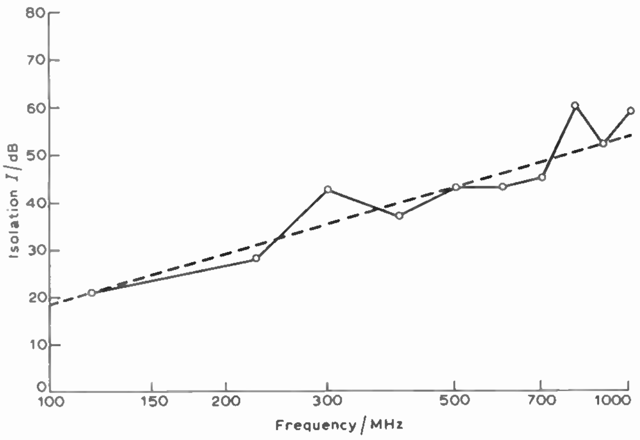


Fig. 6. Isolation between monopole pairs mounted on aluminium cylinder. Aerial pairs have appropriate $\lambda/4$ lengths at each test frequency.

receiving aerial in free space namely

$$P_r/P_t = A_r A_t / D\lambda^2 \tag{1}$$

where

P_t = power supplied to transmitting aerial
P_r = power available from receiving aerial
A_t = effective area of the transmitting aerial
A_r = effective area of the receiving aerial
D = distance between aerials
λ = wavelength

For isotropic transmitting and receiving aerials

$$A_t = A_r = \lambda^2/4\pi$$

and

$$\frac{P_r}{P_t} = \left(\frac{\lambda}{4\pi D}\right)^2 \tag{2}$$

In logarithmic terms the transmission loss or isolation I in dB becomes:

$$I = 20 \log f + 20 \log D - 28 \tag{3}$$

with f in MHz and D in metres.

In Fig. 6, a straight line (shown dashed) was fitted to the results and can be represented by the expression

$$I = 20 \log kf^n + 20 \log D - 28 \tag{4}$$

where k and n are constants (dimensionless), and D is now the separation of the aerials over the cylinder surface (metres).

Note that if $k = n = 1$ we have the classic free-space transmission formula of Friis, shown in (3) above.

For the particular cylinder of Fig. 6 (radius $a = 0{\cdot}56$ m)

$$k = 3{\cdot}7 \times 10^{-2}$$

and

$$n = 1{\cdot}75$$

It has been found possible to accommodate any radius of cylinder by writing k in terms of a.

From the data above $k = a/15$ and

$$I = 20 \log \frac{a}{15} f^{1\cdot75} + 20 \log D - 28 \tag{5}$$

Further measurements made with aerial pairs mounted on the aluminium cylinder established that equation (5) was also valid for aerials displaced by 120° from each other, but for displacements of 60 and 90° equation (3) gave better agreement with measured values.

The expression must now be modified to allow for the in-band/out-of-band status of aerials at the frequency of test and hence becomes:

$$I = 20 \log \frac{a}{15} f^{1\cdot75} + 20 \log D - 28 + (SA) \tag{6}$$

where (SA) is a factor derived from the work of Siarkiewicz and Adams[8] (Fig. 7). For example, using Fig. 7, if both transmitting and receiving aerials have a length of $0{\cdot}2\lambda$, the factor (SA) in the empirical formula is 2×5 dB = 10 dB. For the purposes of the formulae, when aerial length exceeds $\frac{1}{2}\lambda$ the right-hand side of the curve of Fig. 7 is assumed to cycle between 0 and 8 dB. For example, aerial length $\frac{3}{4}\lambda$ gives 0 dB, $0{\cdot}8\lambda$ gives 3 dB and λ gives 8 dB etc.

The calculation for a wide-band aerial at a frequency *below* its lowest specified working frequency is made using that lowest frequency as a datum, e.g. the effective aerial length at 50 MHz for an aerial of working band 225–400 MHz is

$$\frac{\lambda}{4} \times \frac{50}{225}.$$

At frequencies *above* the highest specified working frequency, however, overall agreement between calculated and measured isolation is improved if mid-band frequency is used as a datum. For the example given above the mid-band frequency is $225 + 400/2 = 312{\cdot}5$ MHz, and, at a

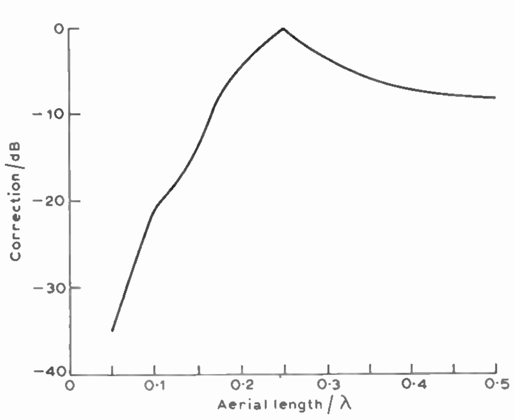


Fig. 7. Aerial length factor (after Siarkiewicz and Adams[8]).

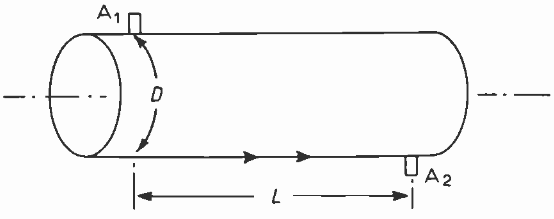


Fig. 8. Distance assessment for calculation of isolation. Where tapering occurs, choose curved path starting at aerial on larger radius of curvature.

test frequency of 800 MHz the effective length of the aerial is

$$\frac{\lambda}{4}\times\frac{800}{312\cdot5}.$$

The formula for isolation of a curved surface has been found to accommodate linear displacement L by merely changing the second term in equation (6), $20\log D$, to $20\log(D+L)$ (see Fig. 8), then

$$I=20\log\frac{a}{15}f^{1\cdot75}+20\log(D+L)-28+(SA) \quad (7)$$

As already stated these formulae have been found to be satisfactory for angular displacements of 180 and 120° However, at 90° (and smaller angles), putting $k=n=1$ gives closer agreement with measurements, and thus equation (6), for example, becomes

$$I=20\log f+20\log D-28+(SA).$$

An additional expression for line-of-sight aerials with some shading by obstruction has been evolved, but to the present, lacks extensive validation. The equation is:

$$I=20\log f+\sec\frac{\theta}{4}\times[20\log(D_1+D_2)]-28+(SA) \quad (8)$$

where D_1 and D_2 are linear distances (Fig. 9).

Equations (6) and (7) have been validated by the examination of 83 aerial combinations on two helicopters and two fixed wing aircraft. Only a limited evaluation of equation (8) has been possible using a single aircraft and two aerial pairs.

The formulae were tested at frequencies of 50, 100, 200, 400 and 1000 MHz. Fuselage radii ranged from 0·23 to 2·13 m and aerial separations from 0·18 to about 15 m. Aerial-to-aerial isolation cannot be calculated with exactitude and it is desirable that any formulae used in such calculations should tend, on balance, to underestimate the magnitude of this parameter thus producing a margin of safety.

With the exception of those at 50 and 100 MHz the results approximate to the normal distribution law. These formulae have been shown to yield calculated values of aerial isolation which are within, at worst, +9 dB to −16 dB (the negative sign indicating an underestimation of isolation), with about 70% certainty. They also have the distinct advantage of being capable of dealing with both in-band and out-of-band situations, and of being able to handle a wider range of geometrical layouts than those previously described.

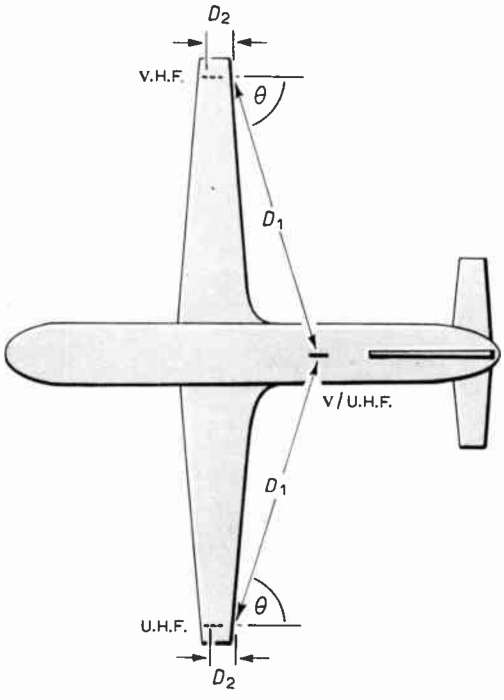


Fig. 9. Illustrating isolation including wing shading. In the plan of aircraft viewed from above it should be noted that the v.h.f. and u.h.f. aerials mounted under wings are not in line-of-sight with v./u.h.f. aerial mounted on top of fuselage.

The ATACAP[3] procedure is claimed to calculate aerial-to-aerial interference to within 10 dB. Gardner[6] *et al.* report 8 and 10 dB accuracies for measurements made on two aircraft, and the AVPAK[5] method has a reported standard deviation of 5·5 dB.

5 Conclusions

It can be stated that:

(i) A suitably filtered transmitter can have a ratio of third harmonic to fundamental power level of −90 dB or better.

(ii) Centre-pins of connectors and other areas where high surface current densities exist are a likely region of harmonic generation, and lubrication is essential.

As an example, consider a u.h.f. transmitter coupling to a co-sited aerial at third harmonic frequency. The received signal level may be expected to range from:

−18 to < −78 dBm without filter or lubricant, but could improve to −78 to < −138 dBm after treatment.

(iii) Aerial-to-aerial isolation may be calculated using the following formulae:

$$\left.\begin{array}{l} I=20\log f+20\log D-28+(SA) \\ I=20\log f+20\log (D+L)-28+(SA) \\ \text{Aerials with angular separation } \alpha<120^\circ \end{array}\right\}$$

$$\left.\begin{array}{l} I=20\log \frac{a}{15} f^{1\cdot 75}+20\log D-28+(SA) \\ I=20\log \frac{a}{15} f^{1\cdot 75}+20\log (D+L)-28+(SA) \\ \text{Aerials with angular separation } \alpha\geqslant 120^\circ \end{array}\right\}$$

$$\left.\begin{array}{l} I=20\log f+\sec\frac{\theta}{4}\times[20\log (D_1+D_2)]-28+(SA) \\ \text{Wing shading (see Fig. 9).} \end{array}\right\}$$

6 Acknowledgment

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Indexing Terms: Electric networks, active, Network synthesis, Mathematical transformations

The synthesis of low-pass active RC ladder networks by the recurrent-continuant method

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SUMMARY

A new procedure for the synthesis of low-pass active RC ladder networks by the recurrent-continuant transformation has been proposed. The transformation scheme consists of simple determinant operations which are carried out subsequent to a frequency transformation and addition of certain parameters to the prescribed all-pole transfer function. The proposed method is more general than those described earlier, and simple as compared with the synthesis procedures based on coefficient matching methods. The method has been illustrated by realizing a number of 3rd and 4th-order characteristic functions. The proposed scheme, with slight modification, is also capable of realizing double ladder networks. The modified scheme has been described with an appropriate example in the third order. The capacitors in all the realized networks have been constrained to be unity. The methods yield multiple sets of values of the network elements and may be used with advantage to obtain optimized networks.

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1 Introduction

The recurrent to continuant transformation method is an elegant and versatile tool for the synthesis or analysis of ladder networks, and a number of transformation procedures have been described by various authors[1–4] for the realization of the low-pass transfer functions by different types of networks e.g. passive lossless and active RC (ARC) ladder networks. In the case of ARC ladder networks the relevant transformation procedure[4] requires a prior assumption of the values of the passive elements, and the procedure in itself does not provide the way in which such an assumption can be made. This limitation has been removed in this paper by proposing a new recurrent-continuant transformation procedure. The procedure has been developed from the one described previously by the authors[3] in the context of passive RC ladders, and consists of frequency transformations and a number of determinant operations. The method is capable of yielding a large number of sets of element values and may be employed to adjust the values of the sum or spread of any particular type of element. In view of the suitability of equal valued capacitor (EVC) ARC structures for realization in microelectronic form, the values of the capacitors have been constrained to be unity in the proposed transformation. The method has been illustrated explicitly by realizing the 3rd and 4th-order EVC-ARC networks. A technique for the realization of double ladder networks through recurrent-continuant methods has also been described and illustrated.

2 Synthesis of EVC-ARC Networks

2.1 The Active Continuant and the Recurrent

The ARC low-pass ladder network proposed to be synthesized is shown in Fig. 1. The reciprocal of the voltage transfer function $V_{\text{in}}/V_{\text{out}} = P_n(s)$ of this network may be described by the following active continuant of order $2n$.

$$\frac{V_{\text{in}}}{V_{\text{out}}} = P_n(s)$$

$$= \begin{vmatrix} R_1 & 1 & 0 & . & 0 & 0 & 0 \\ -1 & C_1 s & 1 & . & 0 & 0 & -A_1 C_1 s \\ 0 & -1 & R_2 & . & 0 & 0 & 0 \\ . & . & . & . & . & . & . \\ 0 & 0 & 0 & . & C_{n-1}s & 1 & -A_{n-1}C_{n-1}s \\ 0 & 0 & 0 & . & -1 & R_n & 1 \\ 0 & 0 & 0 & . & 0 & -1 & C_n s \end{vmatrix} \tag{1}$$

Under EVC constraint $C_i = 1$ ($i = 1, 2, 3, \ldots, n$).

The prescribed low-pass characteristic may be described by an all-pole transfer function $H(s)$:

$$H(s) = \frac{1}{P_n(s)} = \frac{1}{\sum_{i=1}^{n} a_i s^i + 1} \tag{2}$$

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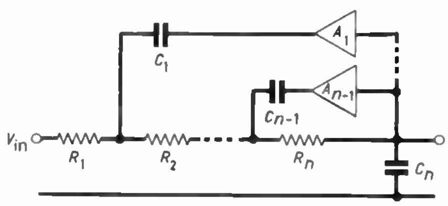

Fig. 1. ARC low-pass ladder network.

where $a_i > 0$. Using the transformation $s = y^2$ and introducing terms $Rb_{2i-1}y^{2i-1}$ $(i = 1, 2, \ldots, n)$ in the denominator of relation (2), $H(s)$ modifies to $H'(y)$ of the following form:

$$H'(y) = \frac{1}{P'_n(y)} = \frac{1}{\sum_{i=1}^{2n} a'_i y^i + 1} \tag{3}$$

where $a'_{2i} = a_i$ and $a'_{2i-1} = Rb_{2i-1}$ $(i = 1, 2, 3, \ldots, n)$, $H'(y)$ equals $H(s)$ when $R = 0$. $P'_n(y)$ may now be written as the following modified recurrent of order $2n+1$:

$$P'_n(y) = \begin{vmatrix} a_n & Rb_{2n-1} & a_{n-1} & . & a_2 & Rb_1 & 1 \\ -1 & y & 0 & . & 0 & 0 & 0 \\ 0 & -1 & y & . & 0 & 0 & 0 \\ . & . & . & . & . & . & . \\ 0 & 0 & 0 & . & y & 0 & 0 \\ 0 & 0 & 0 & . & -1 & y & 0 \\ 0 & 0 & 0 & . & 0 & -1 & y \end{vmatrix} \tag{4}$$

The synthesis technique must secure the transformation of the modified recurrent (relation 4) into the active continuant (relation 1). To establish an equality between $P_n(s)$ (relation 1) and $P'_n(y)$ (relation 4), the transformation procedure must incorporate reverse frequency transformation $y^2 = s$ and R must also vanish.

