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Broader Horizons

PROFESSIONAL engineering institutions have always placed considerable value on their 'learned society' activities and especially on the holding of meetings for the presentation and discussion of papers. This is because, historically, institutions have evolved from the scientific societies which were formed in the days when communication through personal contact was the most efficient method of disseminating information. But today everyone has broader horizons which, coupled with the use of printed material and the attractions of leisure activity, militate against exchanges of information by personal contact: the role and effectiveness of the technical meeting must therefore be re-assessed.

Several kinds of technical meeting are in vogue. In increasing order of complexity, there are the evening meetings at which a single paper is read, the colloquium which groups three or four short papers, the half- or whole-day symposium with half a dozen or more papers on a rather broader theme, and finally, the two- to four-day conference, possibly international, and conceived on a scale ranging from the highly-specialized exploration of a theme in considerable depth to the broad coverage surveying progress, or promise, in a whole branch of technology.

Judging the value of a meeting by attendance indicates that the colloquium, the symposium and the conference are favoured, always provided they are timely and have a sound content. It is sad to say, however, that in the experience of most professional bodies, the evening meeting, at which one paper is read, nowadays seldom attracts the numbers recorded even a decade ago. Many reasons have been advanced to account for this decline, although growing 'specialization' seems to limit attendance of those who are only interested but not actively engaged in a specific field. Subjective considerations such as difficulty of travel, counter-attractions of further education, demanding leisure-time activities, and of course indifference or apathy, all have to be borne in mind.

It is perhaps crying in the wilderness to say that all these considerations ought to be subordinated to the advantages of professional association!

The value of conventions or conferences which provide opportunity for discussing many papers and more time for contact with a number of co-workers is well recognized. Yet another way in which engineers may meet away from the everyday pressures is in the technical visit to a laboratory, factory or operational establishment. If the place visited is in another country the pleasure most of us experience in 'being abroad' heightens the professional benefit. The evening meeting, apart from the obvious opportunity for discussing the single paper informally, presents less time perhaps, but nevertheless more frequent opportunity to meet fellow engineers.

Whatever means are preferred in order to *meet* professional colleagues, it is certain that the printed word alone does not answer the desire of members to participate in 'learned society' activities. How best therefore, to organize personal participation to please the majority of members requires the exercise —almost minimal—of each member to indicate to his national division, local section or even Institution headquarters his preference for type of meeting. If this freedom to influence 'events' is coupled with evidence to participate—perhaps by submitting a paper?—then the twentieth century engineering institution will continue to develop and provide the service which is demanded in an age when the individual has constantly broadening horizons.

G. D. C.

The Institution's 1968 Convention

'Electronics in the 1970s'

King's College, University of Cambridge, Monday, 1st July to Friday, 5th July, 1968

Now that the complete programme has been announced, the broadly ranging theme of the Convention can be appreciated. Its intention is to provide a forward look at the developments of the next decade and several of the Convention sessions have the specific aim of enabling senior engineers to become more familiar with some of the work on 'the frontiers of electronics', which will soon be moving from the 'R & D' stage. Other sessions review the 'state of the art' in particular branches of electronic engineering and discussion at these sessions should be especially valuable.

Applications for accommodation during the Convention at King's College can still be accepted—the charge for the full period (Monday dinner to Friday lunch inclusive) is £18. The registration charge is £5 for members of the I.E.R.E., and £8 for non-members; this covers provision of all available preprints as well as attendance at sessions, and there is not a separate part-time registration charge. Requests for registration forms should be made without delay to the Convention Registrar, I.E.R.E., 8–9 Bedford Square, London, W.C.1. (Telephone 01-580 8443, extension 3).

PROGRAMME AND SYNOPSES OF PAPERS

Monday, 1st July

16.00 onwards. Registration in the Convention Office in King's College.

Tuesday, 2nd July

09.15 **Formal Opening of the Convention by Major-General Sir Leonard Atkinson, K.B.E.,** President of the Institution.

09.30–12.30 **Session 1: Symposium on COMPUTERS AND AUTOMATION**

'Progress in On-line Control by Computer'—D. Best, O.B.E.

'The Application of Computers in Remote Indication and Control'—H. D. Mitchell and W. Renwick.

'Current Developments in Computer Hierarchies for Industrial Control'—S. L. H. Clarke and C. Ayers.

'Simulation Techniques for Traffic Studies'—Dr. M. G. Hartley and E. T. Powner.

'The Representation of an Electronic Circuit Diagram in Digital Form to Permit Drafting by a Digital Computer'—W. F. Hilton.

14.15–17.15 **Session 2: Symposium on ELECTRONIC ENGINEERING EDUCATION**

'Postgraduate Education and Training and the Universities'—Professor G. D. Sims.

'The Role of the Polytechnics'—N. L. Garlick.

'Training after Graduation'—Professor W. E. J. Farvis.

'The Influence of Microelectronics Developments on Education in the 1970s'—Dr. B. H. Venning.

'The Purpose of Teaching Microelectronics'—D. F. Dunster and T. Wilmore.

Wednesday, 3rd July

09.00–12.30 **Session 3: SURVEY PAPERS**

'Applications of Lasers in Electronic and Radio Engineering'—Professor W. A. Gambling and Dr. R. C. Smith.

'High Power Ultrasonics in Industry'—A. E. Crawford.

'Underwater Acoustics: A Review of Progress'—Professor D. G. Tucker.
'Electronics in Oceanography'—R. Bowers.

14.15–17.15 **Session 4: Symposium on** AUTOMATIC TEST EQUIPMENT

'An Introduction to Automatic Testing'—Colonel R. Knowles, R.E.M.E.
'Shipborne Automatic Test Equipment in the Royal Navy'—R. L. Short.
'The Requirements for A.T.E. in the Army Environment'—Lt.-Col. G. Beswick.
'The Use of Automatic Test Equipment by the Royal Air Force—Squadron Leader E. P. Folland.
'Automatic Test Systems for Production and Maintenance'—A. H. Parker.
'Trends in A.T.E. Technology'—D. R. Price.
'The Organization of Test Sequences for Automation'—E. W. Carr.
'The Use of Standard Computer Programs for Equipment Diagnostic Techniques'—J. K. Skwirzynski.
'Automation of Fine Measurement Procedures'—Dr. J. Rawcliffe.

17.30–18.30 **The Sixth Clerk Maxwell Memorial Lecture—M. Maurice Ponte**

Thursday, 4th July

09.30–12.30 and 14.15–16.00 **Session 5: Symposium on** COMMUNICATIONS

'Global Communications: Current Techniques and Future Trends'—R. W. Cannon.
'The U.K. Telecommunication Services'—J. S. Whyte.
'The Impact of Pulse Code Modulation on the Telecommunications Network'—Professor K. W. Cattermole.
'A Communication Network for Real-Time Computer Systems'—D. W. Davies.
'The Control System of the National Grid and its Communication Links'—P. F. Gunning.
'Future Sound Broadcasting Techniques'—Dr. K. R. Sturley.
'Some Thoughts on Future Television Broadcasting'—A. V. Lord.

16.15–17.15 **Session 6:** ELECTRONICS IN THE CAVENDISH LABORATORY

Contributions will be made by Dr. D. M. A. Wilson, D. W. J. Bly and H. Bett on 'Electronics in the Cavendish Laboratory' including the work of the Mond Laboratory (low-temperature physics), the electron microscope group, the meteorological physics group and the radio astronomy group.

19.30 **Convention Banquet in King's College**

Friday, 5th July

09.00–12.30 **Session 7: Symposium on** FUTURE MATERIALS AND COMPONENTS

'Film Electronics in the Seventies'—J. Maddison.
'The Monolithic Linear Integrated Circuit'—Professor W. Gosling.
'The Interconnection of Integrated Circuits'—Dr. S. S. Forte.
'The Impact of the Burghard Scheme'—T. M. Ball.

14.15–16.30 **Session 8:** ELECTRONICS IN THE CAMBRIDGE UNIVERSITY ENGINEERING LABORATORY

An introductory talk by Professor C. W. Oatley, followed by a tour of the Laboratories and discussion.

Synopses of some of the Papers to be presented at the Conference

Session 1: Symposium on COMPUTERS AND AUTOMATION

Current Developments in Computer Hierarchies for Industrial Control

S. L. H. CLARKE, B.A., C.ENG., F.I.E.R.E. (*Elliott Process Automation Ltd., Boreham Wood, Herts.*) and C. AYERS, B.SC.(ENG.), A.M.I.C.E., M.I.MECH.E, M.I.E.E. (*Power and Marine Division, English Electric Co. Ltd.*)

This paper reviews the present trends in the use of computers in process control. Examples quoted to show the wide range of application which has to be covered include steel production, electricity generation, chemical process control and gas production. Conclusions are drawn about the effect which these requirements will make on the design of suitable equipment in the next few years.

Session 2: Symposium on ELECTRONIC ENGINEERING IN EDUCATION

Postgraduate Education and Training and the Universities

Professor G. D. SIMS, PH.D., M.SC., D.I.C., C.ENG., F.I.E.E., F.I.E.R.E. (*Department of Electronics, University of Southampton*)

At the present time many factors are causing Universities and Colleges to revise their attitudes to postgraduate work. Many of the conventional one-year M.Sc. courses receive insufficient numbers of applicants to make them viable and still less do they excite the support of the industries which their products will eventually serve.

Drawing support from the facts and opinions expressed in a number of Government reports, industry has attacked these courses strongly and is even more sceptical about the virtues of traditional University Ph.D. schemes. A consequence of this situation is that in many Universities a considerable amount of re-thinking has taken place and many new and constructive ideas are beginning to crystallize. It is nevertheless far from clear that the needs of industry and the real problems of University/Industry collaboration have been sufficiently well identified. An attempt is made in this paper to survey the present position in regard to postgraduate activity in the Universities and, in the light of the above observations, to outline paths along which future developments might proceed.

The Role of the Polytechnics

N. L. GARLICK, M.SC., C.ENG., F.I.E.E., F.I.E.R.E. (*Vice-Principal, Brighton College of Technology*)

The courses to be offered by the proposed polytechnics will have developed from the former Diploma in Technology courses which were established a decade or so ago under the ægis of the National Council for Technological Awards, and which have now become Degree courses under the Council for National Academic Awards, together with such courses as Higher National Certificate and Diploma. If the polytechnics are to fulfil their intended function it is necessary to avoid duplication of university degree courses and to cater for students who are not at present obtaining places in universities. The present close liaison between colleges of technology and industry must be developed and further extended and facilities must be provided to enable students to obtain qualifications which are appropriate to their ability and therefore transfer between one course and another should be facilitated. It is suggested that the curricula should be very broadly based in the early part of the course and should be much more liberal than is the situation in present courses. In the later part of the course, there should be increased specialization in order to ensure that students have the opportunity of studying to a depth appropriate to the award which they hope to achieve, even although this may mean restricting their breadth of technological study at this stage.

The Influence of Microelectronics Development on Education in the 1970s

B. H. VENNING, PH.D., B.SC. (ENG.), A.C.G.I., C.ENG., F.I.E.E. (*Brighton College of Technology*)

It is obvious that the advent of integrated circuits is already having a considerable influence on the pattern of work undertaken by electronic engineers and that this trend will accelerate in the next few years. System design, utilizing microelectronic modules, is replacing the traditional piece-part circuit construction and educationalists concerned with all levels of work must see that appropriate changes are made. They must also recognize that an increasing number of students will be coming forward with a more modern approach to mathematics, and that the increasing use of computers must make the teaching of many of the traditional analytical procedures unnecessary. Teaching methods too must be reviewed and the means of mass communication used to keep trained engineers in touch with new developments.

The Purpose of Teaching Microelectronics

D. F. DUNSTER, B.SC., C.ENG., M.I.E.E., A.INST.P. and T. WILMORE (*West Ham College of Technology, London*)

Electronics is taught today in basically the same way as it was when valves were the main active elements. Transistors used as separate devices can now be seen to be a transitional phase leading to a mainly micro-electronic future. Thus all those engineers who will need to learn electronics in the future will have to understand the integrated circuit form. This understanding will be best obtained if experimental facilities are available in a semiconductor technology laboratory, which is organized for teaching, not for research.

Simple planar transistors may be fabricated by undergraduates without much difficulty and postgraduate students could make rudimentary integrated circuits. Semiconductor devices other than transistors are being developed in a steady flow, for instance Gunn diodes and thyristors, and a thorough semi-quantitative knowledge of the properties of semiconductor materials and junctions is required if engineers are not to be fobbed off with non-explanations for each device. These requirements go far beyond the scope of present-day 'Physical Electronics'.

Session 3: SURVEY PAPERS

Applications of Lasers in Electronic and Radio Engineering

Professor W. A. GAMBLING, B.SC., PH.D., C.ENG., F.I.E.E., F.I.E.R.E., and R. C. SMITH, PH.D. (*Department of Electronics, University of Southampton*)

Lasers have indirectly made a considerable contribution to the understanding of semiconductors since the study of their bulk interaction nature has pointed the way to new high-power, high-frequency semiconductor devices. The direct application of lasers in radio and electronics is likely to be considerable. The increasing effort being placed in this area is beginning to produce promising results, and applications in ranging, information processing and pattern recognition, optical logic, high-density optical stores and communications will be discussed.

Underwater Acoustics: A Review of Progress

Professor D. G. TUCKER, D.SC., PH.D., C.ENG., F.I.E.E., F.I.E.R.E. (*Department of Electronic and Electrical Engineering, University of Birmingham*)

The paper attempts to review the whole field of civil work in sonar and underwater acoustics, discussing in particular the problems of short-range high-resolution sonar and sonar for fisheries, the influence of micro-electronics on sonar philosophy, the growing importance of the exploitation of non-linear acoustic wave interactions in sonar applications, and propagation, transducers and arrays. It seems clear that civil applications of underwater acoustics are growing rapidly, and that recent developments will lead to sonars of very high performance becoming available at a cost little (if any) greater than that of existing simple commercial equipments.

Electronics in Oceanography

R. BOWERS, B.SC. (*National Institute of Oceanography, Godalming, Surrey*)

Electronics now plays a role of paramount importance in oceanography. This has not always been so; until very recently many measurements in oceanography were not trusted to electronic devices but this has all changed and it is now difficult to think of many oceanographic instruments which do not use some form of electrical instrumentation. This change has been brought about by the introduction of circuits which are very reliable, use little power, and give as good or better resolution and accuracy as classical methods, and by the employment of a greater number of electronic engineers to apply these techniques. This paper primarily describes electronic instruments used to measure parameters of the water column of the oceans with the exception of the very important field of underwater acoustics which is being covered in Professor Tucker's paper.

Session 4: Symposium on AUTOMATIC TEST EQUIPMENT

The Requirements for A.T.E. in the Army Environment

Lt. Colonel G. BESWICK, C.ENG., M.I.E.E., M.I.MECH.E. (*H.Q., R.E.M.E., London*)

This paper examines the Army's repair philosophy in the light of equipment likely to be in use in the 1970s. It is expected that the equipment then in use will be more sophisticated and in greater variety than at present and that due to manpower problems a form of automated testing will be essential. It is shown that multi-system automatic test equipment will be more suitable for Army needs than special system test equipment. The use of a computer as a controller for automatic test equipment is investigated and attention is given to the special environment in which Service equipment may have to operate.

The Use of Automatic Test Equipment by the Royal Air Force

Squadron-Leader E. P. Folland, B.Sc. (*Graduate*), (*Ministry of Defence, Directorate of Electrical Engineering, R.A.F.*)

The paper will show very briefly how the R.A.F. extended the use of automatic test equipment from its original concept as missile check-out equipment to testing aircraft systems; further extension now is to testing aircraft avionics units in a workshop environment. Currently the intention is to extend the application to factory repair organizations within the service. In the course of twelve years as a user, many important lessons have been learned. The major problem areas are defined: such difficulties as test specifications which realistically reflect test necessities; the control and administration of modifications to avionics units which affect testing; the need to establish realistic program proving schedules; and providing diagnostic facilities, are all dealt with in some detail. The paper concludes that as much attention needs to be directed towards administrative and organizational problems as is currently devoted towards the technical problems, and outlines the present thinking in the R.A.F. for a future A.T.E., using computers sophisticated software and hardware probably similar to the Army M.A.T.E., extending capabilities into the r.f. field, at least to u.h.f.

Shipborne Automatic Test Equipment in the Royal Navy

R. L. Short, B.Sc., C.Eng., M.I.E.R.E. (*Admiralty Surface Weapons Establishment*)

At sea human efficiency is impaired and maintenance, now becoming highly complex, becomes increasingly difficult. The shortage of trained maintenance staff could thus have a serious effect on ship survival. It is therefore becoming necessary to evolve highly flexible automatic test equipments to cover a wide variety of equipment types, designs and parameters. However, the cost-effectiveness gain resulting from A.T.E. must be proved, especially in the most efficient employment of highly skilled personnel. The requirements of typical missiles, radar and data handling systems are reviewed and the considerations which determined the design of a particular A.T.E. are described. Possible future development in the automating of automatic test are considered.

Automatic Test Systems for Production and Maintenance

A. H. Parker (*Electronic Equipment Department, Hawker Siddeley Dynamics Ltd., London*)

The paper considers the type of automatic testing currently used by operators and manufacturers in the avionic field. In particular the T.R.A.C.E. (Test Equipment for Rapid Checkout and Evaluation) systems are considered as independent test stations for relatively low throughput testing and as computer controlled multi-station installations where high throughput is being considered. An explanation will be given on how these systems operate, how they are chosen for particular applications and how they are used in practice, together with actual statistics achieved in current installations in the major airline companies.

Trends in A.T.E. Technology

D. R. Price, B.Sc., A.R.C.S. (*Elliott Flight Automation Ltd., Rochester, Kent*)

The application of automatic test equipment poses many problems in specifying the requirement when development of the prime system is going on and in tying down precise test data. The equipment must be as adaptable as possible to maximize cost-effectiveness. The advantages of computer control in providing adaptability are discussed. The choice between software and hardware solutions to a specific problem is not a clear one and some guides are given, together with pointers toward the choice of computer parameters.

The application of A.T.E. raises some intriguing problems of man-machine interface, in particular between the programmer and the machine, and between the machine and its operator. Developments in problem-orientated codes and the use of modern display techniques are discussed. Developments in computer-aided design offer promise for alleviation of the human interface problems between the designer of the prime equipment and the programmer.

Automation of Fine Measurement Procedures

J. Rawcliffe, Ph.D. (*The University of Manchester Institute of Science and Technology*)

The benefits are discussed of automation in the Standards and Calibration Laboratories on the one hand and in Research Laboratories on the other. Two particular approaches, one from each of the above fields, are then specified and some of the anticipated difficulties are mentioned. In particular, the methods of preparation of input data for an automatic calibration system are treated. The impact of automatic calibration systems on measurement laboratories staff and management is also discussed. Proposals are presented for overcoming some of the difficulties which might be encountered and the paper is concluded by a discussion of the results of early experiments relating to the projects.

The Organization of Test Sequences for Automation

E. W. CARR, D.C.AE. (*Honeywell Controls Ltd., Hemel Hempstead*)

Automatic Test Equipment can be used to reduce both the time required for test and the skill necessary for fault diagnosis, simply by automating manual test procedures. However, if full advantage is to be taken of the technique the established procedures must be reorganized.

The first step in this process is the grouping of manual intervention as far as possible and elimination of test point probing by utilization of the more powerful diagnostic routines which utilize external test points. Beyond this, most test sequences resolve themselves into a series of relatively simple go/no-go measurements to establish the serviceability of replaceable modules or sub-units. The usual method of arranging these is in logic order so that each diagnostic routine builds upon the assumption that all tests before it have passed. This gives the shortest total program, but not necessarily the shortest test time to find and repair the fault.

A method of arranging tests for automation is suggested which depends upon the probability of failure in the particular mode tested and the time taken to repair and retest (including the availability of spare components).

The Use of Standard Computer Programs for Equipment Diagnostic Techniques

J. K. SKWIRZYNSKI, B.SC., C.ENG., M.I.E.E. (*The Marconi Company Ltd., Chelmsford, Essex*)

A speculative study is made of ways for using standard network analysis programs for diagnosis of faults, of component deviations and for development of automatic, on-line, computer controlled diagnostic techniques.

It is assumed that a go/no-go manual procedure (or possibly semi-automatic procedure) is used to isolate faulty sub-systems; further, that analysis programs are available to produce frequency responses and time responses. The programs are then used to diagnose the fault. Special attention is paid to the problem of optimizing cost and time of measurement by compromising measurements at few nodes and many signal frequencies, and between measurements at many nodes and few frequencies. As a test of this method a cable corrector in a television camera unit is analysed.

Session 5: Symposium on COMMUNICATIONS

The U.K. Telecommunications Services

J. S. WHYTE, M.SC.(ENG.), C.ENG., F.I.E.E., (*Deputy Director of Engineering, P.O. Telecommunications Headquarters*)

The U.K. telecommunications system has been built up over many years and now forms a very extensive network. Broad band links on coaxial cable and microwave radio systems interconnect more than 150 towns and cities and by 1971 the total wide-band capacity is planned to exceed 170 000 MHz-miles of bothway transmission plant linking some 270 towns and cities. In the junction network the application of digital techniques is permitting multi-channel transmission to be economically employed over distances much shorter than has been possible in the past using analogue transmission and within two years it is planned to have more than 400 000 channel-miles of p.c.m. plant in use. The introduction of digital transmission into the junction network opens the door to new possibilities in the switching field. A variety of new services is in preparation and work in the research laboratories gives some pointers to the communication systems of the future.

Future Sound Broadcasting Techniques

K. R. STURLEY, PH.D., B.SC., C.ENG., F.I.E.E. (*Chief Engineer, External Broadcasting, British Broadcasting Corporation*)

Although television is claiming an increasingly large audience, there is still considerable interest in sound broadcasting and this is likely to continue. The 1970s should see more refined methods of testing and measurement of the performance of studios, and studio design itself should become more of an exact science by the development of the model technique using ultrasonic frequencies. Appreciable advances in acoustic absorber control materials do not seem likely, but it is possible that an electronic treatment may be developed as an alternative. A control point is required between studio and transmitters to link outside sources to the producer, to maintain a degree of continuity and to distribute the programmes. Such an operation is ideally suited to automation and considerable developments in this direction may be expected. It may be assumed also that except for high quality musical programmes, no manual control of volume will be exercised in the 1970s.

It seems probable that distribution will eventually be in digital form which will allow a higher signal-to-noise ratio as well as greatly reduce the distortion problem. For economic and other reasons it is unlikely that there will be any great change in transmitter aerial installations.

The Impact of Pulse Code Modulation on the Telecommunication Network

Professor K. W. CATTERMOLE, B.SC., C.ENG., M.I.E.E. (*University of Essex*)

The basic processes of transmission and switching are cheaper for multiplex digital signals than for their voice-frequency equivalents, and introduce little or no impairment of the message. Conversion between analogue and digital signals introduces much of the cost and almost all of the impairment in a p.c.m. communication system. The long-term plan for a telephone administration should therefore be to introduce an integrated digital network in which only a minimum number of conversions and reconversions is required: ideally only one per connection. A planned transition from present practices towards a digital network is possible, because the partial use of p.c.m. is technically and economically beneficial provided that the applications be carefully chosen. A digital network whose main traffic is telephony can also carry other messages such as telegraph and data.

Session 7: Symposium on FUTURE MATERIALS AND COMPONENTS

Film Electronics in the Seventies

J. C. MADDISON, B.SC. (*Electrosil Ltd., Sunderland*)

This is a general review paper considering the problems and economics of film circuits during the next ten years. It starts by considering why we need film circuits and what are their advantages over assemblies of discrete components and over s.i.c.'s. It then reviews the present types of film circuits and gives some possible suggestions on new types which may be developed during this period. Each of these is considered to try to see which functions it best fulfils with its present and probable future performance and price characteristics.

Consideration is given to the problem of how to combine active devices and s.i.c. with film circuits, and some thoughts are expressed on automatic methods of manufacture and on the optimum size of integrations, with special reference to reliability of the finished product and production problems of testing and burn-in.

The Impact of the Burghard Scheme

T. M. BALL, C.ENG., M.I.E.R.E. (*Elliott-Automation Microelectronics Ltd., Boreham Wood, Hertfordshire*)

'Burghard', or to give it its true title, 'General requirements of specifications for parts of assessed quality', has been the subject of much discussion. Yet it is true to say most engineers are unaware of its contents, or the implications of the system on the electronics industry.

The paper will trace the history from about 1963, when a committee was set up by the Ministry of Aviation under the chairmanship of Rear-Admiral G. F. Burghard, C.B., D.S.O., through to the present work by the British Standards Institution, on whose shoulders rests the responsibility of implementing the Burghard recommendations. Brief explanations will be given of the scheme as laid down in B.S. 9000 and B.S. 9001, and a current 'state of the art picture' given on British Standard Specifications so far issued under the new system.

Reference will also be made to the attitudes of other countries to the scheme and the work that has already started in preparation for a European System. Finally, the author will make some personal comments based on discussions with members of industry to illustrate why the system is taking a long time to implement.

Synopses of nearly all the other papers to be read at the Convention were published in the May issue of *The Radio and Electronic Engineer*.

An announcement about the availability of individual papers in separate form will be made after the Convention has taken place. Only selected papers will be published in the *Journal* and *Proceedings* of the Institution during the next few months.

U.D.C. 621.396.677

A Wideband Circular Array for H.F. Communications

By

I. D. LONGSTAFF, Ph.D.†

AND

D. E. N. DAVIES,
M.Sc., Ph.D., C.Eng., M.I.E.R.E.‡

Summary: This paper describes a study of the use of a single ring array of monopoles for a wideband multiple-beam aerial system for communications. Directional patterns are calculated for 24- and 32-element circular arrays over the frequency band 1·5–10 MHz and for different types of array excitation (or element combination in the case of a receiving array). The calculations are backed by experimental measurements on a model circular array operating over the band 200–1365 MHz. The study shows that low side-lobe levels (−6 to −9 dB) and acceptably small variations of beam-width, can be achieved over the above frequency band.

1. Introduction

Point-to-point h.f. and v.h.f. communication links generally use rhombic aerials for transmission and reception. A single rhombic aerial can only give directivity in two directions and therefore a large communication centre may require extensive aerial 'farms' in order to obtain good azimuth and frequency coverage. This is very expensive in terms of land acreage and hardware. A single aerial system which could replace a number of rhombics would have obvious advantages.

One approach to achieve such an aerial system is the Wullenweber array[1] which is a circular array of radiating or receiving elements backed by a reflecting screen within the active array. These arrays usually form one beam from a group of elements in the form of an arc. This beam is then rotated or repositioned using a capacitance-coupled commutator (or goniometer). A disadvantage of this type of system is the cost of the reflecting screen which should be taller than one quarter-wavelength at the lowest frequency of operation.

A simpler approach to the problem is to employ wideband elements and to form beams with different pointing directions by taking groups of elements in the form of arcs of the circle.[7] The width of these beams will necessarily vary with frequency and this is the principal factor limiting the bandwidth of operation. At the high-frequency limit the beams will not overlap and thus will give gaps in cover; at the low-frequency end there will be excessive overlap of beamshape and if the arc is small compared to λ, the front-to-back ratio will be very poor.

† Formerly with the Department of Electronic and Electrical Engineering, University of Birmingham; now with The General Electric Co. Ltd.

‡ Formerly with the Department of Electronic and Electrical Engineering, University of Birmingham; now with British Railways Board, Electrical Research Division, Derby.

This paper studies simple methods for exciting (or combining) the elements of a single ring of monopoles in order to retain good directional patterns over a wide frequency band. The type of specification that would be required, together with some of the compromises involved are listed below:

(i) Full 360° coverage in bearing should be provided by the beams. Thus the beam-width should not be less than the angle between adjacent beams.

(ii) The number of beams should be neither too great, which would result in confusion due to excessive beam overlap, nor too small which would necessitate wide beams and low directivity. A suitable compromise is considered to be a system with 24 beams.

(iii) Each beam should be capable of receiving signals reflected from the ionosphere arriving with angles up to 50° from the horizontal. Thus the beam should be 'fan' shaped in the vertical plane with no zeros for at least 50°, a 3dB beamwidth of about 40° being sufficient to fulfil this condition.

(iv) Side-lobes should be as low as possible over the whole band to give the best rejection of unwanted signals. Typical values of −6dB to −10dB over the band would be acceptable.

(v) The system should be simple and as inexpensive as possible; thus simple elements with no reflecting screen should be used.

(vi) The operational frequency range should be about 7:1, covering about 1·5 MHz to 10 MHz.

It is expected that the aerial system discussed will only be operated as a receiving aerial unless broadband impedance-matched array elements are used. With a multiple beam system, impedance matching of the transmitter into the aerial will cause multiple signal reflections between elements and the tuning circuit, which will in turn disperse through all the beam

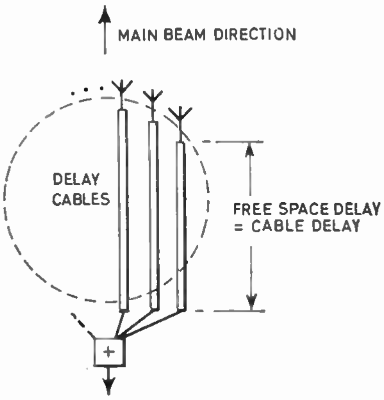

Fig. 1. Co-phasal circular array.

networks and be re-radiated. If the transmitter is not matched, the loss of signal power through incorrect matching will in general be greater than the gain due to the directivity of the array. On reception, impedance matching is not so important. Medium noise is usually much greater than receiver noise in the h.f. band, therefore directivity is more important than reducing signal loss in beam forming networks by matching.

2. Beam Formation

A multiple beam system can be constructed by dividing the signal output of each element a number of ways. By weighting and adding one set of the sub-divided outputs from each element a single beam can be formed. This can be repeated, using the other sub-divided outputs, to form other beams. If the signals from the elements are divided by using lossless hybrid networks and if all the beam forming and adding circuits are matched, then the amplitude of a sub-divided element output will be a constant fractional part of the actual signal received by the element and will not depend on the beam forming networks. It is therefore possible, with such a system, to design one beam without consideration to the others.

3. An Analytic Approach

The synthesis of a directional pattern for a circular array in one plane and at one frequency is quite a simple matter using harmonic analysis techniques.[2] This work has been extended to find the element weightings necessary to produce a good fan beam at a given frequency.[3] However, when the required element feeds are computed, using this technique, it is found that complicated feed circuits are required to produce the correct variations, with frequency, of the amplitude and phase weightings of the signals from the array elements. A system designed on this basis was therefore considered too expensive and complicated. This paper discusses a more heuristic approach to the design of wideband circular arrays with fan-shaped patterns, whereby the best possible pattern is obtained using very simple feed systems.

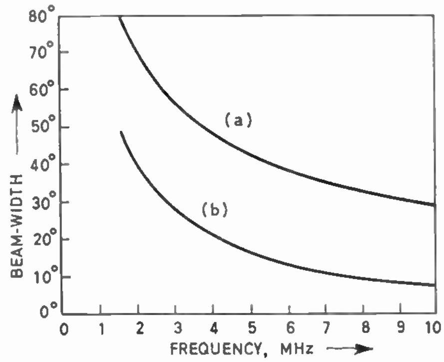

Fig. 2. Beam-widths of a 150 m (500 ft) diameter co-phasal array.
Curve (a): vertical; peak to null-angle.
Curve (b): horizontal; null to null-angle.

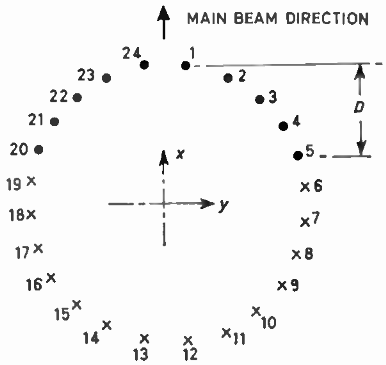

Fig. 3. Sector excited array.
● elements used; × elements unused.

3.1. *A Practical Approach to the Problem of Producing a Wideband Fan Beam*

The most obvious way of producing a wideband array is to use delay lines in the element feed cables so that the signals from the elements are all brought into phase, prior to adding, when a signal arrives from the desired main beam direction. If the delay time of the cable from a given element is made equal to the difference between the time of arrival of the signal at the element and the time of arrival at some arbitrary plane behind the array, then all signals will be co-phasal, independently of frequency, when added (see Fig. 1).

If all the elements are phased in this way prior to adding, then the well-known co-phasal pattern for circular arrays will be produced. It has been shown elsewhere[4] that if the array is co-phasal for a signal arriving from a direction $\theta = 0$ and $\phi = 90°$, then the pattern in the plane of the array is

$$D(\theta)_{\phi=\pi/2} = J_0\left(\frac{2\pi r}{\lambda}\cos\frac{\theta}{2}\right) \qquad ......(1)$$

and the vertical pattern in the main beam direction is

$$D(\phi)_{\theta=0} = D_e(\phi)\, J_0\left[\frac{2\pi r}{\lambda}(1-\sin\phi)\right] \qquad ......(2)$$

where D_e (ϕ) is the vertical pattern of a single element in the array and r is the radius of the array. The first zero of $J_0(x)$ occurs when $x = 2{\cdot}40$, so by equating the arguments of the Bessel coefficient in the previous two equations to 2·40, the null beamwidths can be computed for the main beam in the horizontal and vertical planes. These beamwidths are plotted in Fig. 2 for a 150 metre (500 ft) diameter array over the frequency range 1·5 MHz to 10 MHz. It can be seen that below about 4 MHz the beam in the vertical plane has the required vertical beamwidth. Above this frequency it would be desirable to modify the excitation to improve the vertical pattern. Since it is the front-to-back dimension of the array in the main beam direction (i.e. the x-direction in Fig. 3) that determines the vertical beamwidth, it would be expected that reducing this dimension by switching off elements at the back arc of the array will increase the vertical beamwidth. The dimension D of the remaining arc in Fig. 3 which can be expected to produce the best vertical pattern will be about 1 to 2 wavelengths. If D is made less than this the active part of the array will approach a linear array with a front-to-back ratio of unity. Consequently, care must be taken in selecting the elements to be switched out.

A simple method of implementing this technique would be to divide equally the signals from each element a number of ways, then by selecting the correct number of elements and adding, an output suitable for a specific band of frequencies would be obtained. With a suitable number of such adding circuits as shown in Fig. 4 the whole band can be covered. This multiple-channel system may be quite adequate for many purposes, but in some circumstances switching to different channels may be inconvenient; also the insertion loss due to dividing the element signal into different channels may make the system impracticable.

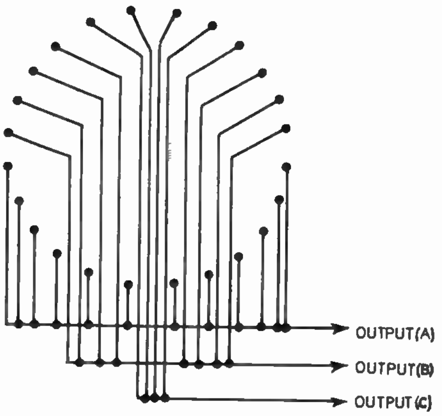

Fig. 4 Possible configuration for a multiple-band circular array.

Low-frequency output (A) + (B) + (C)
Mid-band output (A) + (B)
High-frequency output (C)

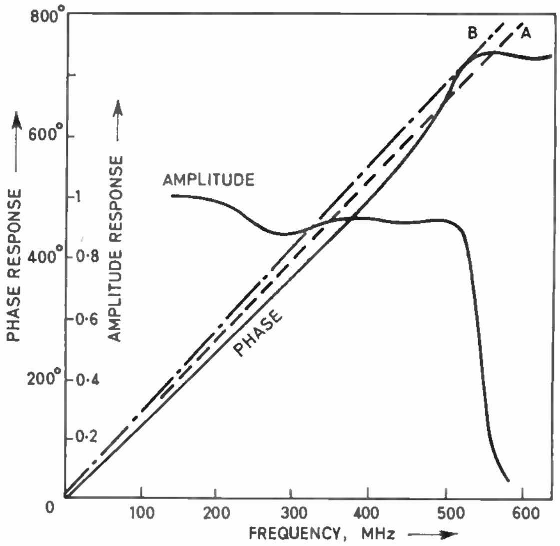

Fig. 5. Frequency response of a low-pass filter.

A possible way round this difficulty is to use low-pass filters effectively to taper out certain elements at a specified cut-off frequency. If the outputs of certain elements are grouped and passed through a low-pass filter prior to final summation, the contribution from these elements will be reduced for frequencies below that of the filter cut-off. Hence the effective arc of the circle in use at any given frequency, may be made frequency dependent. This possibility is investigated further. The main difficulty with this technique is compensating for the phase response of the filter in the pass-band. Figure 5 shows the measured phase response of a typical filter, and the line OA is the phase response of a pure time delay giving the best fit to the filter phase response. The maximum deviation between the two phase responses is 35°, which means that if the signals from the unfiltered elements in an array are delayed by the correct amount there will be, at worst, 35° difference between these signals and the signals from the filtered elements. If half the array is to be switched off with such a system the main beam will be reduced due to the two signals not adding in phase by a factor cos (35°/2), i.e. reduced to 96·5%. A more detailed discussion of the effect of this phase error is made later in conjunction with the results from the experimental model system.