2.2 The Transformation Procedure

For the sake of brevity the following notations have been used for describing the various determinant operations used in the transformation procedure.

$R_{i,j}(q)$: Multiply ith row by q and add it to the jth row.

$C_{i,j}(q)$: Multiply ith column by q and add it to the jth column.

$M_j^i(q)$: Multiply ith row and divide jth column by q.

$d_{i,j}$: The element in the ith row and jth column of the determinant.

To transform the modified recurrent (relation 4) into the active continuant (relation 1) Holbrook's transformation procedure[1] is initially adopted with the following modifications:

(i) The term $d_{1,2}$ is made R instead of unity by the operation $M_2^3(b_{2n-1})$

(ii) The terms $d_{2i+1,2n+1}$ $(i = 1, 2, \ldots, n-1)$ are allowed to survive in the form $(C_i - m_i)y$ (m_i is real number positive or negative) instead of their being made zero during the $R_{i,j}(q)$ operations.

Subsequent to these operations the determinant is made an even-ordered one by the operation $C_{1,2}(-R)$ followed by an expansion of the resulting determinant along the first row.

With these operations, the determinant (4) takes the form

$$P'_n(y) = \begin{vmatrix} R+R_1y & 1 & 0 & . & 0 & 0 & 0 \\ -1 & C_1y & 1 & . & 0 & 0 & -B_1y \\ 0 & -1 & R_2 & . & 0 & 0 & 0 \\ . & . & . & . & . & . & . \\ 0 & 0 & 0 & . & C_{n-1}y & 1 & -B_{n-1}y \\ 0 & 0 & 0 & . & -1 & R_ny & 1 \\ 0 & 0 & 0 & . & 0 & -1 & C_ny \end{vmatrix} \tag{5}$$

Finally R is dropped, and a chain of operations $M_j^i(y)$ $(i = 1, 2, 3, \ldots, 2n)$ followed by a frequency transformation $y^2 = s$ is carried out.

With these steps, $H'(y)$ reduces to $H(s)$ and determinant (5) takes the form of familiar active continuant (relation 1). The diagonal elements R_i and C_i $(i = 1, 2, 3, \ldots, n)$ of the determinant (5) represent the capacitive and resistive elements respectively of the proposed network, and $B_i/C_i = A_i$ $(i = 1, 2, 3, \ldots, n-1)$ give the gains of the amplifier blocks. Obviously R_i and C_i $(i = 1, 2, 3, \ldots, 2n)$ must be positive and further to satisfy the EVC constraint C_i $(i = 1, 2, 3, \ldots, n)$ must be unity. These conditions are met with by an appropriate choice of the constants b_i and m_i.

It is to be pointed out that none of the aforesaid operations necessitates a knowledge of the numerical value of R and hence the transformation is carried out by assuming only the values of the constants b_i and m_i. Corresponding to a prescribed set of values of a_i, different sets of values of b_{2i-1} and also m_i may always be found which satisfy the desired requirements. Consequently the technique will yield multiple sets of element values.

3 Illustration

3.1 The Third-order Network

Let

$$H(s) = \frac{1}{P_n(s)} = \frac{1}{\sum_{i=1}^{3} a_i s^i + 1} \tag{6 a}$$

where $a_i > 0$. Substitution of $s = y^2$ in the above yields

$$H(y)=\frac{1}{P_3(y)}=\frac{1}{\sum_{i=1}^{3} a_i y^{2i}+1} \qquad (6\ b)$$

Now $H(y)$ modifies to $H'(y)$ of the following form (relation 3a)

$$H'(y)=\frac{1}{P'_3(y)}=\frac{1}{\sum_{i=1}^{6} a'_i y^{i}+1} \qquad (6\ c)$$

with $a'_{2i}=a_i$ and $a'_{2i-1}=Rb_{2i-1}$ for $i=1, 2, 3$. As before $H'(y)$ equals $H(s)$ when $R=0$. $P'_3(y)$ may be written as the following modified recurrent (relation 4)

$$P'_3(y)=\begin{vmatrix} a_3 & Rb_5 & a_2 & Rb_3 & a_1 & Rb_1 & 1 \\ -1 & y & 0 & 0 & 0 & 0 & 0 \\ 0 & -1 & y & 0 & 0 & 0 & 0 \\ 0 & 0 & -1 & y & 0 & 0 & 0 \\ 0 & 0 & 0 & -1 & y & 0 & 0 \\ 0 & 0 & 0 & 0 & -1 & y & 0 \\ 0 & 0 & 0 & 0 & 0 & -1 & y \end{vmatrix} \qquad (7)$$

Now carrying out the series of operations:

(i) $M_1^{\,2}(a_3)$, $M_2^{\,3}(b_5)$,

(ii) $C_{1,3}(-a_2)$, $C_{1,5}(-a_1)$, $C_{1,7}(-1)$, $C_{2,4}(-b_3)$, $C_{2,6}(-b_1)$,

(iii) $R_{4,2}\left(\frac{b_3a_3}{b_5}\right)$, $R_{6,2}\left(\frac{b_1a_3}{b_5}\right)$, $M_3^{\,4}(q_1)$,

where $q_1=a_2-\frac{b_3a_3}{b_5}$,

(iv) $C_{3,5}(q_2)$, $C_{3,7}(-1)$,

where $q_2=a_1-\frac{b_1a_3}{b_5}$,

(v) $R_{5,3}(q_2)$, $R_{7,3}(m_1)$, $M_4^{\,5}(b_3-q_2)$,

$R_{6,4}\left\{\frac{q_1(b_1-m_1)}{(b_3-q_2)}\right\}$, $M_5^{\,6}\left\{q_2-\frac{q_1(b_1-m_1)}{(b_3-q_2)}\right\}$,

(vi) $C_{5,7}(-1)$,

(vii) $R_{7,5}(m_2)$, $M_6^{\,7}(b_1-m_1-m_2)$,

(viii) $C_{1,2}(-R)$,

the determinant (7) reduces to the determinant

$$P'_3(y)=\begin{vmatrix} R+R_1y & 1 & 0 & 0 & 0 & 0 \\ -1 & C_1y & 1 & 0 & 0 & -(C_1-m_1)y \\ 0 & -1 & R_2y & 1 & 0 & 0 \\ 0 & 0 & -1 & C_2y & 1 & -(C_2-m_2)y \\ 0 & 0 & 0 & -1 & R_3y & 1 \\ 0 & 0 & 0 & 0 & -1 & C_3y \end{vmatrix} \qquad (8)$$

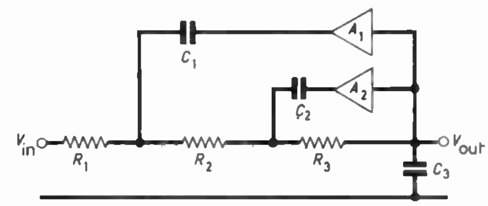

Fig. 2. The third-order active ladder network.

Putting $R=0$ and adopting the frequency transformation $y^2=s$ subsequent to a simultaneous multiplication and division of even rows and odd columns by y, the following active continuant results:

$$P_3(s)=\begin{vmatrix} R_1 & 1 & 0 & 0 & 0 & 0 \\ -1 & C_1s & 1 & 0 & 0 & -(C_1-m_1)s \\ 0 & -1 & R_2 & 1 & 0 & 0 \\ 0 & 0 & -1 & C_2s & 1 & -(C_2-m_2)s \\ 0 & 0 & 0 & -1 & R_3 & 1 \\ 0 & 0 & 0 & 0 & -1 & C_3s \end{vmatrix} \qquad (9)$$

The network corresponding to this active continuant is shown in Fig. 2.

The network elements R_i, C_i and A_i are related to the polynomial coefficients a_i and the arbitrary constants b_i and m_i as follows:

(i) $R_1=\frac{a_3}{b_5}$,

(ii) $R_2=\frac{q_1}{b_3-q_2}$,

(iii) $R_3=\frac{q_2(b_3-q_2)-q_1(b_1-m_1)}{b_3-q_2}$,

(iv) $C_1=\frac{b_5}{q_1}$,

(v) $C_2=\frac{(b_3-q_2)^2}{q_2(b_3-q_2)-q_1(b_1-m_1)}$,

(vi) $C_3=b_1-m_1-m_2$,

(vii) $A_1=C_1-m_1$,

(viii) $A_2=C_2-m_2$.

The capacitors $C_i(i=1, 2, 3)$ become unit valued and resistors $R_i(i=1, 2, 3)$ are positive under the following restrictions:

(i) $b_3=\frac{b_5(a_2-b_5)}{a_3}$,

(ii) $b_1>\frac{b_5(a_1-b_3)}{a_3}$,

Table 1. The element values for third-order EVC-ARC filters

Parameters	Butterworth			Chebyshev ½ dB ripple			Chebyshev 1 dB ripple		
b_5	1·000	0·5000	0·80000	0·75000	0·80000	0·75000	1·40000	2·00000	0·70000
b_3	1·000	0·7500	0·96000	0·53699	0·54416	0·53699	0·42071	0·01200	0·45109
b_1	2·000	0·9870	1·11303	1·23203	1·30593	1·00399	2·07494	0·09336	0·84056
m_1	3·000	2·0000	1·00000	2·00000	2·00000	1·00000	3·00000	3·50000	1·00000
m_2	−2·000	−2·0130	−0·88700	−1·76797	−1·69406	−0·99609	−1·92506	−1·40663	−1·15944
R_1	1·000	2·0000	1·25000	1·86220	1·74581	1·86220	1·45385	1·01833	2·90769
R_2	1·000	0·6902	2·27732	1·09103	1·17609	2·85420	1·52717	3·12167	1·86894
R_3	1·000	0·7245	0·35129	0·68744	0·68023	0·26273	0·91672	0·64072	0·37454
Resistance sum	3·000	3·4146	3·87861	3·64067	3·60213	4·97923	3·89775	4·78072	5·15118
A_1	−2·000	−1·0000	0·00000	−1·00000	−1·00000	0·00000	−2·00000	−2·50000	0·00000
A_2	3·000	3·0127	1·88700	2·76797	2·69406	1·94609	2·92506	2·40663	2·15944

(iii) $$b_1=\frac{-A\pm\sqrt{(A^2-BC+BDm_1)}}{4a_3{}^2/b_5{}^2},$$

where

$$A=\frac{2a_3b_3}{b_5}-\frac{4a_1a_3}{b_5}+a_2;$$

$$B=\frac{8a_3{}^2}{b_5{}^2};$$

$$C=b_3{}^2-3a_1b_3+2a_1{}^2;$$

and $$D=a_2-\frac{b_3a_3}{b_5}=b_5,$$

(iv) $m_2=b_1-m_1-1\cdot$

The amplifier gains are then given by

$$A_1=1-m_1 \quad\text{and}\quad A_2=1-m_2. \tag{10}$$

Some of the different sets of the values of the network elements obtainable from this scheme for different polynomials are given in Table 1.

It is to be noted that the above restrictions are also satisfied by $m_1=1$. This reduces the amplifier gain A_1 in eqn. (10) to zero. In such a case the capacitor C_1 becomes grounded. Thus the scheme is also capable of realizing a third-order EVC–ARC network with single amplifier A_2.

3.2 The Fourth-order Network

Let

$$H(s)=\frac{1}{P_4(s)}=\frac{1}{\sum_{i=1}^{4}a_is^i+1} \tag{11}$$

where $a_i>0$. Using the frequency transformation $s=y^2$ and adding the terms $Rb_{2i-1}y^{2i-1}$ with $i=1, 2, 3, 4$ in the denominator of (11) we get

$$H'(y)=\frac{1}{P'_4(y)}=\frac{1}{\sum_{i=1}^{8}a'_iy^i+1} \tag{12}$$

with $a'_{2i}=a_i$ and $a'_{2i-1}=Rb_{2i-1}$ for $i=1, 2, 3, 4$. $H'(y)$ equals $H(s)$ when $R=0$. $P'_4(y)$ may now be written as the following modified recurrent:

$$P'_4(y)=\begin{vmatrix} a_4 & Rb_7 & a_3 & Rb_5 & a_2 & Rb_3 & a_1 & Rb_1 & 1\\ -1 & y & 0 & 0 & 0 & 0 & 0 & 0 & 0\\ 0 & -1 & y & 0 & 0 & 0 & 0 & 0 & 0\\ 0 & 0 & -1 & y & 0 & 0 & 0 & 0 & 0\\ 0 & 0 & 0 & -1 & y & 0 & 0 & 0 & 0\\ 0 & 0 & 0 & 0 & -1 & y & 0 & 0 & 0\\ 0 & 0 & 0 & 0 & 0 & -1 & y & 0 & 0\\ 0 & 0 & 0 & 0 & 0 & 0 & -1 & y & 0\\ 0 & 0 & 0 & 0 & 0 & 0 & 0 & -1 & y \end{vmatrix} \tag{13}$$

This determinant is now subjected to the following series of steps:

(i) $M_1{}^2(a_4)$, $M_2{}^3(b_7)$,

(ii) $C_{1,3}(-a_3)$, $C_{1,5}(-a_2)$, $C_{1,7}(-a_1)$, $C_{1,9}(-1)$, $C_{2,4}(-b_5)$, $C_{2,6}(-b_3)$, $C_{2,8}(-b_1)$;

(iii) $R_{4,2}\left(\frac{a_4b_5}{b_7}\right)$, $R_{6,2}\left(\frac{a_4b_3}{b_7}\right)$, $R_{8,2}\left(\frac{a_4b_1}{b_7}\right)$, $M_3{}^4(q_3)$;

where

$$q_3=a_3-\frac{b_5a_4}{b_7};$$

(iv) $C_{3,5}(-q_4)$, $C_{3,7}(-q_5)$, $C_{3,9}(-1)$

where

$$q_4=a_2-\frac{b_3a_4}{b_7}\quad\text{and}\quad q_5=a_1-\frac{b_1a_4}{b_7};$$

(v) $R_{5,3}\left(\frac{b_7q_4}{q_3}\right)$, $R_{7,3}\left(\frac{b_7q_5}{q_3}\right)$, $R_{9,3}(m_1)$, $M_4{}^5(b_5-q_4)$;

(vi) $C_{4,6}(q_5-b_3)$, $C_{4,8}(m_1-b_1)$;

(vii) $R_{6,4}\left\{\dfrac{q_3(b_3-q_2)}{b_5-q_1}\right\}$, $R_{8,4}\left\{\dfrac{q_3(b_1-m_1)}{b_5-q_4}\right\}$,

$$M_5{}^6\left\{\frac{q_4(b_5-q_4)-q_3(b_3-q_5)}{b_5-q_4}\right\};$$

(viii) $C_{5,7}(-q_6)$

where

$$q_6=\frac{q_5(b_1-m_1)+q_2(b_5-q_1)}{b_3-q_1};$$

(ix) $R_{7,5}(q_6)$, $R_{9,5}(m_2)$, $M_6{}^7(b_3-q_5-q_6)$, $R_{8,6}(q_7)$, $M_7{}^8(q_6-q_7)$, $R_{9,7}(m_3)$, $M_7{}^9(b_1-m_1-m_2-m_3)$

where

$$q_7=\frac{(b_5-q_1)(b_1-m_1-m_2)}{b_3-q_2-q_4}.$$

Finally, on the application of the step $C_{1,2}(-R)$, the above determinant (13) takes the form:

$$P'_4(y)=\begin{vmatrix} R+R_1y & 1 & 0 & 0 & 0 & 0 & 0 & 0 \\ -1 & C_1y & 1 & 0 & 0 & 0 & 0 & -(C_1-m_1)y \\ 0 & -1 & R_2y & 1 & 0 & 0 & 0 & 0 \\ 0 & 0 & -1 & C_2y & 1 & 0 & 0 & -(C_2-m_2)y \\ 0 & 0 & 0 & -1 & R_3y & 1 & 0 & 0 \\ 0 & 0 & 0 & 0 & -1 & C_3y & 1 & -(C_3-m_3)y \\ 0 & 0 & 0 & 0 & 0 & -1 & R_4y & 1 \\ 0 & 0 & 0 & 0 & 0 & 0 & -1 & C_4y \end{vmatrix} \qquad (14)$$

R is now dropped and a series of operations $M_{2i-1}{}^{2i}(y)$ ($i=1, 2, 3, 4$) are carried out. A frequency transformation $y^2=s$ yields the required active continuant:

$$P_4(s)=\begin{vmatrix} R_1 & 1 & 0 & 0 & 0 & 0 & 0 & 0 \\ -1 & C_1s & 1 & 0 & 0 & 0 & 0 & -(C_1-m_1)s \\ 0 & -1 & R_2 & 1 & 0 & 0 & 0 & 0 \\ 0 & 0 & -1 & C_2s & 1 & 0 & 0 & -(C_2-m_2)s \\ 0 & 0 & 0 & -1 & R_3 & 1 & 0 & 0 \\ 0 & 0 & 0 & 0 & -1 & C_3s & 1 & -(C_3-m_3)s \\ 0 & 0 & 0 & 0 & 0 & -1 & R_4 & 1 \\ 0 & 0 & 0 & 0 & 0 & 0 & -1 & C_4s \end{vmatrix} \qquad (15)$$

The network corresponding to this active continuant is shown in Fig. 3.

The network elements R_i, C_i and A_i are related to the polynomial coefficients a_i and the arbitrary constants b_i and m_i as follows:

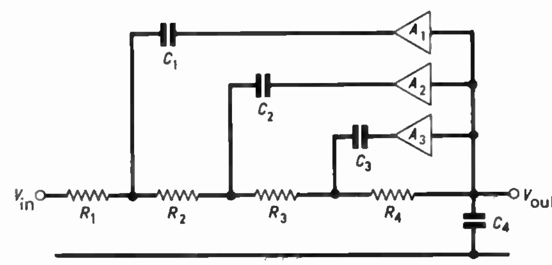

Fig. 3. The fourth-order ARC ladder networks.

(i) $R_1=\dfrac{a_4}{b_7}$,

(ii) $C_1=\dfrac{b_7}{q_5}$,

(iii) $R_2=\dfrac{q_3}{b_5-q_4}$,

(iv) $C_2=\dfrac{(b_5-q_4)^2}{q_4(b_5-q_4)-q_3(b_3-q_5)}$,

(v) $R_3=\dfrac{b_5-q_4}{b_3-q_5-k_1}$

where

$$k_1=\frac{q_5(b_5-q_4)-q_3(b_1-m_1)}{b_5-q_4}$$

(vi) $C_3=\dfrac{(b_5-q_5-k_1)^2}{k_1(b_3-q_5-k_1)-(b_5-q_4)(b_1-m_1-m_2)}$

(vii) $R_4=b_3-q_5-k_1$

(viii) $C_4=b_1-m_1-m_2-m_3$.

It follows from these relations that unit-valued capacitors and positive-valued resistors result under the following restrictions:

(i) $b_5=\dfrac{b_7(a_3-b_7)}{a_4}$ and $A^2>BC$

where

$$A=\frac{2a_4b_5}{b_7}-\frac{4a_2a_4}{b_7}+a_3;$$

$$B=8a_4{}^2/b_7{}^2;$$

$$C=b_5{}^2-3a_2b_5+2a_2{}^2-\left(a_3-\frac{b_5a_4}{b_7}\right)a_1$$

(ii) $\dfrac{a_2b_7}{a_4}>b_3>\dfrac{b_7}{a_4}(a_2-b_5)$

(iii) $b_3=\dfrac{-A\pm\sqrt{(A^2-BC-BDb_1)}}{4a_4{}^2/b_7{}^2}$

Table 2. Element values for fourth-order EVC-ARC filters

Parameters	Butterworth			Chebyshev ½ dB ripple			Chebyshev 1 dB ripple		
b_7	0·89339	1·00000	0·48022	0·25799	1·00000	1·50000	1·56789	1·45678	1·18186
b_5	1·53640	1·61312	1·02426	0·28368	0·81834	0·94322	0·81633	0·80809	0·74110
b_3	2·22643	2·42654	1·36559	0·44000	1·54700	2·29711	2·24311	1·97107	1·64531
b_1	0·53454	0·41312	0·65207	0·20000	0·45000	0·23810	0·06434	0·28183	0·30000
m_1	−1·10532	−1·22118	−1·02971	−0·86398	−0·40103	−0·53058	−1·46500	−0·42232	−0·55000
m_2	2·00942	2·10493	2·60615	1·41237	2·58176	3·53152	3·31883	2·56856	2·30199
m_3	−1·36956	−1·47062	−1·92435	−1·34834	−2·73072	−3·76283	−2·78947	−2·86441	−2·45199
R_1	1·11933	1·00000	2·08238	10·22586	2·63817	1·75878	2·31399	2·49048	3·06980
R_2	1·45431	1·59882	1·05835	1·01749	2·70117	3·30441	2·14214	3·33310	2·28633
R_3	1·05599	0·97799	0·71454	1·24596	0·45648	1·71716	1·70839	1·31128	12·31055
R_4	0·58172	0·63953	0·63501	0·20350	0·81067	0·26435	0·42843	0·33331	0·04199
Resistance sum	4·21137	4·21635	4·49028	12·69282	6·60759	7·04471	6·59295	7·46817	17·70867
A_1	2·10532	2·22118	2·02971	1·86398	1·40103	1·53058	2·46500	1·42232	1·55000
A_2	−1·00942	−1·10493	−1·60615	−0·41237	−1·58176	−2·53152	−2·31883	−1·56856	−1·30199
A_3	2·36956	2·47062	2·92435	2·34834	3·73072	4·76283	3·78947	3·86441	3·45199

where

$$b_1 < a_1 b_7 / a_4 \quad \text{and} \quad D = a_4$$

(iv) $$m_1 < \frac{b_1 q_3 - (b_5 - q_4)(2q_5 - b_3)}{q_3}$$

(v) $$m_2 = \frac{(b_5 - q_5 - k_1)^2 - k_1(b_3 - q_5 - k_1)}{b_5 - q_4} + (b_1 - m_1)$$

(vi) $m_3 = b_1 - m_1 - m_2 - 1.$

The amplifier gains are then given by

$$A_1 = (1 - m_1);\; A_2 = (1 - m_2);\; A_3 = (1 - m_3). \tag{16}$$

Some of the different sets of values of the network elements obtainable from this scheme for different polynomials are given in Table 2.

4 Realization of Double Ladder Networks

Double ladder networks or multiple feedback networks constrained by a single amplifier have evoked considerable interest in recent years and a number of techniques[8–9] have been used for the synthesis of such networks. The recurrent-continuant scheme may also be utilized for the realization of double ladder EVC–ARC networks.

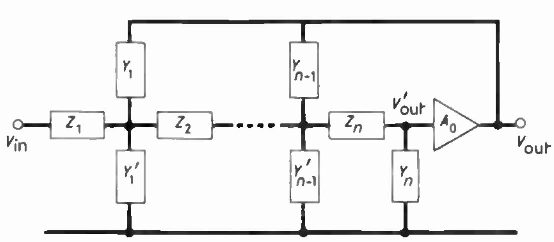

Fig. 4. A double ladder active RC network.

4.1 The Double Ladder ARC Network and the Continuant

The double ladder ARC network to be realized is as shown in Fig. 4. Using Kirchhoff's current and voltage laws the following continuant may be obtained for the voltage ratio V_{in}/V'_{out}

$$V_{in}/V'_{out} = \begin{vmatrix} Z_1 & 1 & 0 & . & 0 & 0 & 0 \\ -1 & Y_1 + Y'_1 & 1 & . & 0 & 0 & -A_0 Y_1 \\ 0 & -1 & Z_2 & . & 0 & 0 & 0 \\ . & . & . & . & . & . & . \\ 0 & 0 & 0 & . & Y_{n-1} + Y'_{n-1} & 1 & -A_0 Y_{n-1} \\ 0 & 0 & 0 & . & -1 & Z_n & 1 \\ 0 & 0 & 0 & . & 0 & -1 & Y_n \end{vmatrix} \tag{17}$$

where $V'_{out} = V_{out}/A_0$.

In the case of the ARC double ladder network one will have

$$Z_i = R_i,\quad Y_i = C_i s \quad \text{and} \quad Y'_i = G_i. \tag{18}$$

However it should be noted that it is not necessary to impose this restriction as Y'_i could be a complete admittance or even a pure susceptance.

4.2 The Transformation Procedure

Decomposing a_i (relation 2) as

$$a_i = x_i + k_i \;(k = 1, 2, 3, \ldots, n-1) \tag{19}$$

where x_i and k_i are positive constants and following Nesbitt's scheme[4] one could write the following determinant:

$$P_n(s)=\begin{vmatrix} a_n & x_{n-1}+k_{n-1} & 0 & . & x_1+k_1 & 0 & 1 \\ -1 & s & 0 & . & 0 & 0 & 0 \\ 0 & 0 & 1 & . & 0 & 0 & 0 \\ . & . & . & . & . & . & . \\ 0 & 0 & 0 & . & s & 0 & 0 \\ 0 & 0 & 0 & . & 0 & 1 & 0 \\ 0 & 0 & 0 & . & -1 & 0 & s \end{vmatrix} \qquad (20)$$

This determinant may be transformed into the continuant (17) through the following procedures:

(i) $R_{2n-1,2n}(-1)$, $C_{2n-1,2n-2}(-1)$, $R_{2n-1,1}(x_1)$, $M_{2n-2}{}^{2n-1}(k_1)$, $C_{2n-2,2n}(-1)$, $R_{2n,1}(x_1)$, $C_{1,2n}\left(-\frac{x_1}{a_n}s\right)$, $C_{2n-4,2n-2}\left(-\frac{m_1}{a_2}\right)$;

(ii) $R_{2n-3,2n-2}(-1)$, $C_{2n-3,2n-4}(-1)$, $R_{2n-3,1}(x_2)$, $M_{2n-4}{}^{2n-3}(k_2)$, $M_1{}^{2n-4}(k_2)$, $M_{2n-3}{}^{2n-2}(k)$, $C_{2n-4,2n-2}(-1)$, $R_{n+2,2n-4}(q)$, $C_{2n-4,2n-1}(q)$, . . .

where

$m_i>0$ and the q are functions of a_i, k_i and m_i.

The final steps consist of a number of alternate $R_{i,j}(q)$, $C_{i,j}(q)$ and $M_j{}^i(q)$ operations, the values of i, j and q being appropriately chosen.

Lastly, the 1st row is divided by a constant α and the entire determinant is multiplied by α, thereby getting α times the continuant (17) with $Y_i=s$, i.e., $C_i=1$.

It is to be pointed out that these operations ensure only the equality of magnitudes of the amplifier gains A_0, or the equality of magnitudes of various feedback voltages, and sometimes the latter may be in opposite phases. Such a situation will require the use of a voltage inverter at the ouput stage. Depending upon the respective phase, the various feedback voltages may be taken either from the output of the amplifier or that of the inverter as shown in Fig. 5.

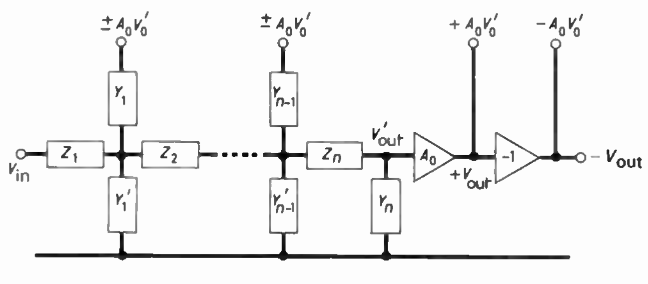

Fig. 5. ARC double ladder with an inverter.

4.3 Illustration

The proposed scheme has been illustrated by realizing a 3rd-order EVC–ARC double ladder network.