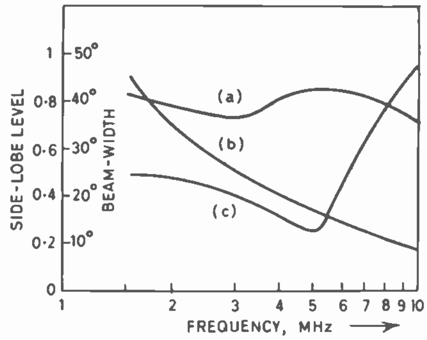

Fig. 6. Computed performance of 150 m (500 ft) diameter array with low-pass filter.

Curve (a): vertical beam-width, peak to 3dB-point.
Curve (b): horizontal beam-width, peak to 3dB-point.
Curve (c): worst side-lobe level, voltage ratio.

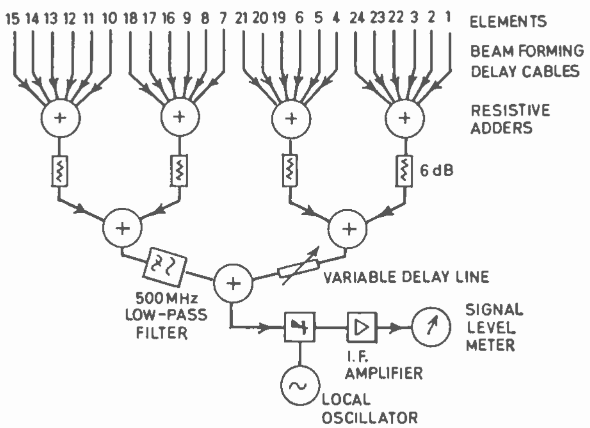

Fig. 7. Schematic diagram of complete system.

No simple theoretical approach has produced a solution to the problem of determining how many elements in the array should be switched off by the low-pass filter. A computer program was therefore devised which calculated the directional patterns of an array with progressively larger sectors of the array filtered out. These computations neglected the effect of coupling between elements but took the phase response of the filter into account.

Patterns were computed for the 24-element 150 m (500 ft) array mentioned earlier for frequencies of 1·5, 3, 5, 7 and 10 MHz. The number of elements connected to the filter was progressively changed as also was the co-phasal elevation angle. The best patterns were obtained when the array employed 12 elements feeding the low-pass filter and when the element-signals were time delayed to make them co-phasal at addition for signals arriving from 20° above the horizon ($\phi = 70°$). The computed patterns for this configuration are summarized in Fig. 6 and the experimental configurations shown in Fig. 7. The high side-lobe levels above 7 MHz are due to the wide element spacings; similar computations for a 32-element array showed that marked improvements could be expected.

4. A Scale Model Array

The particular array configuration suggested by these computations was further investigated using an experimental scale-model array. The advantage of using a scale model is that a small array can be mounted on a turn-table making directional pattern measurements much easier. Also the small size makes any necessary modifications, say, to the radiating elements, trivial compared with the problem of altering towers which are in the region of 50 ft in height.

4.1. *Mechanical Details of the Model Array*

The size of the model was determined by practical mechanical considerations and the most convenient diameter was approximately 1 m (3 ft 8 inches). This gave a scale factor of 136·5 : 1 and a required frequency range of 200 MHz to 1365 MHz. A ground plane was simulated by an octagonal sheet of brass and wire mesh, which was mounted on a caravan roof. The size of the ground plane was approximately 5 metres across. The ground plane was not rotatable but the array was constructed on a central circular brass plate which could be rotated. The array could be hand-rotated from within the caravan by a shaft going through the roof, the element feed cables being brought into the caravan through the centre of the shaft.

Elevated feed monopoles would be a typical choice for the array elements in the h.f. band because they are simple and inexpensive and have been shown to have less variation of impedance with frequency than base-fed monpoles.[5] Also, the radiation resistance is higher at low frequencies.[6] This means that losses in a resistive ground plane are not very important. Another advantage is that the feed point does not occur near a current minimum at high frequencies and the feed current is therefore a better estimate of the average element current. The reasonable practical approximation to an elevated feed monopole at u.h.f. is a sleeve monopole. This is a coaxially-fed monopole with the feed-cable extended up through the ground plane to the feed-point. The centre conductor of the cable becomes the monopole above this feed-point; and below the feed-point the outer conductor forms the element.

The element used had a total length of 9 cm (3·7 in) and a sleeve length of approximately 2·4 cm (0·93 in) with a 50 Ω coaxial feed. The most important difference between this element and the elevated feed monopole was the feed-point configuration, causing the terminal zone capacitance of the two elements to be different. It can be expected then, that the overall performance of the two elements will be similar, with the exception that their radiation impedances will be shunted by different feed-point capacitances.

4.2. *The Beam Forming System*

Only one beam network of the multiple-beam system was constructed. This was formed by delaying the signals from each element with low-loss double screened 50 Ω coaxial cable. The cable lengths were determined to form a co-phasal array with 20° beam elevation. The cables were cut until their measured electrical length, with plugs, was within 3 mm of the required length.

Ideally, the delayed voltages should be added in pairs in lossless hybrid networks. With such a circuit both inputs are isolated from each other and the characteristic impedance is 50 Ω. At the time of starting this work no transformers were available for producing such a network for use at 1·3 GHz and even now they are rather costly when large numbers are required. However, a useful alternative for these high frequencies was found to be resistive adder networks. The input impedance to any port of these devices can be made 50 Ω with all other ports terminated. A signal going into any port is divided equally, with some loss, between the other ports. A seven-port six-way signal adder which was used had a 15·6 dB insertion loss between one input and each of the others. This loss is not important in an experimental system where sufficient received power can be made available; the 15·6 dB isolation obtained between inputs however is very useful. The signal adders used were found to have a maximum input v.s.w.r. of 1·2 and the insertion loss was found to vary by ±1 dB.

The element signals were added in groups of six and the outputs from the 6-way adders were then combined in 2-way adders. The low-pass filter was connected into the appropriate branch of the combining tree which went to the back half of the array. An adjustable constant-impedance line was put in the other branch to compensate for the effective length of the filter. The filter used was a variable coaxial-line type with a performance as shown in Fig. 5. A diagram of the complete system is shown in Fig. 7.

4.3. *Directional Pattern Measurements*

A transmitting system was set up at a point 60 m (200 ft) away from the circular array which is well into the far field. The phase error across the array caused by a wave front being spherical instead of plane is of the order of ±3°.

A log-periodic dipole array was constructed for use as the transmitter aerial. This was found to have an approximate beamwidth of 60° over a frequency range of 200 to 800 MHz. Above this frequency a dipole and corner reflector gave a suitable performance. An estimate of the severity of site effects was made by rotating a single isolated monopole around the circle where the array was to be placed. This was achieved by removing all but one of the array elements. Signal strengths should have been uniform but fluctuations of the order of ±1 dB were noted at 200 MHz increasing to ±4 dB at 1300 MHz. This indicated that only approximate directional patterns would be measured at the highest frequencies. It is also possible that some of these fluctuations were due to the non-circular ground plane.

The circular array and the transforming equipment was set up with the adjustable line set at 106 cm to compensate for the average phase response of the filter (the line OA in Fig. 5). Directional patterns were measured in the horizontal plane with this system and the beamwidths and maximum side-lobe levels obtained are shown in Fig. 8.

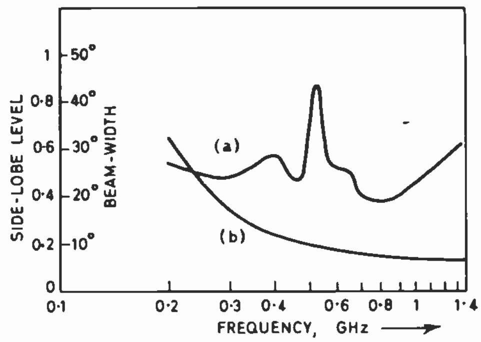

Fig. 8. Measured performance of scale model array with 106 cm delay line.

Curve (a): worst side-lobe level.

Curve (b): beam-width, peak to 3dB-point.

Over the frequency range 500 MHz to 525 MHz poor patterns were obtained with high side-lobes. The reason for this becomes clear if the effect of phase errors between the filter and the compensating delay line are investigated. The pattern in the vertical plane at $\theta = 0$ can be approximately considered as a combination of the patterns of the front and back halves of the array taken separately multiplied by an interferometer term. This term will be the pattern of two omnidirectional elements placed at the phase centres of the two parts of the array. At 500 MHz the spacing between these elements (approximately half the array diameter) will be about one wavelength. The interferometer pattern produced $I(\phi)$ will be

$$I(\phi) = \cos[\pi \cos(\phi - \Phi)] \qquad \ldots\ldots(3)$$

where Φ is the phase shift applied to one signal before adding. In this system the required value of Φ would cause the maximum lobe in this pattern to be in the co-phasal direction of the array ($\phi + 20°$); deviations from this will cause the interferometer term to scan up or down.

An improvement to the horizontal pattern can be made by increasing the length of the compensating delay line. If the line length is chosen as 112 cm to correspond to the line OB in Fig. 5, then all errors will deflect the interferometer term downwards. Figure 9 shows the measured side-lobe levels and beam-widths for this length of delay line and a good improvement is evident. However, the vertical beam with this configuration can be expected to be narrower. Unfortunately pattern measurements required to obtain vertical beam-widths were not possible due to practical difficulties on site. Figure 10 shows three typical experimental patterns in the horizontal plane.

At frequencies above 560 MHz no measurable differences were obtained with different compensating delay lines. From this frequency up to 1000 MHz the patterns obtained were quite good, the measured side-lobe level being less than 0·43 of the main beam amplitude. In the absence of site effects these side-lobe levels would probably be lower. The site effects cause considerable masking of details of the directional patterns above 700 MHz. Above 1000 MHz grating lobes occurred but their maximum amplitude remained below 0·6; the position of the lobes remained fairly constant at $\theta = \pm 140°$ and $\theta = 180°$. Below 560 MHz the directional pattern had side-lobes less than 0·5 for the array with the longer compensating delay. Owing to practical considerations experimental patterns are restricted to the horizontal plane.

4.4. *A Discussion of the Effects of Mutual Coupling*

Comparing the experimental and theoretical directional patterns it was found that moderate agreement was obtained at the lower frequencies: the differences were a narrowing of the main beam and an increase in the first side-lobe level of the measured pattern. The main reason for the degree of agreement obtained is that when the monopole is well below its resonant frequency the self-impedance is much higher than the mutual impedances which can therefore be neglected to a first approximation when computing directional patterns.

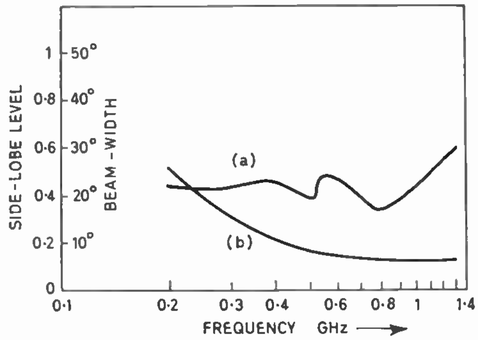

Fig. 9. Measured performance of scale model array with 112 cm delay line.
Curve (a): worst side-lobe level.
Curve (b): beam-width, peak to 3dB-point.

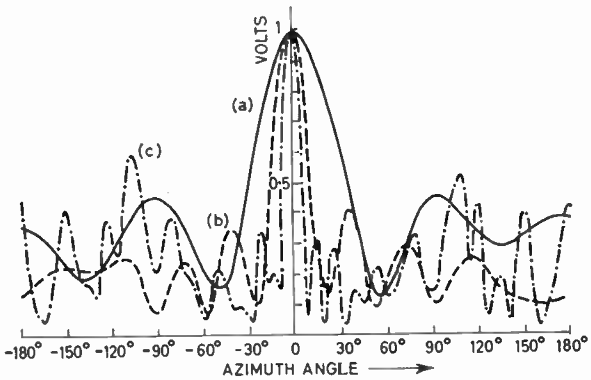

Fig. 10. Some measured directional patterns in the horizontal plane with a 112 cm compensation delay line.
Curve (a): 205 MHz; Curve (b): 480 MHz;
Curve (c): 1350 MHz.

The results at higher frequencies showed that mutual coupling certainly affected the directional pattern inasmuch as the side-lobes were altered but there was no destruction of the main lobe and no strong re-inforcement of any particular side-lobe. In fact at the highest frequencies the amplitudes of the diffraction grating lobes were lower than expected, showing that mutual coupling has improved the performance of the array at these frequencies (there is nothing fundamentally unreasonable about this).

5. Conclusions

A study has been described of the problems involved in designing a communication aerial in the form of a circular array with a simple element feed system which is able to provide a good directional pattern over a wide frequency band. The method is not analytic but relies partly on a selection of the best computed pattern from those due to various array configurations but based on a beam co-phasal system. Pattern computations were made for a 24-element 500 ft diameter array and the best performance over the required band was obtained using the whole array from 1·5 to 4 MHz and the front half only from 4 to 10 MHz.

A model of this system operating at u.h.f. has been built to investigate the feasibility of using a low-pass filter effectively to 'switch out' the back half of the array above the cut-off frequency of the filter. Using a simple delay line to compensate for the phase response of the filter it is found that only a small amount of pattern degradation occurs near the cut-off frequency of the filter. Measured side-lobes of between −9 dB and −6 dB are obtained over a substantial portion of the band and the maximum 3 dB beam-width is 50°. This type of performance would be very acceptable for a wideband array in the h.f. band.

Computation [3] showed that a further improvement in performance would be obtained at the high frequency end of the band by using a 32-element array and two low-pass filters to 'switch off' progressively larger sections of the array as the frequency is increased. However increased difficulty would be encountered in compensating for the phase response of two filters.

6. Acknowledgments

The authors wish to acknowledge helpful discussions with their colleagues in the Department of Electronic and Electrical Engineering, University of Birmingham, and Dr. P. J. D. Gething of the Government Communications Headquarters, Cheltenham. The work was supported by a Ministry of Technology Research Agreement.

I. D. Longstaff is also indebted to the General Electric Company Ltd., for the provision of a scholarship.

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INSTITUTION NOTICES

Electronics Design Conference

The conference on Electronics Design, to be held at the University of Cambridge from 23rd to 27th September 1968, will pay particular attention to the estimating of development cost and the control of design projects. Questionnaires sent to engineers known to be interested in design showed these to be of prime concern. Other topics which were asked for in this Conference were the formulation of product policy, market research techniques, the formation of electronics design teams, value engineering, computer-aided electronics design and decision-making in the design process. The analysis also shows that whilst junior designers are more interested in design method and research into design processes than their seniors, they are much less interested in human factors and the behaviour of design teams.

The conference is being organized by the I.E.E. Electronics Division, the I.E.R.E. and the I.E.E.E. (United Kingdom and Republic of Ireland Section), with the collaboration of the British Institute of Management and the British Association for Commercial and Industrial Education. I.E.R.E. representation on the Joint Organizing Committee, which is under the Chairmanship of Mr. H. V. Beck, are Dr. G. L. Hamburger (Fellow), Mr. J. D. Tucker (Member) and Mr. D. O'N. Waddington (Member).

Further details are available from the Conference Department, Institution of Electrical Engineers, Savoy Place, London, W.C.2.

Switching Techniques for Telecommunications

The Institution of Electrical Engineers, in collaboration with the Institution of Electronic and Radio Engineers and the Institute of Electrical and Electronics Engineers (United Kingdom and Republic of Ireland Section), is arranging a conference on Switching Techniques for Telecommunication Networks, to be held at Savoy Place, London, from 21st to 25th April 1969.

The Organizing Committee, on which the I.E.R.E. is represented by Mr. J. Powell (Fellow), invites contributions of about 1500 words for inclusion in the conference programme. Intending authors are asked to submit synopses of approximately 250 words to the I.E.E. Conference Department without delay. Manuscripts will be required for consideration by 24th December 1968.

The scope of the conference will include:

Switching and control techniques and systems (excluding telegraph and data store and forward techniques);

Network control and management;

Choice of techniques for future systems, and planning development to meet needs.

Contributions are also invited on sub-systems, techniques and the application of devices; hardware and software aspects of control systems; and traffic studies and signalling.

Registration forms and further details of the conference will be available from the Institutions in due course.

The Institution and N.E.C.

The Institution has recently undertaken responsibility for publishing the *N.E.C. Review*, the quarterly journal of the National Electronics Council. Founded in 1964 as the National Electronics Research Council, the N.E.C. is now administered by the Ministry of Technology. Its General Committee, which is under the chairmanship of Admiral of the Fleet the Earl Mountbatten of Burma, comprises nominees of government departments, the universities, learned societies, and industry.

In the latest issue of *N.E.C. Review*, dated April 1968, the contents included a contribution from the Ministry of Technology on Electronic Control in Industry, and two reports in the series on Electronics Research in British Commonwealth Universities, namely from the Departments of Electrical Engineering of the Faculty of Science, University of Manchester, and of the University of Salford.

Requests for information on the submission of contributions and on subscriptions to the *Review* should continue to be sent to The Secretary, National Electronics Council, Abell House, John Islip Street, London, S.W.1.

I.Mech.E. Charter and Bye-Law Changes

The Institution of Mechanical Engineers has announced that the Privy Council has approved certain changes in the description of classes of membership.

In future Honorary Members and Members will be called Honorary Fellows (Hon. F.I. Mech.E.) and Fellows (F.I. Mech.E.); present Associate Members become Members (M.I.Mech.E.). Other descriptions of membership classes remain unchanged.

The I.Mech.E. is thus the third of the major chartered engineering institutions to implement this change toward eventual uniformity in designations. The I.E.E. and I.E.R.E. have already made the necessary alterations.

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Television Network Switching at the Post Office Tower in London

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Presented at a meeting of the Television Group in London on 24th May 1967.

Summary: Television network switching equipment is concerned principally with setting up vision and sound routes between various sources and destinations. Due to the growth of the network system this equipment has developed from the simplicity of patch-panels to a complexity requiring a certain amount of automation.

In the equipment recently installed at the Post Office Tower in London and also in three provincial centres, provision is made for storing two patterns of connections and their switching times. Pre-setting of information is carried out manually but the actual switching operations are under the control of a clock system driven by pulses derived from the speaking-clock TIM. Provision is made for overriding the clock or changing the stored information when last-minute schedule changes occur. The equipment is designed for 625-line colour operation.

1. Introduction

The Post Office provides sound and vision channels to meet the requirements of the B.B.C. and I.T.A. broadcasting services and of many other organizations which use closed-circuit television networks on either a permanent or temporary basis. The 400 vision channels in use at present total approximately 9000 channel miles. These channels are provided by microwave radio links, carrier systems on coaxial cable and by direct video transmission on coaxial cables.