Let

$$P_3(s)=\sum_{i=1}^{3} a_i s^i+1$$

be the prescribed characteristic polynomial. Following Nesbitt's scheme, $P_3(s)$ may be written as the following determinant (relation 20):

$$P_3(s)=\begin{vmatrix} a_3 & x_2+k_2 & 0 & x_1+k_1 & 0 & 1 \\ -1 & s & 0 & 0 & 0 & 0 \\ 0 & 0 & 1 & 0 & 0 & 0 \\ 0 & -1 & 0 & s & 0 & 0 \\ 0 & 0 & 0 & 0 & 1 & 0 \\ 0 & 0 & 0 & -1 & 0 & s \end{vmatrix} \qquad (21)$$

It is now subjected to the following series of operations:

(i) $R_{5,6}(-1)$, $C_{5,4}(-1)$, $R_{5,1}(x_1)$, $M_4{}^5(k_1)$, $C_{4,6}(-1)$, $R_{6,1}(x_1)$, $C_{1,6}\left(-\frac{x_1}{a_3}s\right)$, $C_{2,4}\left(-\frac{m_1}{a_2}\right)$;

(ii) $R_{3,4}(-1)$, $C_{3,2}(-1)$, $R_{3,1}(x_2)$, $M_2{}^3(k_2)$, $M_1{}^2(k_2)$, $M_3{}^4(k_1)$, $C_{2,4}(-1)$, $R_{5,2}(-q_8 s)$,

where

$$q_8=\frac{m_1k_2}{a_2}+1$$

(iii) $C_{2,5}(q_2k_1s)$, $R_{6,1}(q_2k_1)$, $C_{1,6}\left(-\frac{k_1k_2q_8}{a_3}s\right)$, $R_{6,3}(-q_8k_1)$, $C_{3,6}\left(\frac{k_1{}^2q_8}{k_2}s\right)$, $C_{1,6}\left(-\frac{x_2k_1q_8}{a_3}s\right)$, $R_{3,1}(m_1)$, $C_{1,3}\left\{-\frac{k_2(x_2+m_1k_2)}{k_1a_3}\right\}$, $C_{1,2}(-m_2)$, $R_{6,4}(-1)$;

The determinant (21) then takes the form

$$P_3(s)=\begin{vmatrix} \frac{a_3}{k_2} & \alpha & 0 & 0 & 0 & 0 \\ -1 & (s+m_2) & d_{2,3} & 0 & 0 & -A_1s \\ 0 & -1 & k_2/k_1 & 1 & 0 & 0 \\ 0 & 0 & -1 & \left(s+\frac{m_1k_1}{a_2}\right) & 1 & -A_2s \\ 0 & 0 & 0 & -1 & k_1 & 1 \\ 0 & 0 & 0 & 0 & -1 & s \end{vmatrix} \qquad (22)$$

where

$$\alpha=\left(1-m_1-\frac{m_2 a_3}{k_2}\right)$$

and the term

$$d_{2,3}=\frac{k_2}{a_3 k_1}(a_2-k_2+m_1 k_2).$$

Now putting $d_{2,3}=1$, $A_1=A_2=A_0$ and dividing the first row by α and multiplying the entire determinant by α, one gets α times the active continuant:

$$P_3(s)=\begin{vmatrix} R_1 & 1 & 0 & 0 & 0 & 0 \\ -1 & s+G_1 & 1 & 0 & 0 & -A_0 s \\ 0 & -1 & R_2 & 1 & 0 & 0 \\ 0 & 0 & -1 & s+G_2 & 1 & -A_0 s \\ 0 & 0 & 0 & -1 & R_3 & 1 \\ 0 & 0 & 0 & 0 & -1 & s \end{vmatrix} \tag{23}$$

This active continuant represents the reciprocal of the voltage transfer function of a third-order double ladder ARC network shown in Fig. 6.

It is obvious that the transformation of the determinant (21) into the active continuant form (23) is valid under the restrictions:

(i) $$m_1=\frac{a_3 k_1-k_2(a_2-k_2)}{k_2^2} \tag{24}$$

obtained by putting $d_{2,3}=1$.

(ii) $$a_3^2 k_1^2-a_3 k_2(a_2-k_2)k_1^2+(a_3^2 k_2)k_1+ +k_2^2(a_3 k_2+2a_2 a_3-a_1 a_2 k_2)=0 \tag{25}$$

obtained by equating the magnitudes of the amplifier gains A_1 and A_2 and using relation (24).

(iii) $$m_2=\frac{k_2}{a_3}(a_2+m_1 k_2-k_2). \tag{26}$$

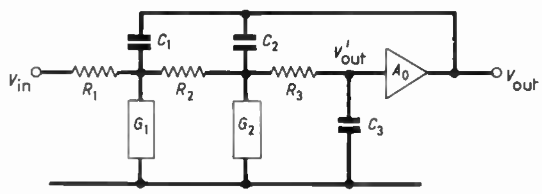

Fig. 6. Third-order double ladder ARC network.

Under these restrictions the other network elements are given in terms of the polynomial coefficients a_i and the arbitrary positive constants k_i and m_i as follows:

(i) $R_1=\dfrac{a_3}{k_2}$

(ii) $R_2=\dfrac{k_2}{k_1}$

(iii) $R_3=k_1$

(iv) $G_1=m_2$

(v) $G_2=\dfrac{m_1 k_1}{a_2}$, and

(vi) $A_0=\pm\dfrac{k_1^2 q_8}{k_2}+2.$

Some of the sets of values of the network elements obtained for different polynomials are given in Table 3.

5 Conclusion

A new recurrent-continuant transformation procedure has been proposed for the realization of active RC ladder networks. Adopting the frequency transformation scheme $s=y^2$ and modifying the prescribed all-pole transfer function by introducing certain odd-powered polynomial coefficient parameters, the requisite active continuant has been obtained by simple determinant operations. There

Table 3. The element values for third-order EVC-ARC double ladder networks

Parameters	Butterworth			Chebyshev ½ dB ripple			Chebyshev 1 dB ripple		
k_1	1·67000	2·19000	2·58310	1·26100	1·60650	1·88650	1·36400	1·64482	1·88400
k_2	2·00000	2·50000	3·00000	3·00000	3·50000	4·00000	4·00000	4·50000	5·00000
m_1	0·41750	0·54340	0·62000	0·61223	0·68306	0·72708	0·62430	0·71829	0·75100
$m_2(=G_1)$	0·56500	0·64150	0·84000	0·40316	0·29290	0·26601	0·24700	0·29120	0·22121
	0·30000	0·20000	0·10000	0·20000	0·20000	0·18000	0·25000	0·15000	0·16000
G_2	0·34861	0·59502	0·80010	0·44100	0·62683	0·78352	0·42330	0·58730	0·70334
$1/G_1$	1·76991	1·55880	1·20000	2·48041	3·41370	3·75927	4·04810	3·43412	4·55060
$1/G_2$	2·86851	1·68060	1·24000	2·26755	1·59530	1·27628	2·36236	1·70270	1·42180
R_1	1·41723	2·00000	3·33333	2·32874	1·99600	1·94062	2·03538	3·01539	2·54423
R_2	1·19760	1·14155	1·16140	2·37906	2·17865	2·12033	2·93255	2·73586	2·65392
R_3	1·67000	2·19000	2·58310	1·26100	1·60650	1·88650	1·36400	1·64482	1·88400
Resistance sum	8·92325	8·57095	9·51783	10·71676	10·79015	10·98300	12·84239	12·53289	13·05475
A_0	4·00000	5·00000	6·30000	3·08000	3·74440	4·36700	3·04250	3·56700	4·00000

is considerable flexibility in the choice of these parameters although they have to satisfy certain restrictions, to ensure positive valued resistor and unit valued capacitors. The method is thus capable of yielding multiple sets of element values and has been illustrated by realizing third- and fourth-order networks with different transfer characteristics in the EVC structures. The results have been presented in Tables 1 and 2.

The proposed method is more general than the Nesbitt technique[4] in that it does not require any pre-fixation of the values of the passive elements and is simpler than those depending upon the coefficient matching methods,[5–7] since, unlike the latter, it does not require the solution of nonlinear equations.

The method has also been extended to the case of double ladder networks and the relevant technique has been described with appropriate examples.

6 Acknowledgment

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Indexing Terms: Radio communication, Radio receivers, Data communication, Microprocessors

Synchronization of a spread-spectrum receiver by a microprocessor control system

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SUMMARY

The use of a microprocessor to control the synchronization of a low-data-rate spread-spectrum system is described. The microprocessor controls a voltage-controlled oscillator and sequence-generator, and carries out the correlation processes needed for the acquisition and tracking modes. Switching between modes is achieved by changing an interrupt vector.

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1 Introduction

Communications systems in which the transmitter processes the information-signal to spread it over a wide bandwidth are collectively known as spread-spectrum systems, and a variety of schemes are available.[1–3] These spread-spectrum techniques can provide signal-hiding and immunity from jamming, may enable a channel to be shared with other, conventional, transmissions, and under some conditions can provide better utilization of a multiplexed channel. Typically, the techniques use a pseudo-random subcarrier which is modulated by narrow-band information (data), producing a wideband noise-like signal which may undergo subsequent modulation processes, or, in some cases, may be transmitted directly. Because the transmitted signal is spread over a wide bandwidth, operation at signal-to-noise ratios of less than unity is possible (and the signal can therefore be hidden from unwanted interception), while the receiver uses correlation to 'de-spread' the wide-band signal and recover the information. A major design problem in such systems is the acquisition and maintenance of synchronization between the pseudo-random signals at transmitter and receiver.

One of the more straightforward methods of spread-spectrum communications uses a binary pseudo-random subcarrier, with two phase ($\pm 180°$) p.s.k. modulation by binary data. This modulation process is equivalent to modulo-2 addition of the data and subcarrier, and has also been called sequence inversion modulation.[4] Various tracking-loops for pseudo-noise signals have been described,[5,6,7] which can be modified to operate in the presence of modulation.[4,8] Prior to tracking the signal, acquisition (frame-synchronization) must take place, and must be repeated if tracking subsequently fails.

Those parts of the receiver concerned with acquisition and with tracking can be mostly digital, excluding those systems which use direct transmission of high-frequency pseudo-noise signals, for which analogue correlation techniques (e.g. using surface acoustic wave filters) are necessary. The extent to which microprocessors may be used in the implementation is therefore of interest, because of the flexibility that should follow from defining the acquisition and tracking procedures by software. Presently available general-purpose microprocessors (other than bit-slice bipolar types) are too slow for this approach to be seriously competitive with conventional digital hardware, though at least an order-of-magnitude speed increase may be expected from future technology improvements.

The following Sections describe microprocessor-control of the acquisition and tracking stages of synchronization of a low-data-rate system. The F8 microprocessor was used, which has advantages of low cost, easy interfacing, reasonable speed, and the availability of a software-compatible single-chip microcomputer. The F8 is comparable in speed to the 8080 and 6800 but is simpler to interface and has more c.p.u. registers.

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Fig. 1

(a) Pseudo-random sequence (period = 127 bits).

(b) Autocorrelation function.

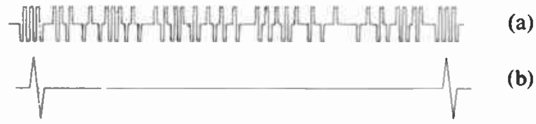


Fig. 2

(a) Three-level sequence.

(b) Cross-correlation function between 1(a) and 2(a).

Typical instructions (such as adding the contents of a c.p.u. register to the accumulator contents) take 2 μs. The objective was to carry out as much of the task as possible in software, subject to keeping within a maximum bit-time of 100 μs for the pseudo-random carrier.

In developing the software, the overriding need was to minimize execution-time. Every available feature of the instruction set which could reduce execution time had to be used. Reducing program length was of no importance at all, as the program memory requirement was far less than 1 Kbyte.

2 Correlation Properties of the Pseudo-random Carrier

The carrier is a pseudo-random sequence chosen for its autocorrelation properties. Such a sequence, together with its autocorrelation function, is shown in Figs. 1(a) and 1(b). Cross-correlation of a locally-generated sequence with the received sequence would enable the relative time-displacement of the received and local sequences to be found, and hence, in principle, would permit initial acquisition. In practice, computing the whole cross-correlation function would take impractically long, so an approximation has to be used, as described below.

After acquisition, the received and local sequences are aligned, and must be maintained in alignment to within one bit-time. This can be achieved by cross-correlation of the received sequence with a suitable three-level sequence (derived from the local sequence), as shown together with the cross-correlation function, in Figs. 2(a) and 2(b). This cross-correlation function forms the error signal for a voltage-controlled oscillator which clocks the local sequence generator.

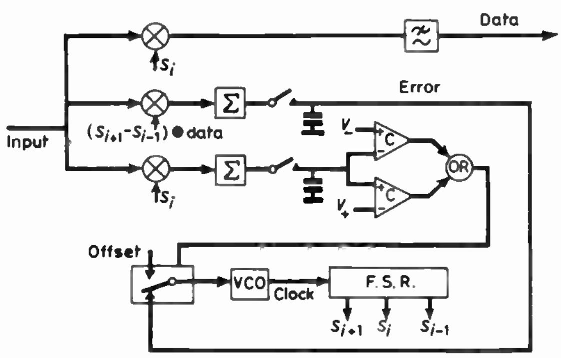


Fig. 3. Implementation without a microprocessor.
V_+, V_- = thresholds to detect peak; F.S.R. = feedback-shift-register; C = comparator.

3 Principles of Operation

Figure 3 shows a block diagram of the main parts of a system for both the acquisition and tracking processes. Although in principle, the whole system could be implemented within a microprocessor, the clock rate would then have to be so low as to be useless. The partition chosen was to implement the voltage-controlled oscillator and sequence-generator in hardware, and the remainder in software, as shown in Fig. 4. The maximum clock rate is still very limited because of speed limitations of existing m.o.s. microprocessors.

The input is assumed to have already been obtained from a higher-frequency carrier by other means, so as to leave the pseudo-random sub-carrier with its binary p.s.k. modulation.

The data may either be asynchronous with respect to the pseudo-random sequence clock, or synchronous, in which case an integral number of data bits (possibly only one), is transmitted for each sequence period. The implementation to be described is applicable to any of these alternatives.

The receiver operates in two alternative modes: searching (acquisition) and tracking.

The clock (voltage-controlled oscillator output) drives the local sequence generator and simultaneously provides interrupts to the processor. The mode is determined by

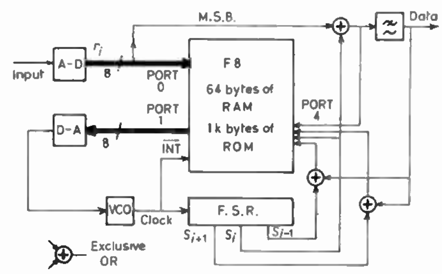


Fig. 4. Microprocessor implementation.

the interrupt vector, which is changed by software to switch between modes.

While searching, the processor sends a constant voltage to the voltage-controlled oscillator which causes a frequency offset between the local clock rate and the nominal clock rate of the transmitter. This causes the local and received sequences to slide past each other, and the cross-correlation estimator generates a value which exceeds a threshold only when they are aligned. (A compromise is needed between a small frequency offset which results in a long average-acquisition-time, and a large frequency offset, for which the cross-correlation peak is small and liable to be missed unless a low threshold is set, which in turn leads to false acquisition in the presence of noise.)

When the threshold is exceeded, it is assumed that the sequences are aligned, and the processor immediately alters the voltage sent to the voltage-controlled oscillator to cause it to run at the nominal transmitter clock-rate, and changes the interrupt-vector to that for tracking. The error signal produced by the tracking process is then added to the voltage sent to the voltage-controlled oscillator, so maintaining the alignment.

The execution time of the software for each mode has to be short enough for all necessary instructions to be completed between interrupts. The processor must be in a 'wait-for-interrupt' condition before each clock pulse occurs. (In the case of the F8, this means repeated execution of a self-branch instruction as there is no wait instruction as such.)

A simplified description of the program intended to explain its structure is given in Appendix 1. The listing of sequences of F8 assembly language mnemonics has been deliberately avoided, and instead, a higher-level pseudo-language has been used which it is hoped will be self-explanatory.

4 Search Mode Details

As mentioned previously, it would be impractically slow to calculate the cross-correlation function accurately, since this would involve summation over a complete period of the sequence, and repeating this for successive time-shifts until the peak were found. Fortunately, summation over a sub-sequence is adequate. This can be seen from Fig. 5 which shows the building-up of the correlation estimate

$$\sum_{i=0}^{m} r_i s_i$$

as a function of m. Clearly, the aligned case can be distinguished from the others when m is far less than the sequence period.

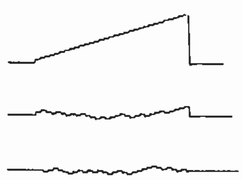


Fig. 5. Build-up of correlation estimate for sequence period 127 and maximum $m=32$, (a) received and local sequences aligned (b), (c) typical examples for sequences not aligned.

Because the pseudorandom sequence is modulated with unknown binary data, the peak in the correlation function may be positive or negative (Fig. 6(a)). Therefore it is necessary to test whether it is above a positive threshold or below a negative threshold.

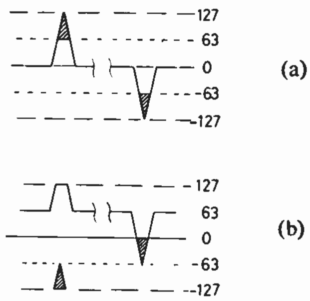


Fig. 6

(a) Alternative values for correlation peak.
(b) Values after adding threshold.

The following description of the method used to detect this peak provides an indication of the execution-time savings which are achieved by careful program design.

The computations are scaled so that the peak has a nominal magnitude of 127 (the maximum positive 8-bit two's complement integer) and the threshold is set at half this value (63). An obvious algorithm to detect the peak is as follows:

if ($\langle$sum$\rangle > 63$) goto *change*
if ($\langle$sum$\rangle < -63$) goto *change*
wait for interrupt

change:

change to tracking-mode
wait for interrupt.

Coding this directly into F8 instructions requires two compare instructions and two conditional branches and execution time is between 26 and 54 μs (depending on whether or not the peak is detected).

If, instead, the threshold is first added to the contents of sum, the overflow properties of two's complement arithmetic change the values as shown in Fig. 6(b). Now if the peak is above the threshold, the result is negative regardless of its original sign. The algorithm becomes

sum $\leftarrow \langle$sum$\rangle + 63$
if ($\langle$sum$\rangle < 0$) change to tracking mode
wait for interrupt.

In F8 machine language, this requires only one conditional branch, and execution time is between 9 and 38 μs (the shorter time applying when the peak is not detected).

The same attention to detail was required in all sections of the code.

5 Track-mode Details

The three-level sequence required to derive the error signal is obtained by subtraction of two versions of the locally generated sequence, one delayed by one-bit time and the other advanced by one-bit time relative to the version aligned with the received sequence. The correlation sum is therefore given by

$$\sum_{i=0}^{m} r_i(s_{i+1} - s_{i-1})$$

the same value for m being used as in search mode.

Because of the modulation, the N-shaped error curve (Fig. 2(b)) may or may not be inverted, depending upon whether the data is 1 or 0 (Fig. 7). However, by forming the exclusive-OR of the sequences s_{i+1} and s_{i-1} with the data prior to calculating the correlation sum, this dependence of the error curve on the data is avoided. (Otherwise, the feedback in the error-control loop would alternate between positive and negative, as the data alternated between 0 and 1.)

In an implementation such as Fig. 3, the searching and tracking correlators are separate and can operate simultaneously. Therefore, during tracking, the search correlator output is still available and can be monitored to ensure that the correlation peak does not fall below the threshold, which would indicate loss of synchronization.

Because of the serial-processing nature of the microprocessor implementation, it is not possible to operate the search correlator while in the tracking mode without incurring a serious penalty in increased execution-time. An alternative method of detecting loss of synchronization is needed, and can be based on the rapid level-transitions which occur on the data-output (prior to filtering) in the unsynchronized condition only. These can be detected during the tracking-mode by loading a register with a constant at the instant of one transition, decrementing the register for each carrier clock-pulse until the next transition, and hence determining the time between the transitions. If the time is shorter than a prescribed minimum, the interrupt address is changed back to that for the search mode. A disadvantage of this method is that under high noise-level conditions, spurious transitions of the data can occasionally cause a false indication of loss of synchronization with a consequent change to search mode. Instead the rate of level-transitions can be measured by a simple diode-pump circuit having its output-level monitored by the microprocessor (only one bit of one I/0 port is used for this).

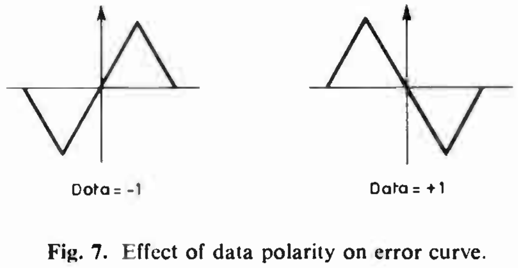


Fig. 7. Effect of data polarity on error curve.

6 Some Alternative Approaches

Using the interrupt-line to synchronize the processor to the voltage-controlled oscillator involves the time overheads associated with interrupt-handling. In those microprocessors which automatically save the status of a number of c.p.u. registers following an interrupt, this time is very long, and an alternative (e.g. polling an input port) would be needed. In the case of the F8 processor, only the program-counter contents are saved following an interrupt, resulting in a comparatively fast interrupt response. It should, however, be noted that the method of switching between acquisition and tracking modes involves changing the interrupt-vector by software. In some implementations, particularly the single-chip-microcomputer version of the F8, the interrupt vector is fixed by internal hardware and a different approach would have to be used.

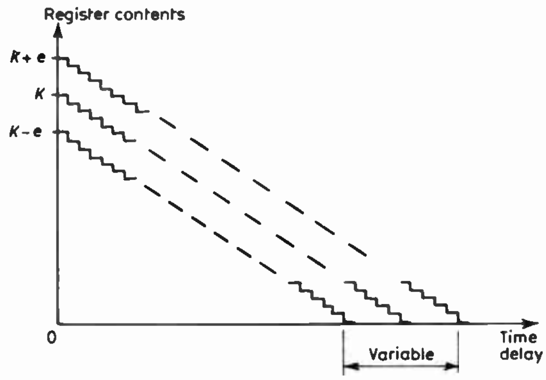
The decision to implement the voltage-controlled oscillator function outside the microprocessor led to the need for the digital-to-analogue converter. The variable clock-rate required during tracking could alternatively have been directly generated by the processor, as follows.

By loading a register with the sum of an initial constant, K, and the calculated error, and then using this register to count down to zero, a time delay dependent upon the value of the error can be generated. This is illustrated in Fig. 8(a), and the required modification to the tracking-mode software is illustrated by Fig. 8(b). The constant, K, is chosen so that the nominal clock rate is generated when the error is zero. A corresponding change to the search mode can be made simply including a preset delay in the computational-loop to produce the required clock-frequency offset for tracking. As a result, interrupts are not needed either for searching or for tracking.

However, because the system clock-rate is no longer determined externally to the processor by the centre-frequency of the voltage-controlled oscillator, any changes to this clock-rate would require software modifications (to change the constant, K, and the search-mode frequency offset)—although provision for several pre-determined alternative rates could be incorporated in the software and selected by reading in an appropriate control word from one of the ports during system-start-up.

7 Conclusions

The use of a microprocessor to control the acquisition and tracking modes of a spread-spectrum receiver offers advantages in being able to define the procedures by software.



(a) Generation of variable clock-period by software (e = error signal from tracking correlator).

```
track:  — — — —
        — — — —
        — — — —
        — — — —
        — — — —
        clock ← 1
        — — — —
        — — — —
        — — — —
        — — — —
        clock ← 0
        delay count ← K + error
        while (<delay count> ≠ 0) decrement delaycount
        goto track
```

(b) Incorporation of variable delay into tracking-loop software.

Fig. 8

The task is simple enough for implementation using the single-chip F8 microcomputer, together with a few external components.

However, presently available m.o.s. microprocessors are too slow for this to be used for other than low data-rate applications.

8 References

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7 Gill, W. J., 'A comparison of binary delay-lock tracking-loop implementations', *IEEE Trans. on Aerospace and Electronic Systems*, **AES-2**, pp. 415–24, July 1966.

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9 Appendix: Simplified Description of Program

set-up: v.c.o.-control ← *offset*
interrupt-address ← *search*

search: sum ← 0
count ← m
perform sum ← ⟨sum⟩ + $r_i s_i$
count ← ⟨count⟩ − 1
until ⟨count⟩ = 0
if (|⟨sum⟩| > *threshold*)
interrupt-address ← *track*
v.c.o.-control ← *centre*
wait for interrupt

track: sum ← 0
count ← m
perform sum ← ⟨sum⟩ + $r_i s_{i+1} - r_i s_{i-1}$
count ← ⟨count⟩ − 1
until ⟨count⟩ = 0
error ← ⟨sum⟩
v.c.o.-control ← *centre* + ⟨error⟩
if (tracking-failure)
interrupt-address ← *search*
v.c.o.-control ← *offset*
wait for interrupt

sum, count, error
are mnemonic names for scratch-pad registers

offset, *centre*, *m*, *threshold*
are constants

setup, *track*, *search*
are labels for addresses

a ← ⟨b⟩
denotes contents of register (b) transferred to register (a)

v.c.o.-control
is a name for the input register of the digital-to-analogue converter which controls the voltage-controlled oscillator

s_{i-1}, s_i, s_{i+1}
three successive time-shifts of the local sequence

r_i
received signal value

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Indexing Terms: Radar, Surveillance applications, Aircraft

Reflection elimination in secondary surveillance radar

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Based on a presentation given at the Eurocontrol Institute, Luxembourg during the October 1978 Seminar on 'Evolution of Radar Systems'

SUMMARY
Secondary Surveillance Radar in Air Traffic Control has evolved from a mere labelling device in support of primary radar to a high integrity plotting system in its own right with primary radar relegated to the support role. This change in concept has placed increased demands on the system and there is a greater awareness of the shortcomings and idiosyncrasies which previously were unnoticed or accepted, among which is the problem of reflections causing false target reports. This paper outlines the nature of the difficulty and some of the principal methods for reducing it to an acceptable level.

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1 Introduction

The internationally accepted standards for Secondary Surveillance Radar (SSR) are detailed in a document published by the International Civil Aviation Organization (ICAO) and commonly referred to as Annex 10.[1] Minimum performance standards and test specifications are also detailed in further documents[2,3] and transponder characteristics are specified in ARINC 572D.[4]

Basically, SSR operates by the transmission via a directional beam from the ground interrogator of two pulses designated P1 and P3 of 0·8 μs each, spaced according to the mode of operation. If identity is requested, Mode A is used in which the separation is 8 μs. Mode C is used to request aircraft height for which the spacing is 21 μs. All interrogations are transmitted on 1·03 GHz. The two modes are normally transmitted alternately so that continuous monitoring of both is maintained. Sidelobe suppression (s.l.s.) is provided by the transmission of a third pulse, P2, 2 μs after the first P1 pulse of the mode pair. This pulse radiates from an essentially omnidirectional aerial. Power levels, tolerances and thresholds are set so that outside the main interrogation beam the transponder recognizes that the P2 pulse is larger than the preceding P1 sidelobe pulse and does not reply. Instead, the transponder suppresses into an inactive state for up to 45 μs during which time re-interrogation by any station is not possible. On a frequency of 1·09 GHz, the aircraft transponder replies to an interrogation with two framing or bracket pulses F1 and F2, 0·45 μs long and spaced 20·3 μs. Up to twelve other code pulses may also be transmitted at specified positions between the bracket pair and these will be interpreted by the ground station as the identity or the height of the aircraft. One pulse position, at 10·15 μs after F1, termed X, is disallowed in SSR, and if present for any reason, would normally invalidate the code. In its original role of providing identity to otherwise anonymous signals derived from primary radars, the system worked well, but as more reliance was placed on its data to provide positional information, situations allowing reflections to be interpreted with as much ease as the direct signals has provoked some re-thinking on the implementation of the system.

2 The Reflection Problem

Reflections in SSR manifest themselves in several ways according to whether the transponder and its reflected image lie simultaneously within the radar main beam or not. If it does, vertical lobing, degraded azimuth estimation and code corruption are the principal results depending on the orientation and range of the reflecting surface. When the transponder and a reflector are separated by more than a beam-width, and the reflector is largely composed of vertical surfaces, it will give rise to an image of a transponder lying on an entirely different

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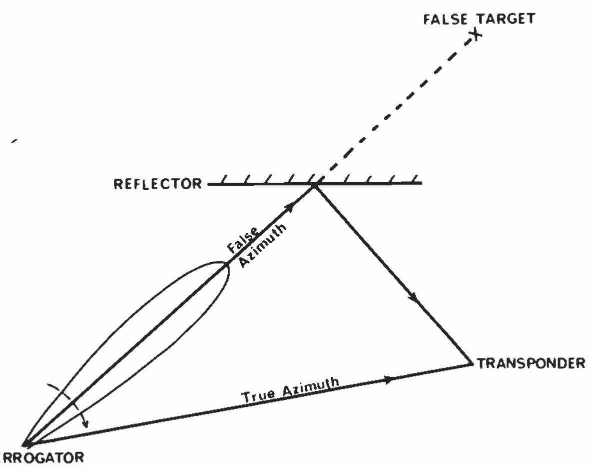


Fig. 1. Generation of false targets.

azimuth (Fig. 1). This image may respond to an interrogation giving rise to a false target and if the reflector is large, these images develop extended tracks over many rotational periods of the aerial. It is this situation which is the subject of this paper.

2.1 Echo Suppression Techniques

There are in fact some facilities built in to the specifications to help overcome significantly delayed reflections. One of these is termed 'echo suppression' and operates within the transponder by effectively desensitizing the receiver to near the level of each signal as it arrives. Recovery then follows at an average rate of 3·5 dB/μs. An optional facility, not used on British civil radars, is an 'improved' s.l.s. (i.s.l.s.) system in which a weak P1 pulse is deliberately radiated via the omni-aerial to ensure that suppression always occurs in the sidelobe region. The reflected and delayed mode pair P1 and P3 from the main beam will then be ineffectual during the suppression time. Echo suppression only operates for delayed and attenuated signals arriving after the main interrogation mode pulses have passed and being limited to the main beam region does not improve the behaviour in the circumstances just described. As we shall see, i.s.l.s. may not always be an improvement.

2.2 Characteristics of Echo Suppression Systems

The effects of reflections may usefully be considered by a similar approach to that used for the effects of sidelobes—this being the case, the interrogator sidelobe suppression system should prevent reflective responses, provided certain conditions can be met. These are:

(1) That the reflected P1 signal amplitude at the transponder does not exceed the control power, and

(2) that the time of arrival of a reflected P1 pulse at the transponder is not delayed by more than 150 ns.

If a sidelobe P1 or an enhanced P1 from an i.s.l.s. system has initiated transponder action in expectation of a P2 pulse, the arrival of a delayed P1 within 1·85 μs may prevent suppression if its amplitude is equal or greater than that pulse. At many sites reflection delays are of less than 2 μs and consequently, i.s.l.s. would not be as effective as may be desired. For a distant transponder, the geometry is simple (Fig. 2), and one can derive the expression for the extra path length, d, for a reflector r metres from the radar and $\theta°$ from the transponder direction. This is represented by the cardioid

$$d = r(1 - \cos\theta).$$

A reflector on a reciprocal bearing to the transponder causes a range increase equal to twice its distance from the radar, and if at right angles the increase is equal to its distance from the radar. In terms of time, each 300 m increase of d is equivalent to 1 μs. For small values of θ, a large distant reflector may only impose a very short delay.