A typical broadcast authority's network, shown in Fig. 1, consists of long inter-city links and shorter links to transmitters, studios, recording centres, monitoring centres, outside broadcast reception points and to the broadcast authority's control point known as a programme switching centre (p.s.c.). In general, all these various links terminate at a Post Office repeater station known as a network switching centre (n.s.c.).

The B.B.C. television services are essentially national and for the majority of time the transmitters radiate centrally produced programmes. In consequence, the B.B.C. vision and sound links are permanently connected through at the network switching centres and switching operations are carried out by the B.B.C.

The organization of the Independent Television service is such that it necessitates considerable flexibility and frequent changes of network configuration. The division of responsibility between the I.T.A. and the programme contractors results in the main inter-city links and transmitter links being rented by the I.T.A. while studio and other local links are

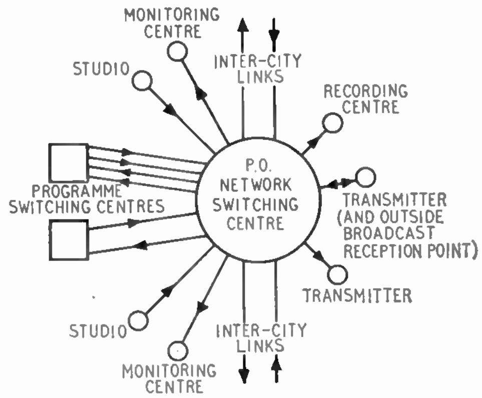

Fig. 1. Typical arrangement of switching centres.

rented by the individual contractors. Programme contractors are free to distribute and interchange programmes and different contractors may operate at various parts of the week. Very often the same recorded programme may be radiated by several transmitters at different times of the day. From the very beginning it was realized that the most economical and flexible way of providing this routing of channels was to locate the switching equipment at the network switching centres. In this way a substantial saving in route mileage can be effected.[1]

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‡ Post Office Engineering Department, Gresham Street, London, E.C.2.

§ Formerly with The Marconi Company Ltd.; now with the Radio and Electronics Section, Greater London Council.

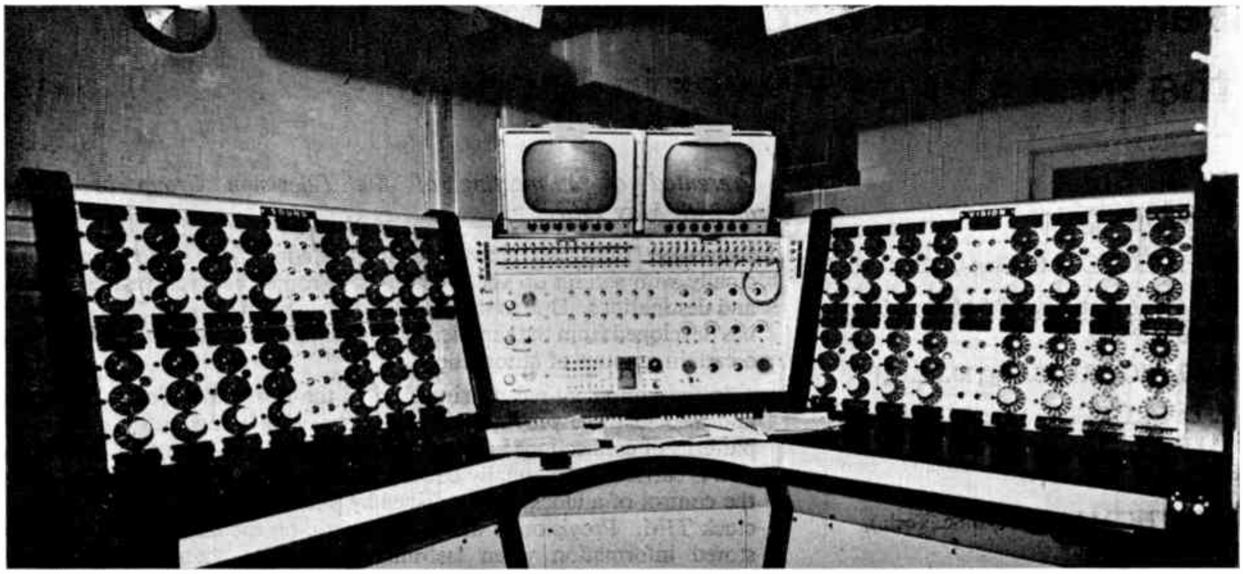

Fig. 2. Old-type control console at the London network switching centre.

2. Present Network Switching Equipment

Commercial television was inaugurated in September 1955 and since May 1956 network switching has been carried out at network switching centres. The inter-connection patterns and switching times required are determined by I.T.A. for inter-city and transmitter links and by programme contractors for local channels only.

It was realized at the very outset that it was essential that all programme sources and switching points should operate to the same standard time and it was agreed that the Post Office speaking-clock TIM would be used because it is accurate and readily available over the whole country.† Changes to the pattern of connections are very rarely made 'on-air' and it is normal to have a period of at least several seconds between the end of use of one connection and the beginning of use of the next one. The times of changes are given in multiples of ten seconds, e.g. 19 hours 32 minutes 10 seconds, corresponding with the ten-second announcements from TIM.

The number of channels involved and the number of switching operations per day have been increasing due to the growth in number of transmitters, network programmes and the hours of broadcasting. The switching equipment, to keep pace with this increase, has also evolved from the simplicity of patch-panels and rotary switches to an advanced state of semi-automated operation controlled by an electronic clock.

† TIM is the original dialling code for the speaking-clock and although superseded by 123 for London and other director exchange systems, it is used in this paper as a convenient designation.

The largest n.s.c. is in London and it handles about 50 changes of interconnection pattern per day. The switching equipment in use at present in London was installed in 1958 and provides for 16 inputs or sources switched to 15 outputs or destinations; separate switching for sound and vision is employed.

Figure 2 shows the old control console in London n.s.c. The sound selection panel is on the left, the vision panel on the right and in the middle portion are located the electronic clock with its controls, monitor selectors and two picture monitors. The switching relays are controlled by banks of 16-way, manually-operated rotary switches and each destination has three switches coloured red, green and blue. Thus at any time, if for example the red switches are operative, the positions of these red switches determine the pattern of actual connections, i.e. 'on-air' pattern. The next two patterns required are set-up on the green and blue switches respectively and the times at which these patterns are required entered on green and blue time entry controls. The switching sequence is fixed: red, green, blue, red, green, etc. Therefore, at coincidence between real time and the pre-set time on the green controls, the relays change automatically to the new pattern which becomes the 'on-air' pattern. The next required pattern should already have been set-up on the blue controls and the pattern after that and its switching time can now be entered into the red controls. Alterations to the scheduled patterns are occasionally made at short notice, e.g. part or the whole of a pattern of changes may be required earlier or later than was expected and the facility is provided for such late alterations to be made manually.

Network switching at centres outside London is on a smaller scale and at some centres very simple arrangements are used for connecting video channels together.

3. Requirements for the New Switching Equipment

Extensions of the Independent Television network have resulted in the present state of affairs where many network switching centres (in particular those in London, Birmingham, Manchester and Carlisle) are trying to cope with a load which exceeds their design capabilities. Additionally, the imminent introduction of colour television has made the need for the new equipment more urgent, as most of the existing switching equipment was designed for 405-lines monochrome operation. In 1962, the Engineering Department of the Post Office started planning the new equipment for London and for the three provincial centres, taking into account the experience gained during the operation of the existing network switching centres.

3.1. *Facilities*

The new switching equipment is required to cater for 30 sources and 40 destinations in London and 15 sources and 15 destinations in the provincial centres. This capacity will be sufficient to meet the requirements for the near future, but the equipment must be capable of expansion to an ultimate switching capacity of 40 sources and 80 destinations in London and 25 sources and 30 destinations in the provinces.

The experience with the existing equipment has shown that considerable simplifications in operational procedures are possible, making it quicker to set up a pattern of connections and easier to check that the correct entries have in fact been programmed. There is no need for separate sound and vision patterns as they are always 'married'. Therefore, only one set of controls is necessary. The present equipment requires complete patterns of connections rather than merely requiring the changes to patterns to be pre-set. The need to enter so much redundant information in cases where the change affects only a few routes increases the probability of operating error. Although the existing fault rate is very low, it might be difficult to maintain it on a larger equipment if the same principle had been followed.

3.2. *Performance Requirements*

The problem of deciding on the performance required of the various equipments between the origin of a television picture (e.g. studio camera) and the broadcast transmitter presented many difficulties. The solution was obtained by postulating a hypothetical longest chain (United Kingdom reference chain), determining the performance requirements for this chain by subjective impairment tests of television pictures, and then allocating a proportion of the total permissible error to each constituent part of this chain according to its complexity. To obtain practical performance specifications it was necessary to determine the law of addition for each type of distortion, i.e. crosstalk, differential phase and gain, linear waveform distortion (K rating), etc. The cumulative distortion of cascaded equipment is the result of the addition of individual distortions according to laws which in general may vary between linear addition and root sum square addition depending on the design of equipment and the parameter under consideration. For planning purposes, however, it was possible to derive empirically the approximate laws of addition of individual parameters and thus establish performance requirements for each type of link in a chain.

The hypothetical United Kingdom reference chain consists of four 'main' links totalling 500 miles, two 'major local' links and six 'minor local' links. Both types of local links have route lengths less than 25 miles, but 'major local' links have intermediate repeater or relay points, while 'minor' links have not. Very few broadcasts take place over a longer chain than this and most of course are over a much simpler chain. All the essential parameters, e.g. linear distortions, non-linear distortions and noise, are defined for each of the three types of links.[2, 3]

It was decided that the new switching equipments should have a performance no worse than that required of a 'minor local' link. The output of the Carlisle switcher will often have come via all four switchers so that Post Office switching alone takes up the equivalent of four 'minor local' links. However, the performance of a 'minor local' link is very stringent and to obtain a performance better than this would be expensive and indeed may be impossible for some parameters.

There are two problems which are, perhaps, peculiar to switching and distributing devices. The first of these problems is the possible level variation on any one destination as other destinations are switched to the same source. The variation is required to be no more than that allowed on a 'minor local' link, namely, $\pm 0{\cdot}1$ dB for vision and $\pm 0{\cdot}25$ dB for sound. The second problem is that of crosstalk, particularly at colour sub-carrier frequency. All switching devices, when open, give crosstalk, the level of which increases with frequency. The number of crosstalking paths in parallel in a distributing switcher depends on the pattern of connections at any given time and the magnitude of the total crosstalk depends on how the various crosstalk signals add and whether the various paths have the same interfering signal or different signals. The permissible interference at colour sub-carrier frequency is not known with precision but is about 60 dB below picture amplitude, and this figure should be met if and when the ultimate 40×80 switcher is completed.

4. New Switching Equipment

The new switching equipments designed to meet the G.P.O. specifications have been recently installed at the G.P.O. Tower in London and at the three provincial switching centres in Birmingham, Manchester and Carlisle. The switchers in all centres are identical in concept and construction, the only difference being in size.

'Married' switching of vision and sound circuits is adopted, and full flexibility is provided. This means that any source can be connected to any number of destinations. Relay matrices are used for vision switching, and high-speed motor uniselectors are employed to control vision matrices and to switch programme sound and displays which show the existing pattern of connections.

4.1. *General Description*

Figure 3 shows the block diagram of the equipment. The actual 'on-air' connections between sources and destinations are made in the vision and sound switcher. Two further patterns of connections together with their

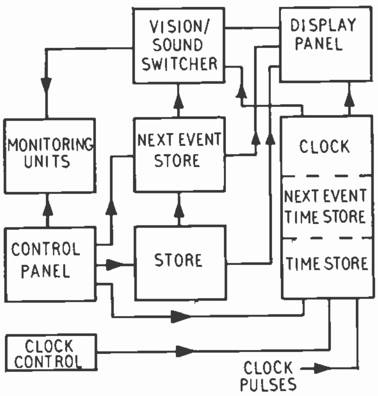

Fig. 3. Block diagram of the network switching equipment.

scheduled switching times can be pre-set and stored. Only the changes to the existing pattern of connections need be entered into the stores. The information for the first and second changes of connections are held in the 'next event store' and 'store' respectively. The clock system has also two corresponding stages of storage. The actual switching is under control of a 24-hour clock driven by pulses synchronized to TIM, the speaking clock. When the time pre-set in the 'next event time store' coincides with the 24-hour clock the switching pattern pre-set in the 'next event store' initiates the required changes of connections in the vision and sound switcher. The information in the stores, both switching and time, is then automatically transferred into the 'next event stores' leaving the stores vacant and ready to receive the next pattern of changes.[4]

Alterations to the data held in the two levels of storage can be carried out at any time, but access to the vision and sound switcher is only possible via the 'next event store'. This was done to avoid accidental changes to the 'on-air' pattern during the operation of the control panel. Provision is made to override the clock control, allowing any number of connections to be changed before the scheduled switch time or to be held as long as required. When not in use for loading the stores, the control panel is used to select, by means of separate monitoring units, sources and destinations for monitoring purposes.

A large alpha-numeric display panel shows the real time, the existing pattern of connections, the two pre-set patterns and their corresponding switching times. A clock control panel allows adjustment of the 24-hour clock and provides facilities for checking this clock against TIM.

The whole equipment is driven from 50 V and 24 V normal telephone exchange batteries. Some of the vision and sound amplifiers require a 240 V a.c. supply but this is derived from the 50 V battery by means of a static inverter. Should the inverter fail, the station a.c. mains supply is automatically connected in place of the inverter output.

4.2. *Equipment Location*

The television switching equipment in a Post Office network switching centre is located in two areas: television control room and the equipment room. A typical arrangement is shown in Fig. 4. The main control panel, clock control panel and two picture monitors are mounted in a control desk in the control room with the wall mounted display panel in front. The rest of the equipment is housed in racks in the equipment room. Figure 5 shows the new control position in the London n.s.c.

All the incoming and outgoing vision and sound lines appear at standard levels (1 V peak-to-peak and 0 dBm) on video and audio distribution racks (v.d.r. and a.d.r.) in the control room where the operator can carry out manual patching of lines in an emergency, on a one-to-one basis only. From the racks the lines are taken to audio and video distribution frames (a.d.f. and v.d.f.) which provide the necessary flexibility in case the allocation of sources and destinations on the switching equipment has to be changed. From the frames the lines are taken to the switching equipment racks where the vision lines are first waveform corrected for cable runs between v.d.r. and v.d.f. The

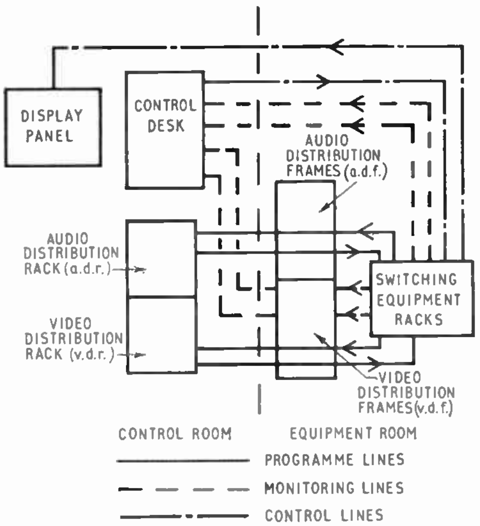

Fig. 4. Typical arrangement of television switching equipment in n.s.c.

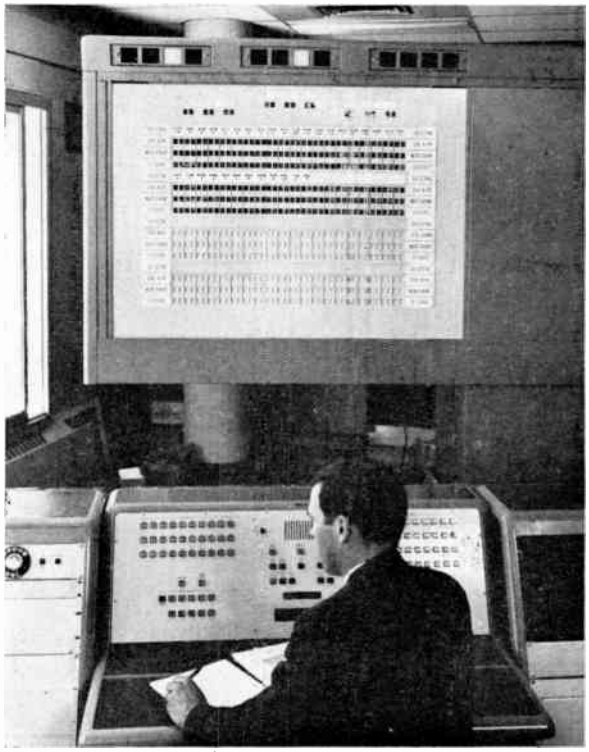

Fig. 5. New-type control console at the n.s.c. in London.

same route is taken by all lines to and from the switching equipment.

4.3. *Vision Circuits*

4.3.1. Design consideration

The design of vision circuits in particular had to be such that the performance specification would be met with the largest switching system (an ultimate capacity of the London switcher of 40 sources and 80 destinations) and yet the design had to provide an economical solution for the smallest system.

The decision was taken to adopt relay matrices for vision switching. At the beginning, a conventional design was contemplated; therefore, it is worthwhile considering the problems of a conventional matrix.

Figure 6(a) shows input and output bus-bars of such a matrix. Relays at each crosspoint between input and output buses are operated to provide the required connection pattern. To achieve the necessary isolation between input and output circuits considerable screening is necessary, and this in turn increases the stray capacitance to earth of the buses. Typically, a large matrix of this type could have stray capacitance values of several hundreds of picofarads on each bus. An isolation amplifier is necessary to deliver the input signal to this capacitance, and also an output amplifier with high input impedance is needed to drive the coaxial output circuit so as to prevent any problems of signal attenuation due to relay contact resistance.

It is common practice to ease the task of the input isolation amplifier by providing on each crosspoint a back loading capacitor C_L (Fig. 6(b)) which is equivalent in value to the total stray capacitance C_S of the output bus. The value of the capacitance which the input amplifier has to drive is therefore held reasonably constant. This fact makes a practical amplifier design possible even though it is still a difficult one.

Figure 6(c) indicates an additional problem, that of the coupling capacitance C_C which represents a crosstalk path between input and output circuits. This problem can be minimized by 'intermediate earth' wiring, a second relay contact 2 shunts the crosstalk signal to earth when the relay is energized (Fig. 6(d)). In this way, provided that there is sufficient isolation between relay contact sets, a reasonable crosstalk performance can be achieved. However, with so much stray capacitance and finite contact resistance of relays, good frequency response will create difficult problems. Previous experience has shown that in practice, to achieve an adequate performance, the largest matrix of this type should be limited to about 10 outputs and 20 inputs. To obtain a larger number of outputs, a number of matrices could be driven in parallel from input distribution amplifiers. To cater for larger numbers of inputs, auxiliary switching would have to be provided to transfer the output connections between two matrices. Thus for the London switcher the signal would have to pass through

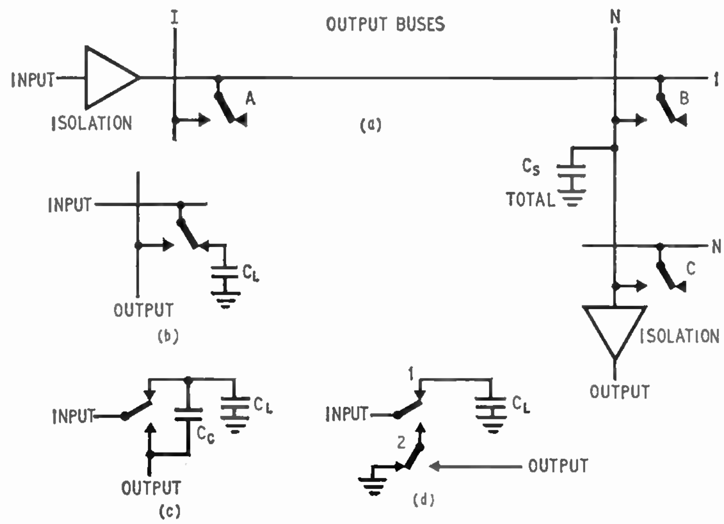

Fig. 6. Development of conventional relay matrix.

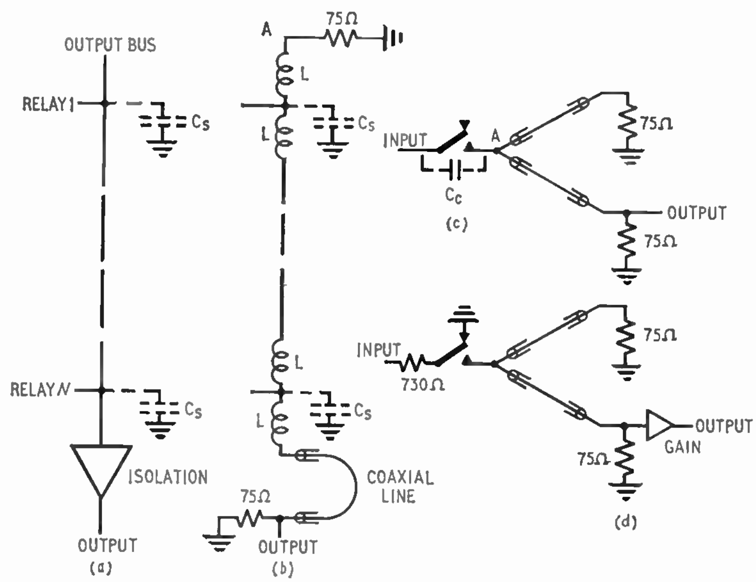

Fig. 7. Development of transmission line relay matrix.

two matrices with their associated amplifiers, thus suffering more degradation than is really necessary.

A new look at the problems of matrix design was called for, since to avoid switching the video signal twice a matrix must be able to accept up to 40 inputs. Figure 7(a) shows the conventional output bus-bar loaded at every crosspoint by capacitance C_S due to relay strays and the screening of the bus-bar itself. The impedance of the bus-bar is very low, because its loading is mainly capacitive and it is mismatched, if considered as a transmission line, when terminated with the high input impedance of the output amplifier. To avoid high-frequency response variation depending on which input is selected to it, the length of the output bus-bar must be restricted to no more than a small fraction of the wavelength of the highest frequency to be transmitted. A bus-bar length of 2 ft would produce a loss of more than 0·1 dB at 10 MHz. To cater for 40 sources the output bus-bar would have to be considerably larger than 2 ft if only to make the 40 source connections to it. The problem was overcome by designing the output bus-bar to have sufficient inductance which, coupled with the distributed capacitances C_S, formed a transmission line of 75 Ω impedance (Fig. 7(b)). An output could therefore be taken from the output bus-bar by means of a 75 Ω coaxial cable eliminating the isolation amplifier required previously. The other end of the bus-bar is terminated also in 75 Ω to complete the matching, thus ensuring a uniform frequency response along the bus-bar.

Figure 7(c) shows, in equivalent terms, the connection of an input signal to the output transmission line. A relay contact closes to feed the line at point A, the signal passes to each end of the line where it is absorbed in terminating resistors, one the load and the other a dummy load. The following two problems, however, exist at this stage: (i) the input impedance at point A is low, being formed by two 75 Ω in parallel, making it difficult to drive more than one circuit from one source: (ii) the capacitance C_C across the open relay contacts produces crosstalk. The first problem was overcome by building out the impedance with a 730 Ω resistor as shown in Fig. 7(d).

The resistor value was chosen such that 10 output circuits can be driven from one source and provide a termination of approximately 75 Ω. The resulting loss of signal amplitude can be easily restored by an output amplifier of sufficient gain. The problem of crosstalk was overcome by earthing the open contact of the relay, thus shunting the signal to earth when not used. The impedance of the input bus-bar is fairly constant: 73 Ω when no output is taken and 77 Ω when 10 outputs are taken, thus matching the input cable.

4.3.2. Practical implementation of vision switcher

The relay matrix is built up from relay strips with one input and 10 outputs. Ten relays with their associated feed resistors and surge suppression diodes are mounted on a printed board (Fig. 8). The top side of the board is nearly all copper to screen the relay coil connections from the video circuits. The video output lead from each relay is screened from its neighbour by small metal screening plates attached to the printed board.

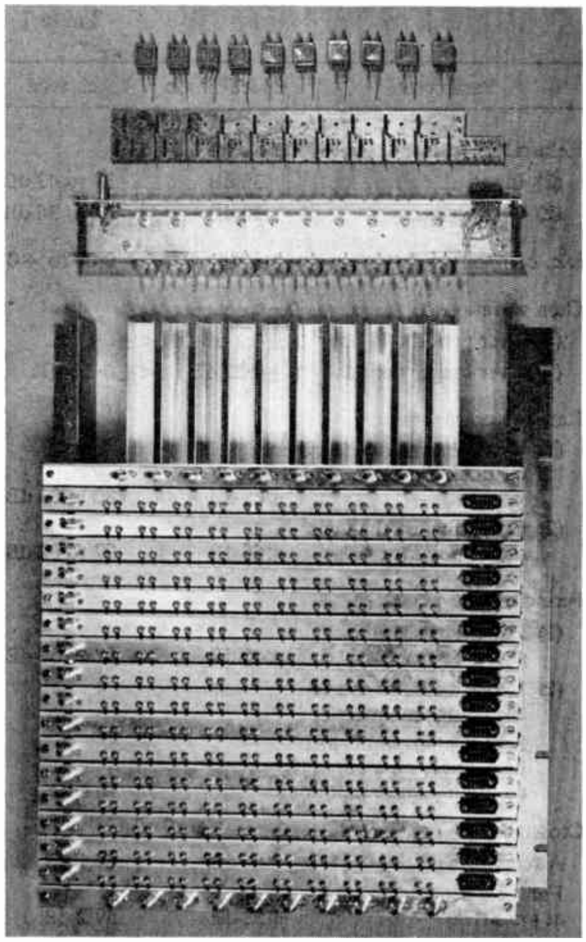

Fig. 8. Constructional details of relay matrix.

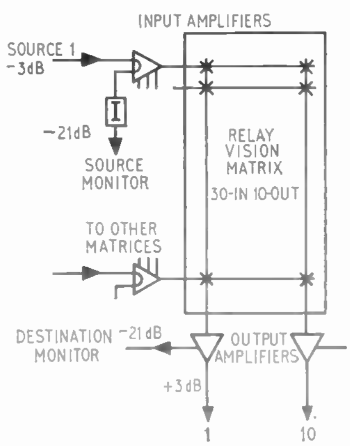

Fig. 9. Simplified schematic of vision circuits.

The matrix is formed by assembling a number of relay strips corresponding to the number of sources required on to a mounting frame. The inner conductors of the output transmission lines are formed by linking together the corresponding outputs of each strip by a tinned copper wire, the outer conductor being formed by a U-shaped channel which clips over the earthing spring. To ensure a good earth connection, and thus reduce the possibility of stray earth couplings which would introduce crosstalk, all matrix metalwork is made of silver-plated brass.

The simplified schematic of vision circuits is shown in Fig. 9. To eliminate gain changes due to slight variation of input impedance which in turn depends on how many outputs take the same input, relay strips are driven from amplifiers having one low-impedance output capable of driving four 75 Ω loads (four

Table 1 Video performance

Parameter	Specification	Achieved
Return loss (measured with 'T' pulse)		
(1) Input	30 dB	39 to 43 dB
(2) Output	30 dB	34 to 36 dB
D.c. component	$\ngtr$5 V	$-0·06$ to $+0·11$
Gain variations		
(1) Over period of 1 hour	$\ngtr \pm 0·1$ dB	$< \pm 0·1$ dB
(2) Over longer periods	$\ngtr \pm 0·25$ dB	$< \pm 0·25$ dB
Random noise		
(1) Monochrome and luminance channel, signal/weighted noise	$\nless$65 dB	80 to 83 dB
(2) Chrominance channel, signal/weighted noise	$\nless$65 dB	77 to 82 dB
Periodic noise		
(1) Power supply hum, signal/noise	$\nless$45 dB	52 to 57 dB
(2) Impulsive noise due to switching circuits. Specification limit: just perceptible on a correctly adjusted picture monitor		In spec.
Crosstalk		
Attenuation up to 1·8 MHz	$\nless$66 dB	67·1 dB
Falling 6 dB/octave; thus at 4·5 MHz	$\nless$58 dB	59·2 dB (worst cases)
Non-linearity distortion		
(1) Of sync. signal, when switching 10% to 90% a.p.l., at		
(a) Normal level	±1%	—
(b) 3 dB up	±2%	1%
(2) In the luminance channel use five-riser staircase, differentiating network and limiter		
(a) At normal level	1%	—
(b) 3 dB up	2%	<0·5%
(3) In the chrominance channel (differential phase and gain). At any a.p.l. between 10% and 90%		
(a) Phase, at normal level	±0·5°	0·06 to 0·4°
(b) Gain, at normal level	±1%	—
(c) Phase, 3 dB up	±1·0°	0·24 to 0·84°
(d) Gain, 3 dB up	±2%	0·2%
Linear waveform distortion (*K* rating)		
(1) With 625-line 2*T* pulse and bar		
(a) 1 path	$\ngtr$0·5%	0·25%
(b) 3 paths		<0·5%
Luminance-chrominance inequalities		
(a) Gain, normal level	±1%	0 to +1%
(b) Delay, normal level	±10 ns	<2 ns
(c) Gain, 3 dB up	—	+1 to +1·5%
(d) Delay, 3 dB up	—	2 ns
Long-time step response		
With 0·7 V step one path	$\ngtr$25%	4 to 7% 6 seconds to restore

Note: Normal level is + 3 dB rel. 1V.

matrices). The output bus-bars are fed into output amplifiers which restore the loss of gain produced in the matrix. The overall gain is 6 dB (input −3 dB and output +3 dB on 1 V composite level) to allow passive equalization of incoming and outgoing lines. Low-level feeds (−21 dB rel. 1 V) are provided from inputs and outputs of the switcher for source and destination monitoring.

4.3.3. Performance of vision circuits

Table 1 gives the summary of specification and actual performance of vision circuits for the London switcher. It can be seen that for most parameters the performance is well within the specified figures, the exceptions being crosstalk and differential phase on some routes which are just within specification. It is interesting to note that crosstalk performance is not decided by the matrix, which gives a figure of about 70 dB at 4·5 MHz for the worst combination of routing, but by the coupling between the frames of adjacent input amplifiers. Therefore the extension will not make the crosstalk any worse. In practice, the crosstalk performance will be much better as it is very unlikely that all input amplifiers giving the worst combinations for crosstalk will carry synchronous signals.

4.4. *Sound Circuits*

The programme sound is switched on motor uni-selectors, one uniselector per destination is used. The

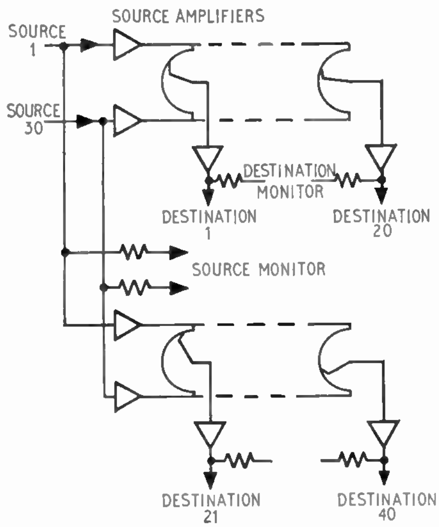

Fig. 10. Simplified schematic of sound circuits.

two levels carrying balanced sound circuits are gold plated to ensure low contact resistance. The simplified schematic of sound circuits is shown in Fig. 10. Each sound source is connected to two source amplifiers whose outputs are 'multipled' over the same position on 20 uniselectors. The output of each uniselector is fed to line via a destination amplifier. The output impedance of source amplifiers is low (about 35 Ω) and the input impedance of the destination amplifiers is high (about 50 kΩ), thus the output of any source amplifier can be connected to one or 20 destinations without any appreciable change in programme level. The source and destination monitoring feeds are supplied via hold-off resistors.

Table 2 gives the summary of performance in comparison with the specification. Crosstalk is within the specification—better than 80 dB at 10 kHz. Losses due to change in output level when one source feeds one or 40 destinations ('Teeing' losses), exceed the specified figure at 16 kHz, but the losses at lower

Table 2 Audio performance

Parameter	Specification	Achieved
Return loss		
(1) Input		
(a) 50 Hz to 100 Hz	>15 dB	38 dB
(b) 100 Hz to 10 kHz	>20 dB	38 dB
(c) 10 kHz to 16 kHz	>15 dB	16 dB
(2) Output		
(a) 50 Hz to 5 kHz	>20 dB	38 dB
(b) 5 kHz to 10 kHz	>12 dB	30 dB
(c) 10 kHz to 16 kHz	>9 dB	14 dB
Teeing loss		
With any change in number of destinations	≯0·25 dB	
With increase from 1 to all destinations		
(a) 50 Hz		−0·05 dB
(b) 1 kHz		−0·05 dB
(c) 16 kHz		+0·4 dB
With increase from 1 to at least 13 destinations in the same group, i.e. in group 1–20 or group 21–40, at 16 kHz		+0·25 dB
Frequency response		
(a) Range 50 Hz to 10 kHz	±0·2 dB	+0·15 to −0·2
(b) Range 10 kHz to 16 kHz	+0·2 to −0·5 dB	−0·05 to −0·45

Parameter	Specification	Achieved
Gain change with level		
With constant nominal gain, a change of output level from +10 dBm to −20 dBm shall produce a gain change at		
(a) 50 Hz	≯0·3 dB	0·34 dB
(b) 100 Hz to 16 kHz	≯0·1 dB	0·12 to 0 dB
Harmonic distortion		
Signal/harmonic distortion ratio		
(1) 100 Hz		
(a) At +10 dBm	≮50 dB	51·2 to 56·8 dB
(b) At +17 dBm	≮42 dB	46·5 to 56·8 dB
(2) 1 kHz		
(a) At +10 dBm	≮55 dB	56·4 to 65 dB
(b) At +17 dBm	≮47 dB	50·8 to 69 dB
Noise (measured on t.p.m.)		
(1) Random		
(a) Unweighted	≯−58 dBm	−66 to −69 dBm
(b) Weighted	≯−68 dBm	−66 to −68 dBm
(2) Impulsive, unweighted	≯−58 dBm	−59 to −65 dBm
Crosstalk		
Attenuation at		
(a) 50 Hz to be greater than	80 dB	>90 dB
(b) 1 kHz	90 dB	> 100 dB
(c) 10 kHz	80 dB	86 dB
(d) 16 kHz	76 dB	82 dB (worst cases)

frequencies are well within specification. The weighted random noise does not differ appreciably from unweighted noise and is very slightly worse than the specified −68 dBm.

4.5. *Monitoring Units*

To enable the operator to keep a check on inputs and outputs of the vision and sound switchers, source and destination monitoring units are provided. Although in this system quality checking is carried out elsewhere, the monitoring units are designed to broadcast quality standard; for example, vision crosstalk is better than 60 dB at 4·5 MHz, differential phase is better than 0·2°, etc. Each unit is capable of switching up to 50 vision and sound signals and also provides control of source or destination indicators mounted under the picture monitors in the control console.

Vision signals are supplied to the monitoring units at a level of −21 dB rel. 1 V and after the selection of the required input on the motor uniselector wiper the signal is amplified to 1 V. Sound circuits are entirely passive and a loudspeaker amplifier and small loudspeaker are provided on the main control panel. During the rotation of the uniselector, outputs of the unit are muted to prevent 'flashing' of unwanted inputs as the wipers travel to the selected position.

Each monitoring unit consists of a motor uniselector, uniselector driver, vision amplifier and muting relays mounted in a dust-proof box. The control of monitoring is carried out by the same source and destination buttons that are used to load the stores.

4.6. *Store*

One motor uniselector per destination is used as the basic memory element in the 'store' and 'next event store'. All storage uniselectors carry only d.c. circuits and they are wired to accept the ultimate number of sources in each centre.

4.6.1. Motor uniselector control

The motor uniselector (m.u.) is well known to telephone exchange equipment designers.[5] Basically, control of the m.u. consists of operating and releasing the latch magnet. Because of the high speed of the m.u. (about 200 steps/s) any control circuit must respond very quickly to a signal telling the m.u. to stop. The latch magnet must be de-energized within about 1 ms of the uniselector reaching the required stopping point.[6]

A standard circuit for control of motor uniselectors using a high-speed relay would require an excessive marking current as in this application up to 40 uniselectors can be running at the same time. Therefore a new control circuit was designed using a silicon controlled rectifier instead of the high-speed relay.[7] The circuit of this new uniselector driver is shown in Fig. 11. Operation of relay A from an external switch (dest.) causes the s.c.r. to fire via C1, R1, and contact A1, energizing the latch magnet (LM). The latch magnet contact, lm, in addition to powering the motor coils operates relay B and locks up relay A via contact A2.

The required stopping position is marked by an earth connection to a point on one of the uniselector levels whose wiper is connected via contact A3 and C2 to the cathode of the s.c.r. During the period while the m.u. is running C2 charges and when the wiper reaches the earthed point the charged capacitor is effectively connected across the s.c.r., turning it off and releasing the latch magnet.

Obviously signals can pass either way through the uniselector contacts. For network switching purposes what is termed a 'source seeking' switcher is used, that is one uniselector is used per output or destination and inputs or sources are connected in a multiple along the banks of all selectors. In this way one destination may be connected to only one source at a time, but one source may feed many destinations.