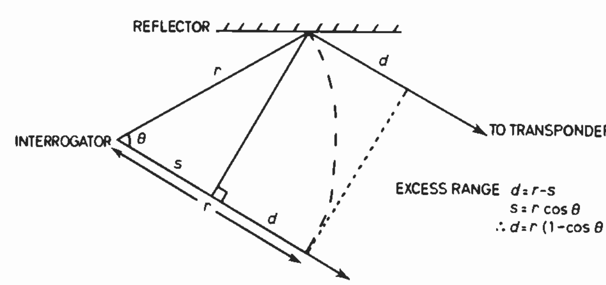


Fig. 2. Geometry of signal paths (simplified).

With i.s.l.s., the simultaneous transmission of a main beam and omnidirectional P1 from different aerials also causes interference problems, causing uncertain suppression cover and main beam blackouts. Most British civil SSR installations use an integral control aerial with simple s.l.s. The control pattern generates little P2 power in directions at right angles to the main beam where many reflectors may lie. It may therefore be surmised that such installations are prone to the generation of false targets.

It was with this background that RSRE joined with the CAA in the effort to examine methods for alleviating the problem. This paper covers a number of concepts, many of which would be difficult to employ at existing operational sites but are included to indicate areas where duplication of thought may lead to unfruitful conclusions.

3 Solutions

There are several approaches, but in general all may come under one of two headings:

(1) Up-link interrogation methods.

(2) Down-link reply methods.

3.1 Up-Link Methods

The areas where anti-reflection measures may be applied and which are relevant to the interrogation path are in:

siting;
reducing the cross-section of the reflector;
improving aerials;
modifying interrogation signal formats;
modifying transponder circuits.

3.1.1 *Siting*

Because of the particular problems arising from reflections, the requirements for an SSR site are somewhat more stringent than those for a simple primary radar. This is because of the inverse square law applicable to SSR against the inverse fourth power law for primary radar. Where joint primary-secondary equipment is to operate, the choice ought therefore to be dictated by the secondary radar. The wide vertical beamwidth of current SSR aerials allows ground reflections to cause vertical lobing and high tower mounting is sometimes used to generate a finer lobe structure with reduced null depth. This also allows the radar to see a larger area of ground with a proportionally larger number of buildings potentially capable of causing false targets. The alternative of having a mounting very close to the ground to avoid lobing is precluded by the inability to interrogate low angle targets obscured by buildings. A counterpoise supported by a high tower may be impractically large and a better solution would be to mount the radar on a high building having a large roof area, thereby gaining the advantages of good low angle cover, reduced lobing and a degree of screening from the local environment afforded by the building itself (Fig. 3).

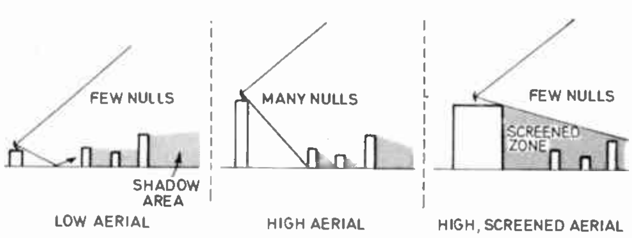


Fig. 3. Effects of aerial mounting.

It must be admitted, however, that mountain top sites have sometimes shown serious within-beam multipath problems and it may be that such sites would be closer to the ideal if the tops were levelled or even hollowed over a reasonable area and the radar mounted level with the site perimeter.

3.1.2 *Reflector elimination*

To gain a simple idea of the effectiveness of a reflecting surface one may compare the signal strengths received by a transponder both directly and via a reflector (Fig. 4). For a direct beam the beacon equation applies and the receiver signal power is

$$P_r = \frac{P_t G_t G_r \lambda^2}{(4\pi)^2 R^2}$$

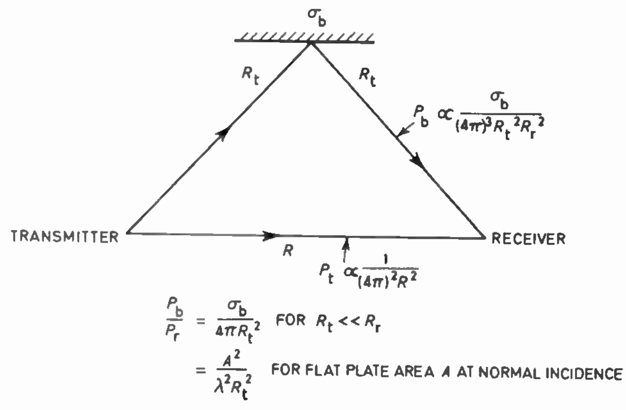


Fig. 4. Reflection loss determination.

For the reflective case, the situation is identical to a bistatic radar for which one assigns a bistatic cross-section to the reflector, σ_b, and the reflected power is

$$P_b = \frac{P_t G_t G_r \lambda^2 \sigma_b}{(4\pi)^3 R_t^2 R_r^2}$$

where

P_t, G_t = transmitter power and aerial gain
G_r = receiver aerial gain
λ = wavelength (0·275 m and 0·291 m)
R = interrogator-transponder range
R_t = interrogator-reflector range
R_r = transponder-reflector range
σ_b = effective bistatic cross-section of the reflector.

The ratio of reflected to directly received signal is then

$$\frac{P_b}{P_r} = \frac{\sigma_b}{4\pi} \cdot \frac{R^2}{R_t^2 R_r^2} \approx \frac{\sigma_b}{4\pi R_t^2}$$

when $R = R_r$.

This would apply to both up and down links. The ratio P_b/P_r is seen to be the ratio of the bistatic cross-section of the reflector to the surface area of a sphere of radius R_t. For reflectors which are simple structures such as flat surfaces or masts, the assignation of σ_b could be relatively simple provided that they are small compared with the spatial beamwidth of the aerial.

When the reflector is close to the interrogator, $R_t \ll R_r$ and $R \approx R_r$, then

$$\frac{P_b}{P_r} = \frac{\sigma_b}{4\pi R_t^2}$$

$$= \frac{A^2}{\lambda^2 R_t^2}$$

for a flat plate area A at normal incidence.

In many situations, the reflectors are wider than the beam and become more like perfect mirrors for which little loss may be expected, indeed, a flat reflector equal in size to the first Fresnel zone may actually increase the signal power.

The reflector material also determines the magnitude of the reflected signal and although many structures

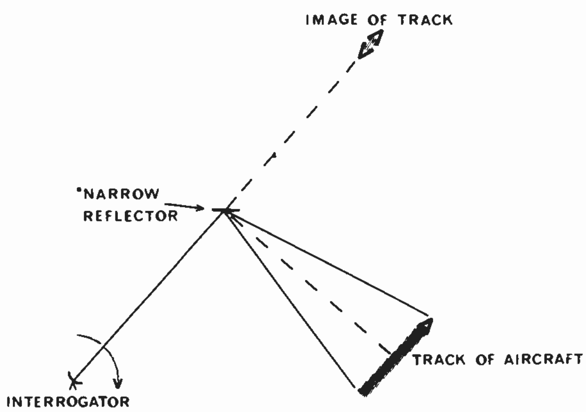


Fig. 5. Compressed image in narrow reflector.

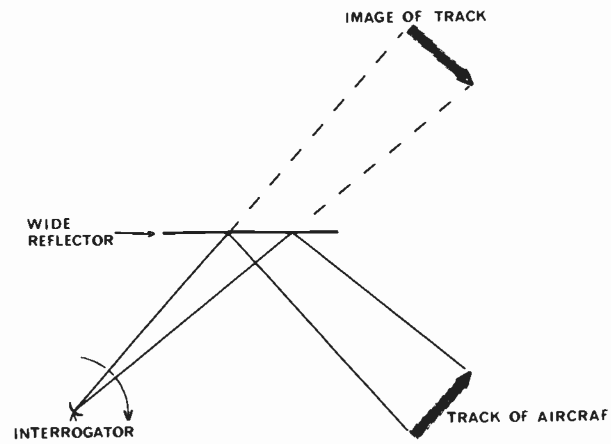


Fig. 6. Track image in wide reflector.

such as aircraft and aircraft hangars, are built of metal, other materials will have higher ohmic losses. As an example we may consider a tail fin of a large aircraft as a large flat plate of area approximately 80 m^2. For normal incidence $\sigma_b = 4\pi A^2/\lambda^2 = 950{,}000$ m^2, nearly one million square metres which, if positioned 1 km from the radar, gives a loss of $10^6/4\pi \times 10^6 \simeq 0{\cdot}078$ or 11 dB, assuming the fin to be completely covered by the beam.

This is not a very large attenuation and this sample calculation emphasizes the problem at airport-sited radars. A low reflection loss implies a limited angle of coverage by the reflector and this is determined by the linear dimensions, a, of the reflector as if it were an aerial aperture in its own right; approximately $51\lambda/a$ degrees. The tail fin averaging 9×9 m say, would cover 1·6° horizontally and vertically, a rather limited angle, but is effectively much wider because of sidelobes and also because a tail fin is often more nearly triangular than square in shape.

A tall thin structure would give a limited vertical coverage but being narrow would scatter widely in azimuth, and aircraft in different directions may be interrogated whilst the interrogator aerial is directed towards the reflector and many false plots may appear on the one bearing. An aircraft flying tangentially to the reflector generates a succession of reflected plots in this structure, appearing to be almost stationary for the range changes are slight (Fig. 5). A long building (Fig. 6) will also cause reflections over a wide arc but generating perfect mirror images of tracks over long periods of time. Although both types of reflector are capable of covering a wide field of aimuthal arc, the difference is that whereas the thin structure gives simultaneous cover, the long structure develops a reflected beam which sweeps as the aerial rotates.

It is self-evident that if a reflector can be identified, then some measures may be possible to reduce its effectiveness, as has been done, for example, at Tulsa where the control tower was fitted with radiation absorbent panels. Such treatment is not usually possible, either on account of cost or impracticability as would be the case where the reflector is an area of metallized solar control window. During investigations, a portion of the 747 Hangar at Heathrow was identified as a serious reflection problem but this was reduced by covering with galvanized wire mesh orientated to deflect the incident radiation upwards.

Security fences and some other structures which may be under the control of or accessible to the radar managers might be adjusted by re-alignment, tilting or local screening but in practice little can be done with most reflecting structures.

3.1.3 *Aerial design*

Aspects of main beam aerial design are well appreciated, and improvements can be made, principally by adopting sharp bottom-edge cut-off in order to reduce the power directed at low angles where reflections usually lie. Some caution is however necessary because although a cut-off rate is usually quoted as the figure of merit, what actually matters is the low angle gain and unless this could be brought down to very low values, it will not be very effective, particularly in terminal areas where transponder ranges are short and received signal levels correspondingly high on both up and down paths. A mere 10 or 20 dB of reflection loss would need to be supplemented by a considerable reduction of aerial gain to bring the power down below receiver threshold. On the down link, swept gain would assist in providing extra desensitization. A further difficulty is that long range aircraft appear at lower elevation angles than some tall buildings and there is the likelihood of losing detectability of these, whilst gaining no protection against high angle reflections.

It was mentioned that control patterns are not always truly omnidirectional and the lack of control cover in significant directions allows reflective interrogations which might otherwise be suppressed. A true omnidirectional control pattern would overcome this problem, although other considerations may make practical difficulties of implementation.

3.1.4 *Modified interrogation formats*

To avoid reflective interrogations we must bear in mind that we are dealing with a special sidelobe problem and that existing suppression techniques based on relative amplitudes are inadequate, particularly when signal timing is also critical. Brooke-Footitt[5] approached the problem by using a pre-suppression pulse pair, preceding the main interrogation and transmitted in the direction of the target when the main aerial is expected to make a reflective interrogation. This could be effective in a simple static environment where one can learn the reflection problem and build in the necessary azimuth switching but it has the disadvantage that it does not cope with the ephemeral reflector, manoeuvring aircraft for example, nor with the exceptional very long delay reflection for which P1 and P3 arrive at the transponder after suppression has ended. One may, of course deliberately delay the presuppression signal to accommodate this, but it introduces a further complication to the adaptability. Another drawback of this scheme is that it increases the number of suppressed transponders thereby reducing their detectability to other interrogators.

It is interesting to observe that if one considers the old two-pulse s.l.s. in which P1 was radiated omnidirectionally, a reflection delay imposed on the directional P3 causes the separation between P1 and P3 to increase, and if this exceeds 0·5 μs or so, the transponder will not recognize a mode pair and therefore not reply. Tests made at RSRE have shown that about half the reflections observed there may be eliminated by using an omnidirectional P1 (Fig. 7). However, a P3 sidelobe, if present, would elicit a reply.

A number of modified formats have been examined, in which extra pulses are radiated to overcome the limitations of three-pulse s.l.s. in a reflective environment. Most of these are non-starters because conditions can be anticipated whereby a transponder may reply to a different mode to that intended, or suppression induced unnecessarily.

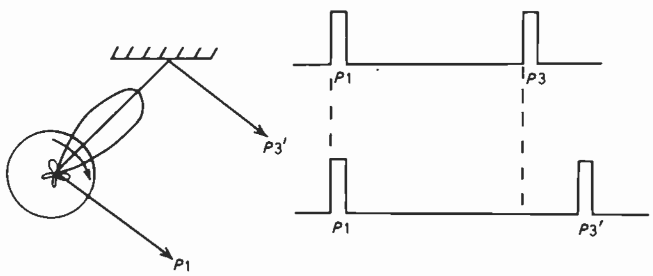


Fig. 7. Effect of omnidirectional P1 with directional P3.

In one proposal examined at RSRE, a pulse was to be radiated in the side-lobe region at a time which would deliberately cause the transponder to respond to Mode A when Mode C was actually interrogated. Only aircraft outside the main beam would so respond if a main beam reflection was also present, but the difficulty of guaranteeing a deep null in the omnidirectional beam along the main beam axis renders such an arrangement impracticable without added complexities. The plot extractor would need to be modified to accommodate and recognize the interlace pattern so developed and would not be suitable for any existing installation.

A second scheme showed more promise and was examined in more depth. In this proposal, P2 is replaced by another pulse, designated a screening pulse S1, timed 1 μs before P3, and radiated omnidirectionally (Fig. 8).

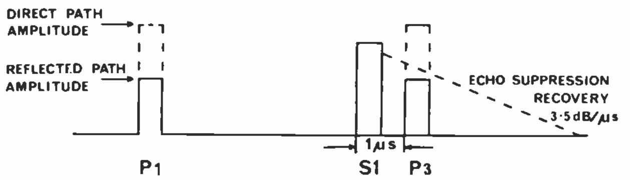


Fig. 8. Use of screening pulse S1.