4.7. *Time Control System*

The network switching equipment, as well as providing for pre-setting and storing of two patterns of connections, allows for pre-setting their scheduled switching times. The actual switching of routes according to the pattern set-up in the next event store is under the control of a 24-hour real-time clock driven by pulses derived from speaking-clock TIM.

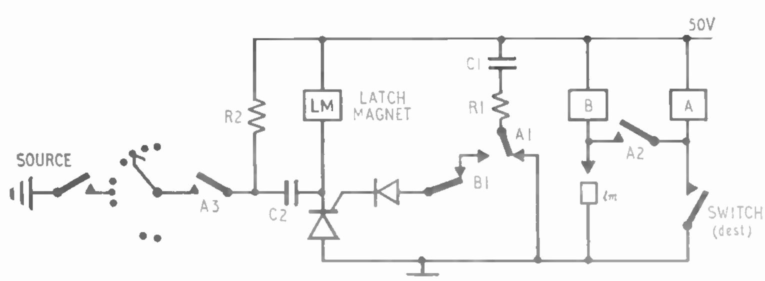

Fig. 11. Motor uniselector control.

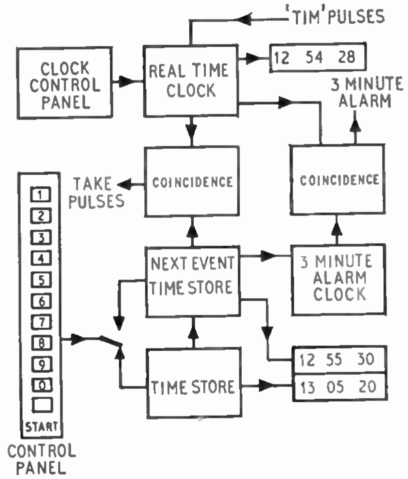

Fig. 12. Block diagram of time control system.

4.7.1. General description

The block diagram of the time control system is shown in Fig. 12. Sequential operation of the required non-locking time entry buttons on the main control panel will enter the desired time of the switching operation into the time store.

When the next event is transferred to 'on-air' the time store entry moves into the 'next event store' and appears on the 'next event' time display. The time entry pre-set in the 'next event time store' is compared with the real-time clock, and on coincidence a 'take' pulse changes the connections in the vision and sound switcher to positions determined by the 'next event store'. Then information is transferred automatically from stores to the 'next event stores'. Corrections of the time entry in both stores can be carried out at any time. An alarm clock gives an aural warning (chimes) to the operator 3 minutes before a clock 'take' is due.

The real-time clock consists of a semiconductor ring counter (for units seconds), driven by 1-second pulses, and two parallel chains of miniature uniselectors (for other digits) driven by 10-second pulses. Both 1-second and 10-second pulses are derived from TIM. The ring counter does not take any part in the switching operations, the units seconds display is provided simply for the convenience of the operator.

Alarm circuits warn the operator of any malfunctioning of the clock, and also immobilize the clock 'take' circuits. The operator must now control manually until the clock has been repaired, reset and restored to service.

The clock control panel mounted in the control console is provided for adjustment of the real time clock. A loudspeaker mounted on the panel enables the operator to check the clock against TIM.

4.7.2. Circuit details

The real-time clock and time stores are housed in a single unit. This contains the miniature plug-in uniselectors used for the real-time clock and stores, auxiliary relays and a solid-state ring counter.

The type of uniselector used has six independent levels of contacts and each level has 12 positions, which enables a single uniselector backed up by relay logic to perform the many functions required of a clock digit store. The first 10 positions correspond to the digits 0 to 9.

Miniature uniselectors can be either pulsed on one step at a time, or self driven via the interruptor contact, the wiper stepping continuously until the energizing voltage is removed. Both methods of driving are used in the real-time clock shown in Fig. 13.

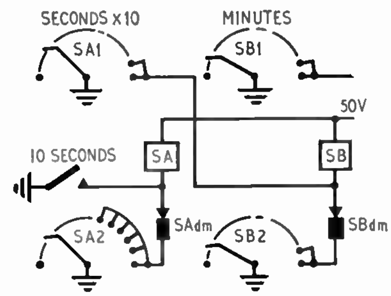

Fig. 13. Real-time clock circuit.

The relay contact driven from the G.P.O. clock pulse unit, closes for 100 ms every 10 seconds. Each time this happens the coil of the seconds × 10 uniselector SA is energized and the wipers step on one position. After six 10-second pulses have elapsed, in other words one minute, the wipers step on to position 6 and the uniselector self-drives back to zero position by virtue of the circuit on level 2 of the uniselector. As the wipers move past the last two positions, uniselector coil SB is pulsed once and the wipers step on one position. The stepping continues with each sixth pulse on the seconds × 10 uniselector moving the minutes uniselector wipers one position, and each tenth pulse on the minutes uniselector moving the minutes × 10 uniselector wipers one position, etc. Thus the chain of five uniselectors acts as a 24-hour clock.

The time is read into the time store uniselectors by means of time entry buttons on the main control panel and Fig. 14 illustrates the basic principle of running the uniselector to the desired position. When a button is pressed the uniselector coil is energized via a diode, the normally-closed relay contact A1 and the interrupter contact. At the same time a position of a

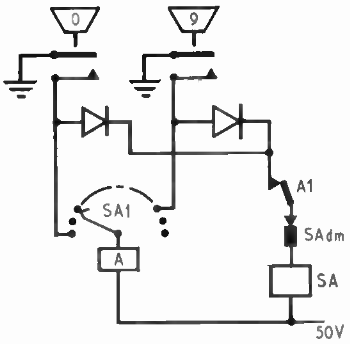

Fig. 14. Store read-in control.

level of the uniselector is marked with an earth from this button. Consequently the uniselector wiper arm steps until it reaches the marked position when the circuit is completed to energize relay A. The contact A1 opens to de-energize the uniselector coil and arrest the wiper at this selected position. This simple control circuit is common to all the store time uniselectors and is applied to each in turn by means of another uniselector acting as a sequence switch.

When the real time coincides with the time pre-set in the alarm clock chimes are sounded. Similarly when the real time coincides with the time entry in the 'next event time store' a 'take' occurs. The detection of coincidence is achieved by a series circuit on a separate level of all uniselectors in the real time clock and pre-switch alarm clock (Fig. 15). A similar circuit compares the real time clock and the 'next event time store'. The circuit is completed when the corresponding uniselectors of each chain are on the same position and the relay A operates.

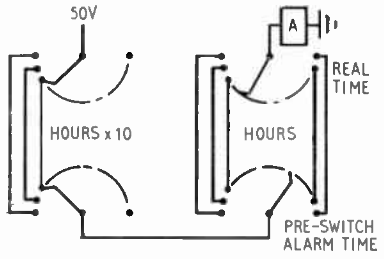

Fig. 15. Coincidence circuit.

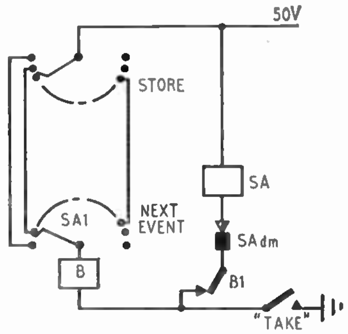

Fig. 16. Store/next event transfer.

When the 'take' occurs, the 'on-air' pattern is changed and the information from the store is transferred into the 'next event store'. The principle of this transfer is illustrated in Fig. 16 in the case of the hours digit. When the 'take' occurs, the circuit to the coil of the 'next event time' uniselector is completed and the uniselector wipers step until the wiper position of this uniselector coincides with that of the store time

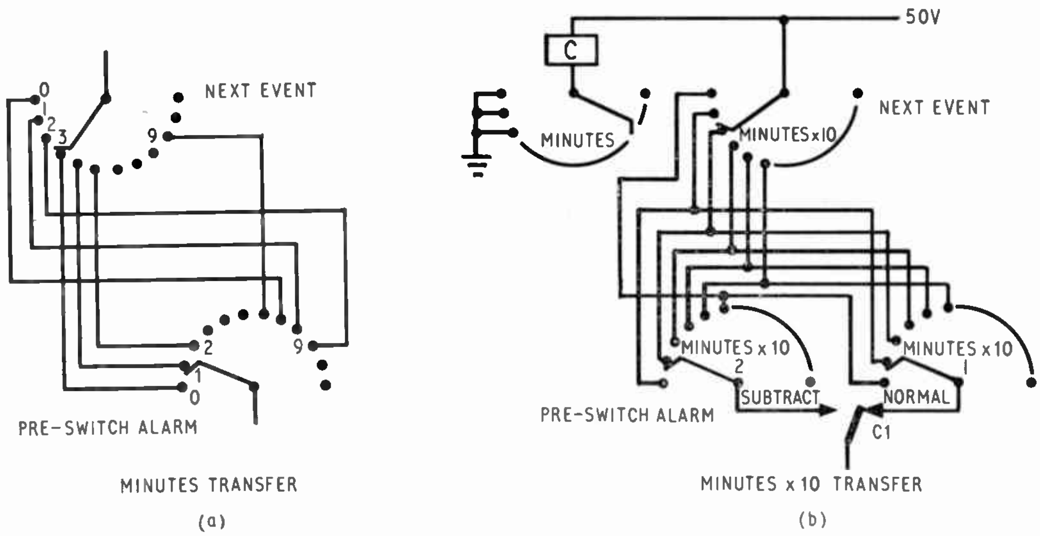

Fig. 17. Next event time/pre-switch alarm transfer.

uniselector, in this case one position. Then the relay B operates to break the circuit to the coil and arrest the uniselector at this position. All digits are transferred simultaneously using identical circuits.

Upon completion of this transfer the 'next event' time has to be transferred to the pre-switch alarm store with a three-minute subtraction. If, for example, the 'next event' time is 09 49 00 then the required pre-switch alarm time is 09 46 00. The seconds × 10 digits can be transferred as just described. The minutes transfer is effected by connecting the positions of the 'next event' level to the corresponding positions less three on the pre-switch alarm level, i.e. 9 to 6, 8 to 5, etc., through to 0 to 7 as shown on Fig. 17(a). However, in addition, when the 'next event' time minutes digits are less than three, the minutes × 10 digit will require subtraction by 1. For instance the pre-switch alarm time for a 'next event' time of 09 hours 42 minutes is 09 hours 39 minutes, a reduction in the minutes × 10 digit from 4 to 3. This requires additional circuits shown in Fig. 17(b) to perform the subtraction. Normally a straightforward transfer takes place on level 1 when position 0 is connected to position 0, etc. If the minutes 'next event' time selector wiper is on positions 0, 1 or 2, relay C is energized and contact C1 changes over so that 1 is connected to 0, etc., and subtraction occurs. This process may of course be required for higher digits as well.

5. Conclusions

The new network switching equipment has now been installed and thoroughly tested, and at present is being used for training the operating and maintenance staff. Because of the reliability demanded of equipment whose maloperation may be evident to millions of viewers, it is likely that the switchers will be operated for some time to the normal daily schedules but not using the line traffic.

Initially, the loading of the stores will be carried out manually from the control panel. But the system is designed in such a way that an automation module can be added in parallel with the manual panel so that punched tape or cards can be utilized for reading-in of the information into the stores. The tape or card reader could be located either in the network switching centre or at a remote point with a suitable data link connection. It must, however, be remembered that due to last minute changes and emergencies, even on very rare occasions, complete automation will not eliminate the operator. A skilled operator will still be required to carry out rapid re-routing.

After a carefully planned changeover of circuits to the new equipment it is confidently expected that in a short time the operators will find their duties less onerous and that the fault-rate, due to either mis-operations or equipment failure, will be reduced still further.

6. Acknowledgments

The authors wish to thank the Senior Director of Engineering of the British Post Office and the Director of Engineering, The Marconi Company Limited, for permission to publish this paper.

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The *Iris* Satellite

Iris, known before launch as *ESRO II*, is the European Space Research Organization's first satellite to be placed in orbit, and was launched from the Vandenberg Western Test Range in California on 16th May 1968. It was placed in an orbit very close to the planned one, having a perigee of 326 km, and an apogee of 1086 km. The satellite completes a circuit of the Earth once every 98·9 minutes, and as the orbit is inclined at 97·2° to the Equator, the satellite will be in continuous sunlight for 192 days. These are virtually ideal conditions for the accomplishment of *Iris's* scientific mission which requires that the satellite must pass over the poles and be continuously in sight of the Sun, and must pass through the lower layer of the Van Allen radiation belt. In passing close to the poles the satellite will traverse the various levels of the radiation layers and will also be able to measure the particles in this region that reach the Earth directly from the Sun.

All the on-board systems are working perfectly and most of the scientific experiments have already been switched on by telecommand and have started sending back their first observations.

In addition to the British prime contractor, three other British, seven French and three American companies have participated in the project. Although other satellites have been built in Europe, mainly by groups from one nation, the international character of the ESRO project is worthy of particular note, as it has involved experimenters, industrial enterprises and project staff from six different nations.

In addition to the seven experiments, the satellite carries its own electronics for operation and control, and sufficient externally-mounted solar cells to provide power during the one year that it will be in orbit. The weight limit imposed by the specifications was 80 kg, about 6 kg of which is accounted for by the standard adapter and separation system. In order to make as much of this weight as possible available for the scientific and technical equipment, most of the structure has been manufactured in magnesium alloy or aluminium.

Scientific Aims

The main objective of *Iris* is the study of solar astronomy and cosmic rays. To achieve this objective, the seven scientific experiments carried by the satellite will investigate:

- the electromagnetic radiation emitted by the Sun, with a view to reaching a better understanding of the Sun itself, especially during its active phases,
- the corpuscular radiation from the sun during solar flares,
- protons trapped in the inner Van Allen belt,
- the electron component of primary cosmic radiation,
- the modulation mechanism of cosmic rays in interplanetary space.

The *Iris* experiments have their counterparts in the NASA *OSO* series and *Iris* will provide continuity to the solar radiation observations, carried out by *OSO-D* approximately a year prior to the launch of *Iris*. In addition, the particle experiments will continue similar measurements carried out by the *Ariel I* satellite.

The design, development, construction and financing of the experiments has been the responsibility of the institutions sponsoring them, the design being compatible with agreed interface specifications issued by ESRO. Details of the experiments are as follows:

*Experiment S*25. Routine monitoring of energetic particle flux (Sponsored by a group headed by Prof. H. Elliot and Dr. J. J. Quenby, Imperial College, London).

The experiment will measure the flux of energetic charged particles in the vicinity of the earth, using two Geiger-Müller counters. One of these, an Anton 302 counter, has been widely used in earlier satellites and the data obtained from it in *Iris* will be correlated with some of the earlier data. In particular, they will be correlated with the measurements made by *Ariel I* on the particle fluxes near the lower edge of the inner radiation belt with a view to relating variations in the intensity of these particles to changes in atmospheric density. This type of study is important in establishing the lifetimes and origin of trapped particles in this region.

The second counter, an Anton 112, will study particle fluxes below the main trapping region of the inner belt. The geometrical factor of this counter, fifty times greater than that of the Anton 302, will enable the low intensities of the trapped radiation to be recorded with better statistical accuracy.

*Experiment S*27. Measurement of solar protons and inner Van Allen belt protons (Prof. H. Elliot and Dr. R. J. Hynds, Imperial College, London).

The aim of this experiment is to measure the energy spectrum of protons in the range 1 to 100 MeV that are found in the inner Van Allen belt and that also form part of the energetic solar flare particle flux. The detector is designed to acquire data on the spectra and intensity variations of these two distinct particle populations, and to obtain information on the geomagnetic thresholds at higher latitudes and on the way in which these change during periods of geomagnetic activity.

One of the problems in the design of the telescope used is to reduce to a minimum the contribution of alpha particles to the proton counting rate. The method used to do this makes available data on alpha particles in the energy ranges 8 to 60 MeV and 37 to 60 MeV. Such data are of considerable interest, since the existence of alpha particles in the magnetosphere would be a very significant clue to the mechanism of particle injection into the magnetosphere. The proton detector used is a telescope comprising four solid-state detectors separated by small amounts of absorbing material.

*Experiment S*28. Solar and galactic alpha-particles and protons (Prof. H. Elliot and Dr. J. J. Quenby, Imperial College, London).

The prime purpose of this experiment is to measure the time-dependence of the flux ratio of protons and alpha-particles of the same magnetic rigidity (0·4 to 0·8 GV) which are emitted from the sun during energetic

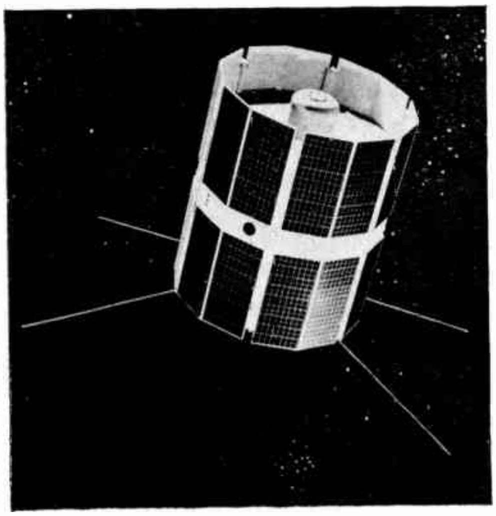

Fig. 1. The *Iris* satellite.

particle events. These data will help to throw light on the modulation mechanism that acts upon cosmic-ray particles in interplanetary space. A further objective is the investigation of geomagnetic threshold rigidities; the instrument will monitor the flux of relativistic protons and alpha-particles. The sensor consists of a telescopic arrangement of four detectors.

Experiment S29. Primary cosmic ray electrons (Prof. J. G. Wilson and Dr. P. L. Marsden, University of Leeds).

The main purpose of the experiment is to measure the flux and energy distribution of primary cosmic ray electrons in the GeV range. Such data are relevant to theories on the origin and acceleration of cosmic ray particles, and when combined with radio noise observations provide a sound basis for estimation of the galactic magnetic field strength.

The principle of the experiment is to use a gas Cerenkov counter with a beta-threshold sufficiently high (more than 20 GeV) for protons to reduce the contaminating background by a factor of about 20, and to distinguish electrons from these protons by requiring a certain degree of cascade multiplication in a lead absorber.

Experiment S36. Study of the variations of solar x-rays between 1 and 20 Å (Prof. E. A. Stewardson, University of Leicester, and Prof. R. L. F. Boyd, University College, London).

The experiment will measure the flux and spectrum of solar x-rays in the wavelength range 1 to 20 Å. The more energetic part of this radiation is closely related to solar flares and shows great variation in time. Its study will contribute to the knowledge of solar flare phenomena and their correlation with the ionosphere. The experiment will also provide precise information on the absolute intensity, spectral slope and variability of the 'non-flare' Sun, mainly in the softer region (about 10 to 20 Å).

An array of five gas-filled proportional counters is used to detect the x-ray photons. In the region 1 to 9 Å, where the largest time variations occur, two pairs of counters with different sensitivities provide a large dynamic range.

Experiment S37. Study of the variations of solar x-rays between 44 and 60 Å (Prof. C. de Jager, University of Utrecht).

The experiment has been designed to monitor the flux of solar x-rays in the wavelength band between 44 and 60 Å, in order to obtain data which will assist in explaining the various phenomena that occur during solar disturbances. Furthermore, the intensity of radiation emitted in this range during quiet periods is of great interest in solar and ionospheric studies.

The sensors are two proportional counters filled with a mixture of argon and methane at a total pressure of 0·5 to 1·0 atmosphere. The windows of the counters are made of mylar, 6·4 microns thick, and allow photons in the wavelength band 44 to 60 Å to penetrate without too much absorption. The counters are placed on opposite sides, at the equator of the satellite.

Experiment S72. Flux and energy spectra of solar and galactic cosmic ray particles (Dr. J. Labeyrie, Centre d'Etudes Nucléaires de Saclay).

The main object of this experiment is to measure the flux and energy distribution of protons between 35 and 1000 MeV that either belong to the galactic cosmic rays or have been emitted by the Sun during a flare. The experiment will also measure the flux and spectrum of alpha particles with energy between 140 and 1200 MeV, as well as the flux of relativistic lithium, beryllium and boron nuclei. It is hoped that the results obtained will contribute to the understanding of solar flares, particularly regarding the acceleration of energetic particles on the Sun. When evaluated in conjunction with other observations, the experiment should also provide information on the modulation of galactic cosmic rays in interplanetary space.

The sensor consists of three solid-state detectors, two of which are disks forming a telescope in coincidence. The

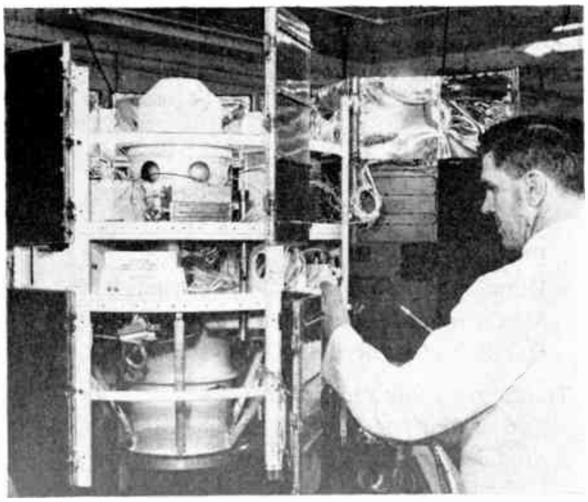

Fig. 2. A cosmic ray experiment being installed in a prototype *ESRO II* satellite for electrical integration tests. Four of the twelve solar cell panels fitted with protective covers are open revealing some of the thermal control blankets.

third has the shape of an open rectangular box surrounding the telescope in order to eliminate nuclear shower events by means of an anti-coincidence circuit.

Design of the Satellite

The British company, Hawker-Siddeley Dynamics Ltd., was selected as main contractor for the project, with Société Engins Matra of France as main sub-contractor.

Under the provisions of the contract, HSD was responsible for:

- structural, dynamic and thermal analysis of the project,
- design-study and development of the structure and control system (rate of spin, attitude of the spin axis), the power-supply system, the instrumentation ensuring the operation of the on-board equipment, and the antenna system,
- construction of mock-ups in order to determine the location of the scientific apparatus and technical equipment, the behaviour of the structure under the stresses and vibrations that occur at launching, and the thermal balance in the anticipated flight conditions,
- construction of two prototypes for various tests, and of two flight models,
- assistance to the Organization in the various trials and launching tests, particularly by supplying certain handling and measuring equipment.

Telemetry and Data Storage

The telemetry system employs pulse code modulation and will provide two data links (real-time and recorded data), each transmitting identical scientific information. In addition, 'housekeeping' or monitor data will be transmitted to enable the satellite performance to be checked and its operating mode verified.

The real-time link will operate continuously, transmitting both scientific and 'housekeeping' data as they are produced. Transmissions may be received at any suitably equipped station and are compatible with conventional STADAN receiving equipment. The carrier frequency of the real-time transmitter will also be used for tracking purposes.

Transmitted data characteristics are:

Bit rate	128 bits/second
Word length	8 bits
Frame length	32 words
Duration of frame	2 seconds
Master frame length	8 frames
Duration of master frame	16 seconds

Transmitter characteristics are:

Carrier frequency	136·89 MHz
Modulation	p.c.m./f.m./p.m.
Radiated power	0·2 W continuously
Information bandwidth	10 kHz

Data will also be stored on board by a magnetic tape recorder in order to preserve information obtained during periods when the reception of real-time transmission is not possible. The stored data can be played back 32 times as fast as recorded, so that data from one orbit (about 100 minutes) can be replayed in about 3 minutes, and transmitted by the high-power, stored-data transmitter. The telemetry link for this data is initiated by ground command and has a transmission rate of 4096 bits/second.

Transmitter characteristics are:

Carrier frequency	136·05 MHz
Modulation	p.c.m./p.m.
Radiated power	2 W on command
Information bandwidth	50 kHz

The two telemetry transmitters are coupled via a duplexer and hybrid ring to a four-element turnstile antenna, each element being about 60 cm long. They are deployed parallel to the satellite longitudinal axis and are attached to the lower end of the solar cell longerons.

The satellite will be controlled in flight by 36 commands, using a tone-digital command system compatible with GSFC standards and operating on a frequency of 148·25 MHz.

Power System

The power for the satellite is provided by the conversion of solar energy to electrical energy in 3456 n-on-p solar cells. In addition to supplying power to the load, the solar cell array also charges a 16-volt nickel-cadmium battery, which provides sufficient power for full satellite operation during eclipse periods of up to 37% of orbit time.

Attitude Control System

As deviations from the nominal injection data, launch date and launch time could introduce errors of up to 50° in the required initial attitude of the satillite's spin-axis, an automatic, on-board attitude-control system is used to maintain this axis within $\pm 10°$ of the normal to the solar aspect line. The control is achieved by a magnetic torquing system in which control torques are generated within the satellite. The satellite field is obtained energizing an electromagnet—commonly referred to as a magnetorquer—positioned parallel to the spin-axis. The information for the magnetorquer actuation is obtained from a combined sun angle/earth's magnetic field sensing system. The sensor information is fed to, and analysed by, an on-board logic system, which forms the main part of the attitude-control system, and which generates the appropriate current polarities and switch-on times. The automatic on-board system is subject to ground-command override, should the on-board logic fail. In such a case, attitude information is obtained from the on-board magnetometer and an analogue sun sensor measuring the pointing-angle error.

The data transmitted by the satellite are picked up by the French CNES and American NASA ground stations, the Norwegian station at Tromsö, and the stations of ESRO (ESTRACK) network—the latter being in operation for the first time. The data received by the ESTRACK station at Redu (Belgium) and the Tromsö station are transmitted in real-time to the ESRO Operations Centre at Darmstadt (Germany)—also operational for the first time—where they undergo real-time computer processing.

Photographs by courtesy of Hawker-Siddeley Dynamics Ltd.

U.D.C. 621.373

Propagation of Uniform Plane Waves in a Fabry-Perot Resonator containing Ammonium Dihydrogen Phosphate (ADP) as Dielectric

By

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AND

C. H. CLAYSON†

Summary: The propagation of uniform plane electromagnetic waves in a uniaxial medium has been developed from an electrical rather than an optical viewpoint. Explicit formulae for the ratios of the magnitudes of electric field components along the three co-ordinate axes are worked out for ordinary and extraordinary wave normals, and the direction of energy propagation for both waves is also developed. The results are applied to a Fabry–Perot resonator with a dielectric of ADP and it is shown that the wave normal resonance angle is the same when the incident and reflected wave velocities inside the resonator differ by a maximum as when they are identical. This leads to circular interference fringes as with an isotropic medium which is a desirable feature in a Fabry–Perot intensity modulator.

1. Introduction

The use of ammonium and potassium dihydrogen phosphate (ADP and KDP) in optical modulators[1-4] and switches has presented the electrical engineer with the problem of rethinking the conventional optical approach to birefringence in terms with which he is familiar. Some recent effort has already been made in this direction[5] and in the course of some recent research work involving a Fabry–Perot resonator containing ADP it was found desirable to carry this further.

The initial problem is to examine the nature of the propagation in a medium where the dielectric constant along one crystal axis, ε_z, is different from those along the other two axes which have ε_x equal to ε_y. The direction of the propagation may be at any angle to this ' different ' z-axis, which is commonly referred to as the optic axis of the uniaxial crystal.

The result of this anisotropy in permittivity is that there is no longer a single-phase velocity which is a characteristic of the medium, but two separate velocities, referred to in optics as those of the ordinary and extraordinary wave normals. This birefringence is already familiar to the electrical engineer in the solution to the helical aerial problem when it is analysed by the sheath helix method.[6] There, however, the anisotropy is one of conductivity rather than dielectric constant, since the solution results from an assumed infinite conductivity in the helical direction and zero conductivity along the two orthogonal axes. Also these axes refer to a set of cylindrical co-ordinates, whereas here a medium is to be considered in which rectangular co-ordinates are appropriate.

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When such a medium is placed as a dielectric between two parallel, infinite and perfectly conducting sheets, a low-loss Fabry–Perot resonator is formed. A resonator of this type filled with an iosotropic dielectric acts as a filter having a number of resonances of extremely narrow bandwidth.[7] With a uniaxial anisotropic dielectric the behaviour is described in the following sections, this analysis is applicable to the case of ADP. In this case, however, where the crystal is electro-optic as well as uniaxial, an additional property is available in that the filter resonant frequencies are a function of the electric field strength applied to the medium. This property enables the Fabry–Perot, in resonance with a monochromatic source, to be rapidly switched to the anti-resonant or other state and back again if necessary.

2. Uniform Plane Wave Equation for the Infinite Uniaxial Medium

The wave equation for a uniform plane wave travelling in an infinite anisotropic medium takes the form

$$\beta^2 \mathbf{E} - (\boldsymbol{\beta}\mathbf{E}) = \omega^2 \mu \mathbf{D} \qquad \ldots\ldots(1)$$

where β, $\boldsymbol{\beta}$ are the scalar and vector phase constants, and **E**, **D** are the vector electric and displacement fields (see Appendix 1). When propagation takes place in an isotropic medium it is well known[8] that there is no component of electric field in the direction of propagation $\boldsymbol{\beta}$. Hence the scalar product term in eqn. (1) is zero and there remain the three scalar equations

$$\beta^2_{xyz} = \omega^2 \mu \varepsilon_{xyz}$$

Consider what happens, however, when the medium is anisotropic. Let the wave be launched into this medium by a uniform plane wave in air normally incident on it, as shown in Fig. 1.

As shown in Appendix 1, Maxwell's equation gives:

$$-\boldsymbol{\beta}\times\mathbf{H} = \omega\mathbf{D} \qquad \ldots\ldots(2)$$

it follows that **D** is perpendicular to **β**. Using the relations

$$D_{xyz} = \varepsilon_{xyz}E_{xyz}$$

it follows from the figure that for $\varepsilon_z \neq \varepsilon_{x,y}$, there is a component of the resultant electric field **E** in the direction of propagation unless propagation is along the optic axis itself. Hence **β**.**E** is in general not zero,

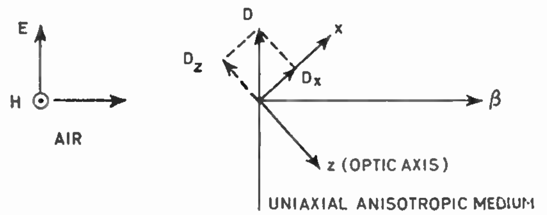

Fig. 1. Path of plane wave in an anisotropic medium.

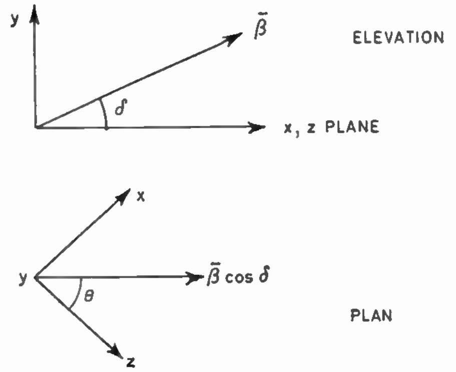

Fig. 2. General case of propagation in any arbitrary direction.

but it can be seen from the scalar components of eqn. (1) and as shown in Appendix 1, that very simple results are obtained under the following particular conditions:

(i) When **β** is orthogonal to one of the co-ordinate axes the phase velocity is determined by the dielectric constant along that axis.

(ii) When a rectangular component of electric field is zero, the ratio of the other two components is readily obtained. Thus if E_x is zero, then

$$\frac{E_y}{E_z} = -\frac{\beta_z}{\beta_y}$$

For the general case of propagation in any direction, let **β** make an angle δ with the xz plane, and its projection on this plane an angle θ with the optic (z) axis as shown in Fig. 2.

Appendix 2 gives the permissible values for β^2 as can be found in standard text books[7] but proved by other methods.

For the ordinary wave

$$\beta^2 = \omega^2\mu\varepsilon_y = \omega^2\mu\varepsilon_x \qquad \ldots\ldots(3)$$

and for the extraordinary wave

$$\beta^2 = \frac{\omega^2\mu\varepsilon_x\varepsilon_z}{\varepsilon_x\sin^2\phi+\varepsilon_z\cos^2\phi} \qquad \ldots\ldots(4)$$

where ϕ is the angle between **β** and the optic axis. It should be noted that eqns. (27–31) are individually satisfied for both ordinary and extraordinary waves.

3. Relations between Electric Field Components

3.1. *Ordinary Wave*

From eqn. (30) it can be seen that

$$\frac{E_x}{E_z} = \frac{\beta^2-k_z^2}{\beta^2-k_x^2}\tan\theta$$

It follows that for the ordinary wave with β equal to k_x, the value of E_z must be zero for E_x to be finite. This means that there is no electric field component along the optic-axis for this wave. Use of this result in eqns. (27), (28) or (29) gives the ratio of the other two field components for this wave

$$\frac{E_x}{E_y} = -\frac{\tan\delta}{\sin\theta} \qquad \ldots\ldots(5)$$

It will be noted that when β is in the xz plane, i.e. δ is equal to zero, then E_x is zero, unless θ is also zero. For this case propagation is along the optic-axis and as can be seen from eqn. (33) the two waves coalesce.

However, not only is the resultant total electric field perpendicular to the optic-axis, but it is also perpendicular to the direction of propagation **β**. This can be shown from the fact[8] that the cosine of the angle between two vectors (θ_1, ϕ_1) and (θ_2, ϕ_2) is given by

$$\cos\alpha = \cos\theta_1\cos\theta_2+\sin\theta_1\sin\theta_2\cos(\phi_1-\phi_2) \qquad \ldots\ldots(6)$$

Here (θ_1, ϕ_1) and (θ_2, ϕ_2) are respectively

$$\left(\frac{\pi}{2}-\delta, 0\right) \quad \text{and} \quad \left(\tan^{-1}\frac{E_x}{E_y}, \frac{\pi}{2}\right)$$

Substitution of these values in eqn. (6) gives

$$\cos\alpha = \sin\delta\frac{E_y}{\sqrt{E_x^2+E_y^2}}+\cos\delta\sin\theta\frac{E_x}{\sqrt{E_x^2+E_y^2}}$$

Using the value for E_x/E_y in eqn. (5) the right-hand side simplifies to zero. This shows that the resultant electric field is perpendicular to **β** also. Thus for the ordinary wave the resultant electric field is normal to the plane containing the optic-axis and the direction of propagation. The corresponding magnetic field from eqn. (22) in Appendix 1 can be seen to be perpendicular to $\boldsymbol{\beta}$ and the resultant electric field.

3.2. *Extraordinary Wave*

From eqns. (32) and (27) in Appendix 2, it can be seen that for the extraordinary wave

$$\frac{E_x}{E_y} = \frac{\sin\theta}{\tan\delta} \qquad \ldots\ldots(7)$$

Equations (5) and (7) indicate orthogonal transverse polarizations since the product of the gradients of these fields is -1. However, the orthogonality is applicable to more than the transverse fields. It can be shown that it is applicable also to the total fields. Using subscripts 'o' and 'e' for ordinary and extraordinary waves respectively, the following two equations can be written:

$$\mathbf{E}_o = \mathbf{i}E_{xo} + \mathbf{j}E_{yo}$$

$$\mathbf{E}_e = \mathbf{i}E_{xe} + \mathbf{j}E_{ye} + \mathbf{k}E_{ze}$$

Taking the scalar product of the above vectors we get

$$\mathbf{E}_o \,.\, \mathbf{E}_e = E_{xo}E_{xe} + E_{yo}E_{ye}$$

Using eqns. (5) and (7) to eliminate E_{xo} and E_{xe} we have

$$\mathbf{E}_o \,.\, \mathbf{E}_e = 0$$

This means that the total electric fields for the two waves are orthogonal. For normal incidence the $\mathbf{E}_e$ vector will thus lie in the same direction as the magnetic vector $\mathbf{H}_o$, and correspondingly the $\mathbf{H}_e$ vector will lie in the same direction as $\mathbf{E}_o$.

4. Direction of Energy Propagation

From Maxwell's equations for plane waves we have

$$-\mathbf{j}\boldsymbol{\beta}\times\mathbf{H} = \mathbf{j}\omega\mathbf{D} \qquad \ldots\ldots(21)$$

and

$$-\mathbf{j}\boldsymbol{\beta}\times\mathbf{E} = -\mathbf{j}\omega\mu\mathbf{H} \qquad \ldots\ldots(22)$$

Vector-multiplying eqn. (22) by $\mathbf{E}$ gives

$$-\mathbf{j}\omega\mu\mathbf{E}\times\mathbf{H} = -\mathbf{j}\mathbf{E}\times(\boldsymbol{\beta}\times\mathbf{E})$$

so that the Poynting vector becomes after expansion of the right-hand side,

$$\mathbf{P} = \mathbf{E}\times\mathbf{H} = \frac{1}{\omega\mu}[E^2\boldsymbol{\beta} - (\mathbf{E}.\boldsymbol{\beta})\mathbf{E}] \qquad \ldots\ldots(8)$$

This means that

$$\mathbf{P} = \frac{1}{\omega\mu}[E^2(\mathbf{a}_x\beta_x + \mathbf{a}_y\beta_y + \mathbf{a}_z\beta_z) - \\ -(\beta_x E_x + \beta_y E_y + \beta_z E_z)(\mathbf{a}_x E_x + \mathbf{a}_y E_y + \mathbf{a}_z E_z)] \qquad \ldots\ldots(9)$$

where $\mathbf{a}_x$, $\mathbf{a}_y$ and $\mathbf{a}_z$ represent unit vectors along the x, y and z crystal axes. The direction of the Poynting vector is referred to as the ray.