Its purpose is to reset the transponder echo-suppression level so that a sidelobe or reflected P3 becomes screened with the result that the transponder makes no reaction, either to reply or suppress. It was found initially that a test transponder behaved as predicted when P3 did not exceed S1 by more than 6 dB, and main beam interrogations would be unaffected. The advantages of non-suppressive sidelobe cancellation are self-evident and it was found in practice to perform as effectively as normal P_2 suppression. Because the echo-suppression action persists over a period of time specified in Annex 10 by a recovery rate of 3·5 dB/μs and maximum recovery time of 15 μs, it was expected that reflection delays imposed on P3 would be less restricting on its behaviour with reflections than normal suppression.

The system was tested using P2 and S1 on alternate interrogations and it showed great promise in that a subjective assessment suggested a 50% reduction in the number of reflections. Unfortunately, it was observed that an occasional transponder failed to reply at all, even in the direct beam and further investigation showed that some transponders may be completely silenced by a signal appearing above minimum threshold if timed earlier than 0·5 μs in advance of P3. For this reason, the scheme cannot be progressed but it has opened a possible lead for the future if non-suppressive sidelobe and reflection cancellation schemes were to be considered by ICAO, whilst some shortcomings in transponder design have also been highlighted.

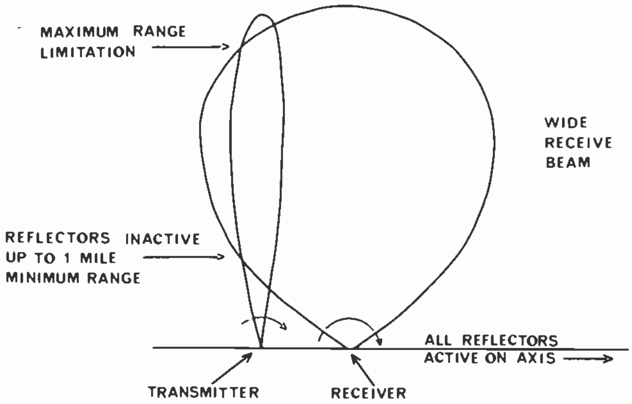


Fig. 9. A configuration for bistatic SSR.

Another method of interrogator control which has received some attention is that of controlled reduction in interrogation power, the concept being to exploit the losses expected on reflection. As yet, there is little experience of this method in the UK, but it is a simple modification applicable to appropriate sites which do not suffer from deep vertical lobing.

3.1.6 *Transponder modification*

It would be conceptually possible to recognize a reflective interrogation from the aircraft by noting the apparent direction of rotation of the interrogator. With *a priori* knowledge that an aerial rotates clockwise, a reflected image of the aerial would appear to rotate anti-clockwise. To make this recognizable requires a complex airborne array and signal processor but the scheme is of dubious value because many reflectors are not large enough to behave differently from a point scatterer and a complete image is not possible.

Without the force of international agreement, any modification to transponders or their aerials to recognize reflective interrogations cannot be considered as practical, although other schemes have been proposed.[6]

To conclude this Section on up-link reflection elimination, it would seem that little can be done at present to perfect an electronic solution. The use of the screening pulse seems promising, but unless shortcomings in existing transponder specifications can be overcome there will be unacceptable problems.

3.2 Down-Link Methods

For the reply path, anti-reflection measures may be applied in:

siting;
reducing the cross-section of the reflector;
improving aerials;
recognizing particular characteristics of reflected signals.

We may note that three items, namely siting, aerial design and reflector cross-section reduction, would be effective for both up and down paths and have already been referred to.

3.2.1 *Reflection recognition by a bistatic arrangement*

A system which was considered as exploiting the reflector characteristics is a bistatic arrangement in which the receiver is associated with an aerial separated from the interrogator transmitter aerial (Fig. 9). Both rotate synchronously and the idea was that the reflection pattern from the one aerial is different from that of the other so that neither see the same reflectors simultaneously. In principle this seems attractive but there are several problems which render such an arrangement ineffectual. These are associated with the aerial beamwidth necessary to permit common cover down to minimum range; the complexity of plot processing to accommodate apparent target range variations, and the reduced efficacy against reflectors lying on the joint axis of the aerials. Overall, it is the polar characteristics of the reflector which would determine the performance and these are by no means uniform. Estimates of the aerial separation required at RSRE (unpublished) show this to be about 5 km, suggesting a non-rotating omnidirectional receiver aerial with its limited range and interference problems. Taken to its limits, the concept becomes a multi-station integrated system which is what is already in existence, but it does not resolve the reflection problems at each site.

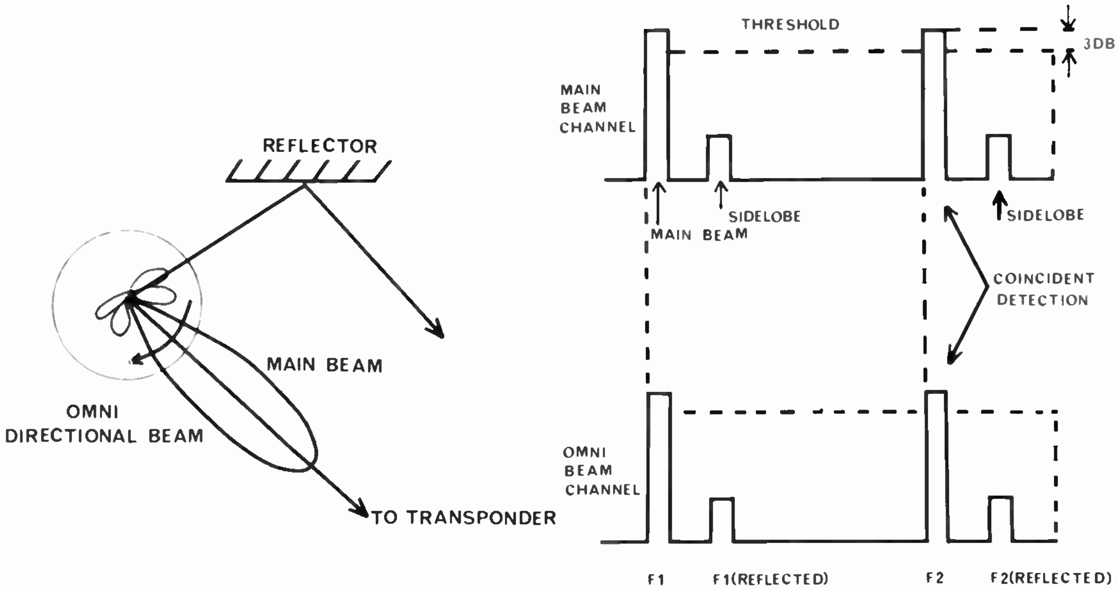
3.2.2 *Reflection recognition by hardware*

Reflection detection using hardware may be conceptually applied to recognize whatever parameters are deemed to be modified when signals are reflected. These are:

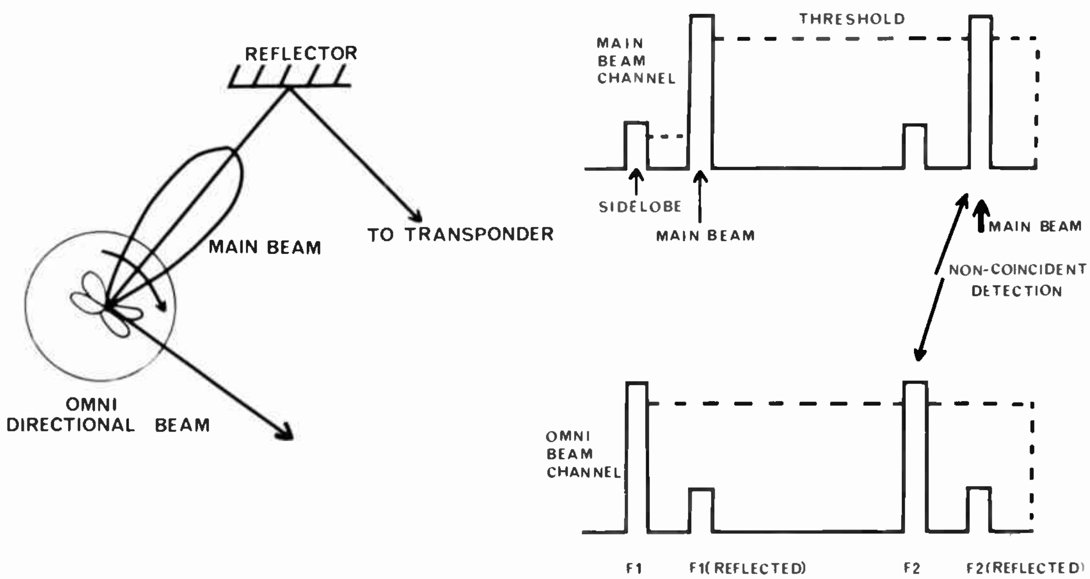
de-polarization;
elevation angle of arrival;
amplitude;
time of arrival.

The first of these would be extremely variable and difficult to measure satisfactorily and would probably only be significant where sloping reflector surfaces dominate. The second implies a form of elevation monopulse and presumes that all reflectors are at low elevation angles and consequently only particular reflections would be recognizable. A further drawback is that a high building would reflect an image at a similar elevation angle to the true target. Sharp bottom edge cut-off aerials offer a cheaper form of elevation discrimination for the circumstances which may generally apply.

Amplitude discrimination is already available in operational systems by virtue of the sensitivity time control (s.t.c.) or swept gain circuits. On-route radars are improved less than terminal radars where targets are at short range and s.t.c. attenuation is larger. Tests made at London (Heathrow) show at least a 90% reduction in the number of false plots as the s.t.c. is switched into operation. Adaptive s.t.c. represents an advancement which allows



(a) Target directly interrogated.



(b) Target reflectively interrogated.

Fig. 10. Operation of reflection detector with adaptive threshold circuits.

the initial attenuation to be programmed to vary according to the azimuth of the aerial.

The one factor which is inherent in all reflections is that the path length is always longer and the time of arrival must be later than that received directly from a transponder.

To implement the detection of the relative times of arrival, an omnidirectional aerial is required to ensure that a directly received reference signal is available from the transponder, even though the main beam is pointing to the reflector and therefore receiving a delayed signal.

A processor developed at RSRE measures the time difference on a signal resulting from successful bracket pair detection in each channel. A figure of 50 ns was chosen since this represents the highest common factor of bracket pair separation and pulse lengths, whilst permitting these to vary within their specified limits.

Amplitude discrimination is also included by incorporating adaptive threshold circuitry which automatically rejects all but the strongest signals in a 25 μs period commencing at F1. These latter circuits were originally developed to overcome a within-beam multipath

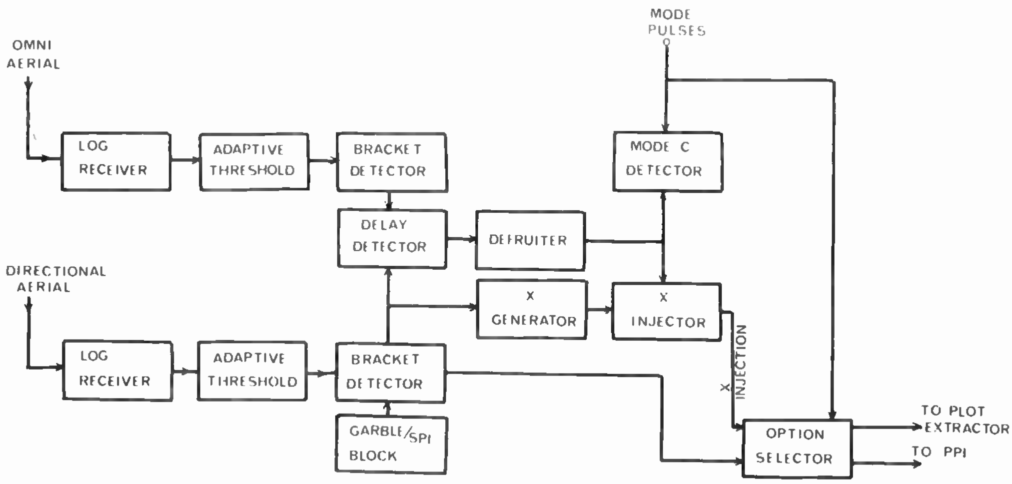


Fig. 11. Transponder reflection detector. Basic block diagram.

problem at a mountain top radar, by suppressing weaker, delayed signals which were synchronously garbling the main responses.[7]

Figure 10(a) shows the action. A signal arriving directly from a transponder appears simultaneously in both channels and no delay detection can occur. If delayed reflections are received via sidelobes, they may be expected to be weaker than the directly received signals and will be suppressed by the adaptive threshold circuit.

If the main beam interrogates via a reflector (Fig. 10(b)), there is an initial delay imposed by the up-link path and therefore common to both channels. The down-link signals processed by the omni-channel are otherwise identical to those for a directly interrogated transponder, but the main beam signals are received via the reflector and are delayed further. Provided these latter are larger than any directly received sidelobe signals, they alone will be fed to the delay detector. The situation for which the sidelobe signals are larger than the main beam reflected signal may be accommodated by standard receiver sidelobe suppression techniques using the omni-signal as a reference since the latter should always be larger than any sidelobe.

The circuit arrangement is shown in Fig. 11. The main beam and omni-channels comprise log-receivers, followed by adaptive threshold circuits and bracket pair detectors.

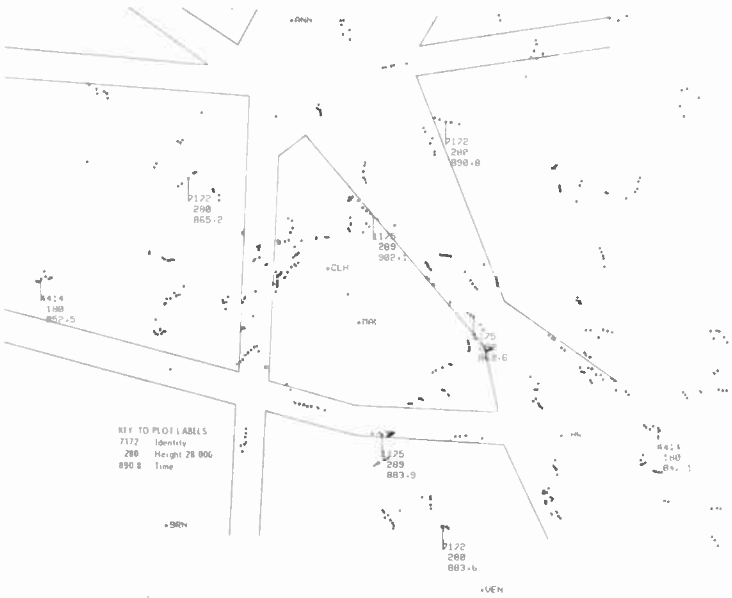
The log-receivers are essential to preserve relative amplitude. The adaptive threshold circuits are combined with pulse defination circuits and produce fixed logic level pulses of duration equal to the 3 dB width. The threshold rises to within 3 dB of the largest signal present for a period of 25 μs to cover the time from F1 to SP1. Any weaker signals occurring during this time (from sidelobes, fruit, garbling, interference or delayed within-beam reflections†) are suppressed and do not appear at the output.

The bracket detectors that follow are comprised of 416 bit shift registers clocked at 20 MHz and therefore have a total delay of 20·8 μs and a resolution of 50 ns. A selection of parallel outputs near the input and output are AND gated to give an output for any bracket pair complying within the limits permitted by Annex 10. The main beam bracket detector includes a lock-out to prevent certain code combinations or garbles causing a detection which may be interpreted as reflection.

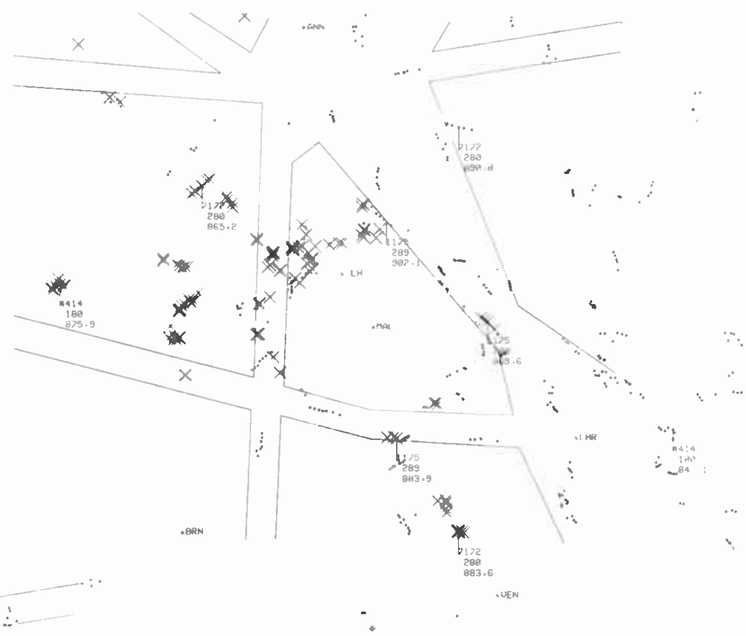
The delay detector operates by generating an output if the leading edge of the main beam signal appears more than 50 ns after an omni-signal, with an optional maximum of 25 μs. A facility is included to allow the display of the actual delay time which would assist in the accurate plotting of reflector position.

Having made a decision as to whether a reflection has occurred, the delay detector must make the fact known to the radar operator or the plot extractor. For a p.p.i., the simple expedient of tagging the response with a second pulse has been found adequate. An occasional detection in an otherwise clear run is visual evidence that the target is not to be interpreted as a reflection. The possibility of false accusations due to fruit or other effects has been reduced by the 'defruiter' following the delay detector. At present this operates on a 2 out of 2 or 3 out of 3 basis. Any hardening beyond this level would reduce the likelihood of detections when presented to a plot extractor. The defruiter operates out to 120 n.m. and employs a single chip 4K static r.a.m. addressed at 2·5 MHz.

† 'Fruit' = asynchronous interference due to other interrogators. 'Garbling' = synchronous interference between closely-spaced aircraft.



(a) Normal.



(b) 'X' = reflection.

(c) Reflections deleted.

Fig. 12. Plot displays.

Most applications would require the plot extractor to be fed with a suitable signal to indicate a reflection. Assuming that no internal modifications to the plot extractor are permissible, the option selected has been the deliberate injection of an 'X' pulse into the data stream input to the plot extractor. It must be noted that no other modification is applied to the data apart from an inherent overall delay, and in the basic form, some garble repression arising from the adaptive threshold technique. Where this is not desirable, as in a terminal area radar, the data stream comes directly from the main beam receiver with an equalizing delay. On receiving a plot containing an 'X' pulse, the plot extractor passes on the indication in the target report and this may be used by the software to mark up a suitable symbol on the labelled plan display. 'X' pulse injection does not invalidate the reply in Mode A, but Mode C is, or may be, invalidated by the plot extractor, although target detection is not affected. It is possible that spurious 'X' pulses arising either directly from the originating transponder, by garbling or by injection may cause an indication of a reflection when in fact it is not and although the defruiter helps to reduce the eventuality, experience shows that some wild indications are still given.

The manner in which the reflections are labelled may be supplemented by deliberately delaying the suspect replies by 3 μs or so, thereby placing them in a different range cell. An un-typical indication will then be rejected by the plot extractor rather like defruiter action, whilst the rest of the run will be unaffected but leaving a small error in plotted azimuth. A reflection which is confirmed by all or most of its replies being delayed will be reported as a target at a slightly increased range but the 'X' pulses will indicate that a correction is to be applied.

A useful improvement could also be gained in 'ring-around' situations resulting from a multiplicity of nearby reflectors surrounding the radar site. Target detection is then normally difficult because of the excessive run-length. By using the delay technique, the major portion of the ring-around is placed in a separate range cell from the replies received directly from the transponder and giving an opportunity of displaying the true target position.

Figure 12 shows display situations containing 500 plots gathered over about 6 aerial revolutions. Some plots have been labelled and two identities are triplicated showing that some of them are reflections (Fig. 12(a)). Figure 12(b) is the same situation with about 30% of the plots labelled with a cross indicating a suspected reflection. If desired all of these could automatically be deleted as in Fig. 12(c), but it is considered incautious to do so during evaluation.

3.2.3 *Reflection recognition by software*

Hardware and software solutions are complementary; it is usually assumed that the latter will use whatever data are already available from existing receivers but the hardware fix provides extra information for the software to act upon. Indeed, a hardware method requires software to apply labelling to suspect plots, although complete deletion could be implemented independently. J. M. Shaw and K. J. Fooks (RSRE) have developed a software solution which follows a set of rules:

(1) a reflected plot will be accompanied by another real plot with

(a) identical code,

(b) an extrapolated range equal or less than the suspect plot,

(c) a similar height,

(2) a reflected plot has no primary radar correlation,

(3) the calculated reflector range is realistic.

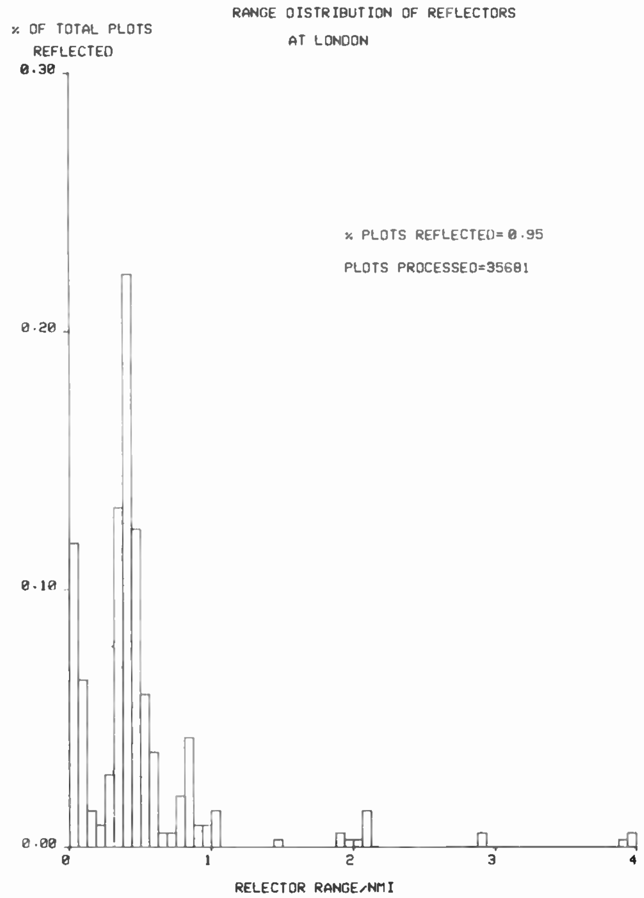
The detection of a reflection permits the listing of position and orientation of the reflector and this memory is used to support further decisions on suspected plots. The learning ability is adaptive to allow the automatic addition and removal of reflector characteristics from the list to allow for ephemeral reflectors. The package includes facilities for plotting histograms of the distribution and intensity of reflectors as a function of range and azimuth as illustrated in Figs. 13(a) and (b), for data collected at London (Heathrow). An interesting point revealed here is that for the traffic conditions prevailing during this time, one reflector, at 42°, is responsible for about a third of all detected reflections, which total nearly 1% of all plots. In contrast, an azimuthal histogram for Malvern is shown in Fig. 14 where some 30% of the plots are declared false.

4 The Final Decision

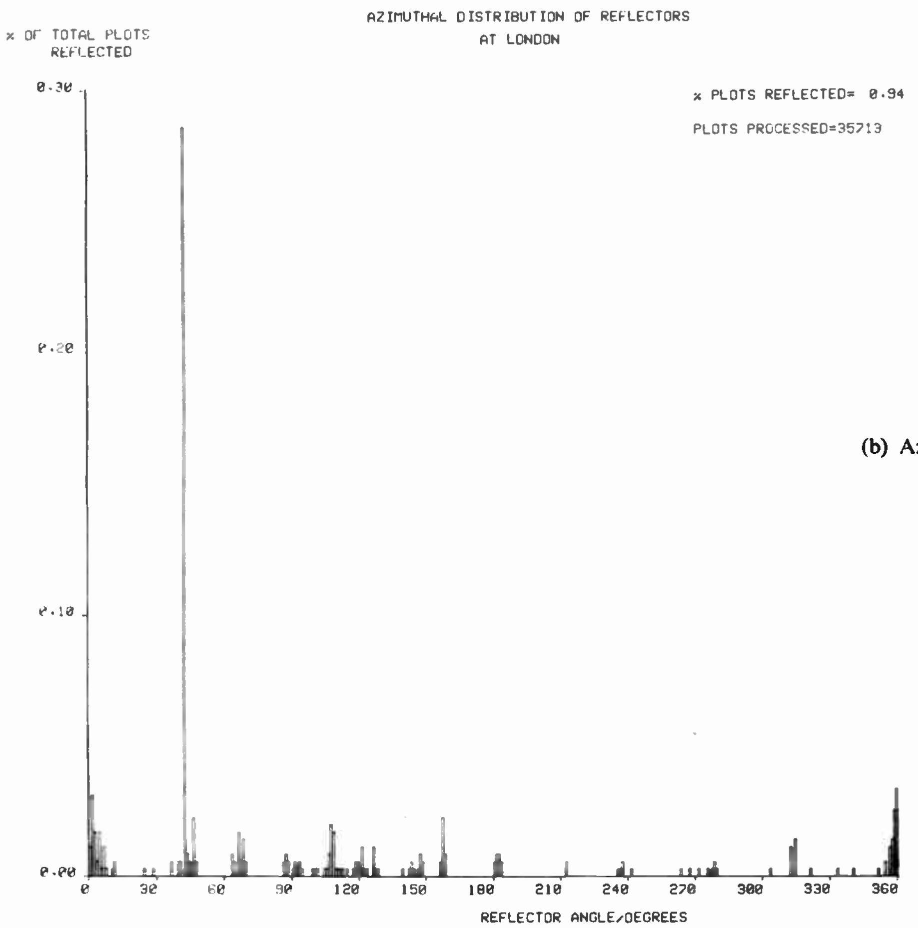
Overall, apart from sensible site evaluation and appropriate aerial design, there appear to be only two effectual remedies for existing sites, using either hardware or software reflection detectors, or both, in the down-link receiver system.

The question inevitably arises as to which of these two approaches is the best. It is probably too early to give a clear cut answer and to some extent is dependent on the installation it is to be applied to: a radar having no plot extractor but only raw video could only use a hardware system.

A final answer could only be given after it had been determined which process gave the highest proportion of false target detections or the minimum number of falsely accused genuine plots. Indications so far are that at least 90% of reflections are detectable by each. A hardware system requires extra maintenance and procedures for routine testing, but is relatively easily installed with existing equipment. It is potentially capable of detecting reflection delays which are very small but only to a maximum range limited by the signal/noise ratio of the omni channel. Its indications are however instantaneous. A software system alone can only be taken to a



(a) Range distribution of reflectors at London.



(b) Azimuthal distribution of reflectors at London.

Fig. 13

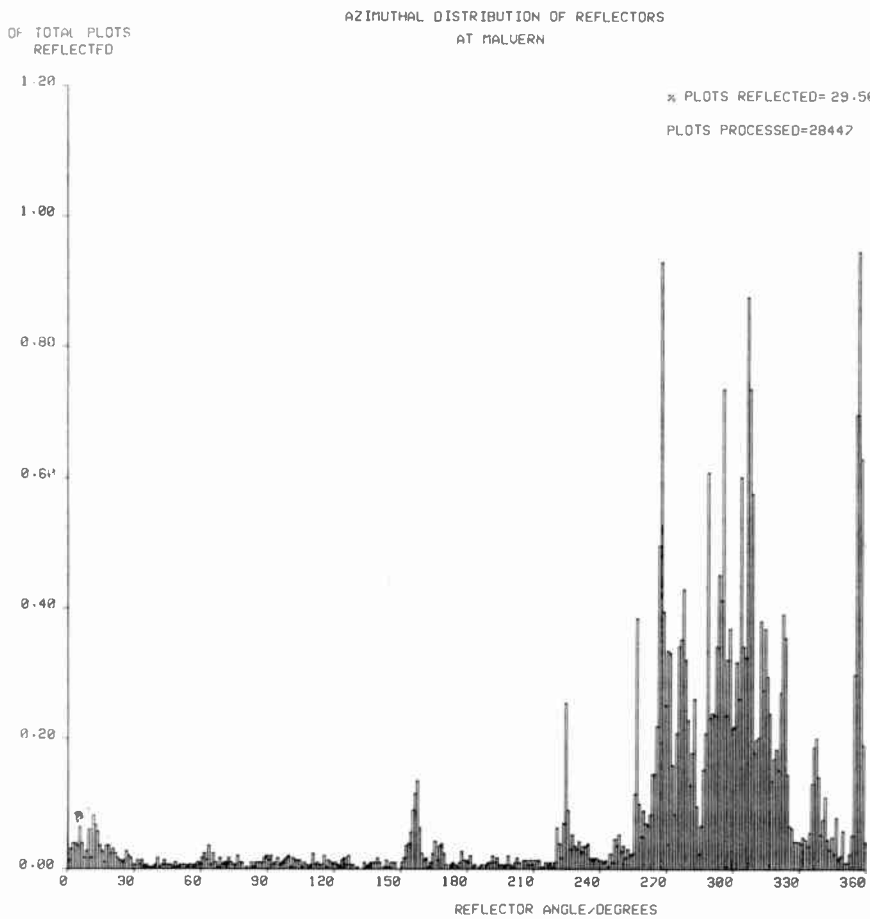


Fig. 14. Azimuthal distribution of reflectors at Malvern,

level of optimization dependent on cost. It is capable of detecting reflections out to any range but the plot extractor range cell size limits the resolution and its indications take time to establish. The best overall concept would be to consider the two as complementary, with the ability to place a high degree of certainty on those plots which both systems agree to be false (to the extent of permitting complete extinction) and a level of suspicion on those which are in dispute.

5 Acknowledgments

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