4.1 *Ordinary Ray*

It has previously been proved that $\mathbf{E}$ is perpendicular to $\boldsymbol{\beta}$, the Poynting vector. From this eqn. (8) simplifies to

$$\mathbf{P} = \frac{E^2}{\omega\mu}\boldsymbol{\beta} \qquad \ldots\ldots(10)$$

showing that it is in the same direction as the wave normal. In this equation $\boldsymbol{\beta}$ refers to the ordinary wave normal of eqn. (3).

4.2. *Extraordinary Ray*

From eqn. (9) the three components of $\mathbf{P}$ for the extraordinary ray are

$$P_x = \frac{1}{\omega\mu}[(E_x^2 + E_y^2 + E_z^2)\beta\cos\delta\sin\theta - \\ -\beta\cos\delta\cos\theta\, E_x^2 - \beta\sin\delta\, E_x E_y - \\ -\beta\cos\delta\cos\theta\, E_x E_z] \qquad \ldots\ldots(11)$$

$$P_y = \frac{1}{\omega\mu}[(E_x^2 + E_z^2)\beta\sin\delta - \beta\cos\delta\sin\theta\, E_x E_y - \\ -\beta\cos\delta\cos\theta\, E_y E_z] \qquad \ldots\ldots(12)$$

and

$$P_z = \frac{1}{\omega\mu}[(E_x^2 + E_y^2)\beta\cos\delta\cos\theta - \\ -\beta\cos\delta\sin\theta\, E_x E_z - \\ -\beta\sin\delta\, E_y E_z] \qquad \ldots\ldots(13)$$

In eqns. (11), (12) and (13) the appropriate value of β is given by eqn. (33). Each expression may be evaluated in terms of one unknown, e.g. E_x, E_y or E_z by utilizing the ratios E_x/E_z and E_x/E_y of eqns. (32) and (7). Hence the direction of the resultant ray may be found.

As in the case of a resonator with air as dielectric, resonance will occur when the phase difference along BE differs by a multiple of 2π from that along (BC+CD). (Fig. 3.) This in turn depends on whether

5. Application to Fabry–Perot Resonator with ADP as Dielectric

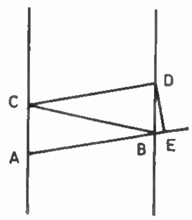

Fig. 3.

ordinary or extraordinary wave normals are travelling along the zigzag paths. In the former case the phase velocity is a constant independent of the direction of the optic-axis, and the Fabry–Perot fringes will therefore be circular in form. The existence of ordinary wave normals alone may be arranged by rotating the direction of polarization of the incoming wave by 90° clockwise from that shown in Fig. 1.

For the extraordinary wave alone, launched as shown in Fig. 1, the phase velocity depends on the angle between the wave normal and the optic-axis.

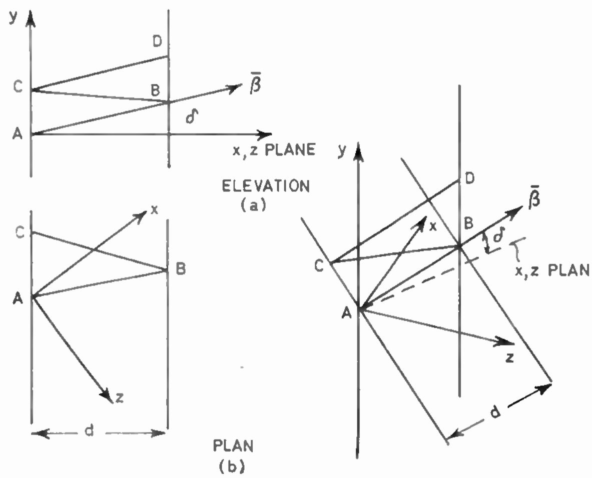

Fig. 4. Paths of a wave in two planes mutually at right angles.

Consider paths in two planes at right angles, corresponding respectively (as in Fig. 4) to $\theta = 45°$ say, with a non-zero δ, and containing the y-axis, and $\delta = 0°$ in the xz plane with θ equal to $(45°+\varepsilon)$. The angles δ and ε for resonance are considered small as in a conventional Fabry–Perot resonator, and this can readily be confirmed from the numerical results obtained in the following sections.

Case (i): $\theta = 45°$, $\delta \neq 0°$

The angle ϕ between the wave normal $\boldsymbol{\beta}$ and the optic-axis is such that for the path AB,

$$\cos\phi = \cos 45° \cos\delta$$

An identical result holds for the reflected path BC, so that the phase velocity is the same for both incident and reflected wave normals, as it is in an isotropic medium. This assumes that the angle δ is the same for both, which it very nearly is, since the law of reflection at the boundary gives

$$\frac{\delta_2}{\delta_1} \cong \frac{\sin\delta_2}{\sin\delta_1} = \frac{\beta_1}{\beta_2} \simeq 1$$

Then the condition for resonance is, writing the outgoing wave normal angle as δ'

$$\frac{2\beta d}{\cos\delta_1} - 2kd\tan\delta_1 \sin\delta' = 2N\pi \quad \text{......(14)}$$

where N is an integer. Use of the law of refraction at the boundary gives

$$\frac{\sin\delta'}{\sin\delta_1} = \frac{\beta}{k}$$

this becomes

$$2\beta d \cos\delta_1 = 2\pi N \quad \text{......(15)}$$

For the dielectric ADP, for which $\varepsilon_x = 2{\cdot}31627$ and $\varepsilon_z = 2{\cdot}15764$, and a resonator spacing of 1 inch, the phase constant β can be evaluated from eqn. (33) and is given by

$$\beta \simeq \omega\sqrt{\mu}1{\cdot}4947(1-0{\cdot}01772\delta^2)$$

For small value of δ eqn. (15) can be approximated by

$$2\beta d\left(1-\frac{\delta_1^2}{2}\right) = 2\pi N$$

and hence if there is resonance for axial wave normals there will be adjacent resonances given by

$$\beta d\delta_1^2 = 2\pi \quad \text{or} \quad \delta_1 = 4{\cdot}08 \text{ milliradians} \quad \text{......(16)}$$

This is the angle of incidence from the normal within the resonator at which the first resonance appears in the plane $\theta = 45°$, containing the y-axis.

Case (ii): $\theta = (45°+\varepsilon)$, $\delta = 0°$

For the path AB the angle made with the optic-axis is such that

$$\cos\phi_1 = \cos\left(\frac{\pi}{4}+\varepsilon_1\right)$$

but for the reflected path BC

$$\cos\phi_2 = \cos\left(\frac{\pi}{4}-\varepsilon_2\right)$$

As in the previous case $\varepsilon_1 \simeq \varepsilon_2$, but the phase constants β_1 and β_2 are not now the same as in an isotropic medium or as in the orthogonal plane. The condition for resonance is

$$\frac{\beta_1 d}{\cos\varepsilon_1} + \frac{\beta_2 d}{\cos\varepsilon_2} - kd\tan\varepsilon_1\sin\varepsilon' - - kd\tan\varepsilon_2\sin\varepsilon' = 2M\pi \quad \text{......(17)}$$

where M is an integer and ε' is the outgoing wave normal angle.

The following laws of refraction at the two mirrors

$$\frac{\sin\varepsilon'}{\sin\varepsilon_1} = \frac{\beta_1}{k} \quad \text{and} \quad \frac{\sin\varepsilon'}{\sin\varepsilon_2} = \frac{\beta_2}{k}$$

give

$$(\beta_1+\beta_2)d\cos\varepsilon_1 \simeq 2M\pi \quad \text{......(18)}$$

As before by replacing $\cos\varepsilon_1$ by

$$\left(1-\frac{\varepsilon_1^2}{2}\right)$$

and by assuming resonance for axial wave normals adjacent resonances are obtained when

$$(\beta_1+\beta_2)d\,\frac{\varepsilon_1^2}{2} = 2\pi \quad \text{......(19)}$$

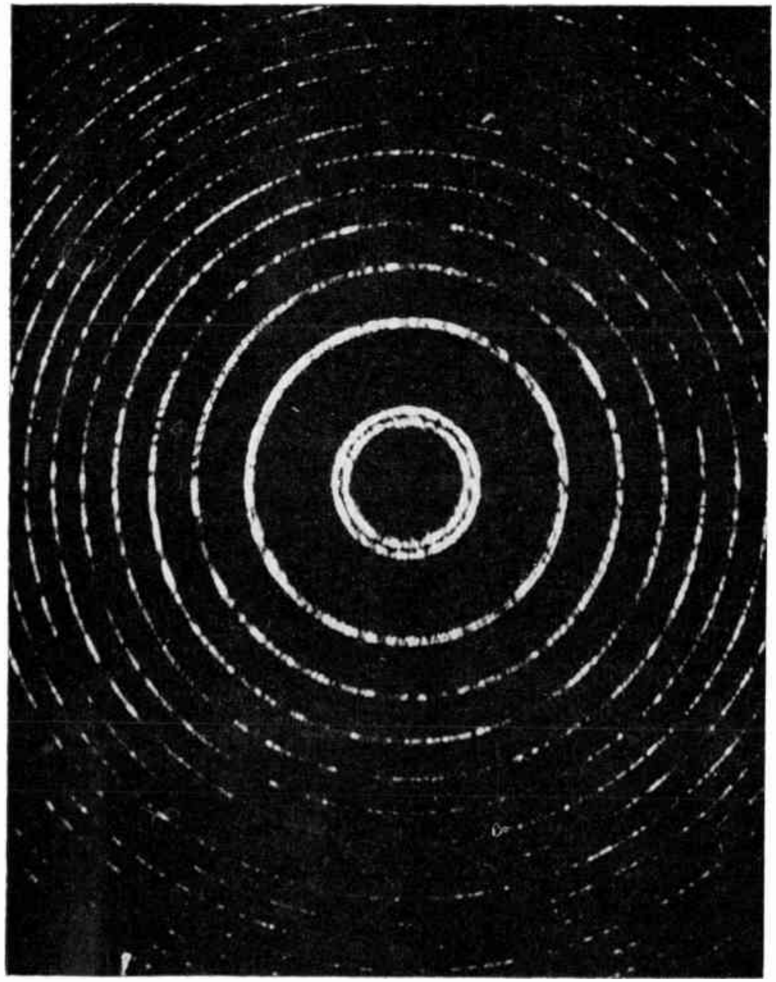

Fig. 5. Fabry–Perot fringes obtained with air as dielectric.

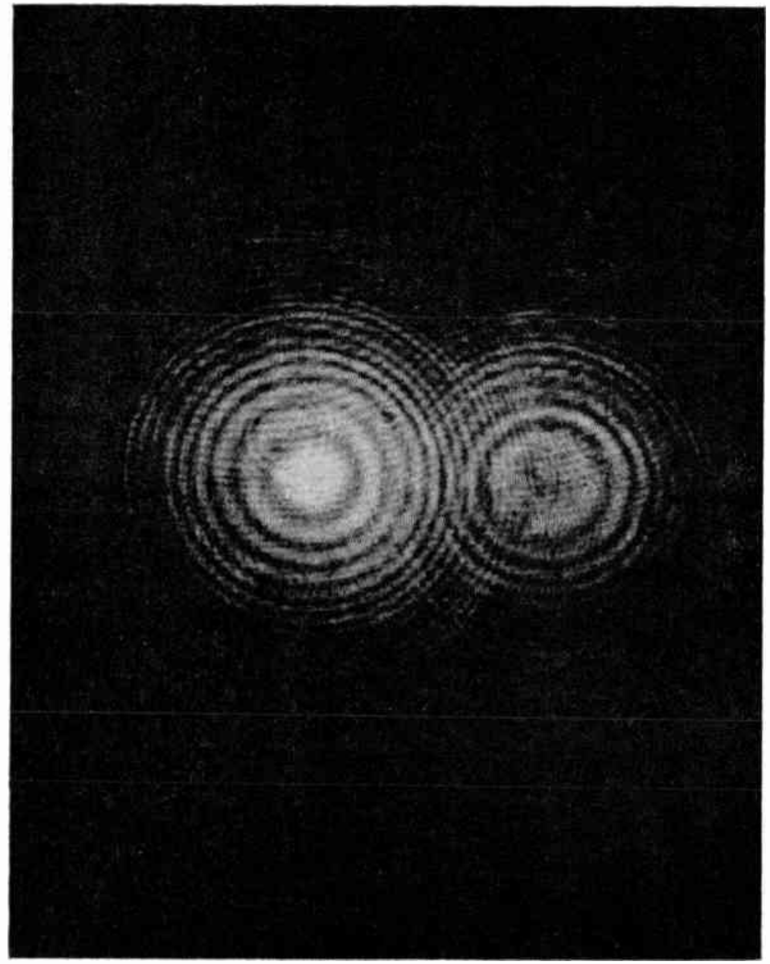

Fig. 6. Fabry–Perot fringes for ordinary and extraordinary waves in ADP.

For the numerical example cited in Case (i), eqn. (33) gives

$$\beta_1 \simeq \omega\sqrt{\mu}1{\cdot}4947(1-0{\cdot}03546\varepsilon_1)$$

$$\beta_2 \simeq \omega\sqrt{\mu}1{\cdot}4947(1+0{\cdot}03546\varepsilon_1)$$

so that resonance occurs when

$$\varepsilon_1 = 4{\cdot}08 \text{ milliradians} \qquad \ldots(20)$$

The rather surprising conclusion is reached that in two orthogonal planes between two parallel mirrors, where three different phase velocities apply, the resonance angles are approximately equal. Although these results apply strictly to wave normals rather than the observable rays, application of eqns. (11–13) leads to the same result.

Confirmation of this was obtained from an *étalon* constructed with a reflector spacing of approximately 1 inch. An ADP 45° y-cut square-shape crystal (side 0·25 inch) was inserted between the reflectors, with the crystal y-axis normal to the resonator axis. In the absence of the ADP the interference fringes were as shown in Fig. 5, where the double rings are evidence of two longitudinal axial modes in the Ferranti laser, separated by a frequency of 410 MHz. When the ADP is present and the input polarization is such as to excite both ordinary and extraordinary waves equally, the results obtained are shown in Fig. 6, where the fringes are again circular, but displaced because of the divergence of the two rays.

The fact that the rings are circular is, indeed, a desirable feature when the Fabry–Perot *étalon* with 45° x- or y-cut ADP filling is to be used as an intensity modulator, since this means that a simple anastigmatic collimating system is all that is required to produce a suitable input light beam. When, on the other hand, a single 45° x- or y-cut ADP crystal is used, alone, as a polarization/intensity modulator, the angular aperture of the modulator differs very considerably in the plane containing the crystal z-axis and (respectively) the y- or x-axis from that occurring in the plane containing the beam axis and (respectively) the crystal x- or y-axis. The practical result of this is that the depth of modulation will vary over the beam cross section in a complicated manner, unless an astigmatic collimating system is used, whereas, in the case of the Fabry–Perot electro-optic modulator, the depth of modulation falls off radially about the beam axis in a symmetrical manner.

6. Conclusions

An electrical rather than optical approach to the study of plane wave propagation in an anisotropic medium has been applied to investigate a Fabry–Perot resonator using anisotropic dielectric. It has been shown that even in the extreme case of maximum phase velocity difference between the incident and reflected wave normals the resonance angle is identical with that in the orthogonal plane where this difference is zero. The interference patterns are circular as with an isotropic dielectric, and this leads to advantages in its operation as an intensity modulator.

7. References

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8. Appendix 1

8.1. *Uniform Plane Wave Equation for the Infinite Uniaxial Medium*

Consider first a uniform plane wave travelling in any medium along the z-axis, say, with a phase constant β. Then the del operator defined by

$$\nabla = \mathbf{i}\frac{\partial}{\partial x}+\mathbf{j}\frac{\partial}{\partial y}+\mathbf{k}\frac{\partial}{\partial z}$$

becomes simply $-\mathrm{j}\beta\mathbf{k}$ for a time factor $e^{j\omega t}$. Generalizing this to the same wave travelling in an arbitrary direction gives

$$\nabla = \mathbf{i}(-\mathrm{j}\beta_x)+\mathbf{j}(-\mathrm{j}\beta_y)+\mathbf{k}(-\mathrm{j}\beta_z)$$

If β is now written as a vector quantity $\boldsymbol{\beta}$ this may be abbreviated to

$$\nabla = -\mathrm{j}\boldsymbol{\beta}$$

so that Maxwell's equations become

$$-\mathrm{j}\boldsymbol{\beta}\times\mathbf{H} = \mathrm{j}\omega\mathbf{D} \qquad \ldots\ldots(21)$$

$$-\mathrm{j}\boldsymbol{\beta}\times\mathbf{E} = -\mathrm{j}\omega\mu\mathbf{H} \qquad \ldots\ldots(22)$$

These equations give

$$-\mathrm{j}\boldsymbol{\beta}\times\left(\frac{\boldsymbol{\beta}}{\omega\mu}\times\mathbf{E}\right) = \mathrm{j}\omega\mathbf{D} \qquad \ldots\ldots(23)$$

and making use of the vector identity

$$\mathbf{A}\times(\mathbf{B}\times\mathbf{C}) = (\mathbf{A}\,.\,\mathbf{C})\mathbf{B}-(\mathbf{A}\,.\,\mathbf{B})\mathbf{C}$$

one obtains for the uniform plane wave equation

$$\beta^2\mathbf{E}-(\boldsymbol{\beta}\,.\,\mathbf{E})\boldsymbol{\beta} = \omega^2\mu\mathbf{D} \qquad \ldots\ldots(1)$$

The scalar wave equations form of eqn. (1) are

$$\beta^2 E_x - (\beta_x E_x + \beta_y E_y + \beta_z E_z)\beta_x = \omega^2 \mu \varepsilon_x E_x \quad \ldots\ldots(24)$$

$$\beta^2 E_y - (\beta_x E_x + \beta_y E_y + \beta_z E_z)\beta_y = \omega^2 \mu \varepsilon_y E_y \quad \ldots\ldots(25)$$

$$\beta^2 E_z - (\beta_x E_x + \beta_y E_y + \beta_z E_z)\beta_z = \omega^2 \mu \varepsilon_z E_z \quad \ldots\ldots(26)$$

9. Appendix 2

9.1. *Solutions of Wave Equation*

Referring to Fig. 2 the components of the phase constant β along the three reference axes of the crystal are

$$\beta_x = \beta \cos\delta \sin\theta, \quad \beta_y = \beta \sin\delta, \quad \beta_z = \beta \cos\delta \cos\theta$$

where δ is the angle which $\boldsymbol{\beta}$ makes with the xz plane and θ the angle which the projection of $\boldsymbol{\beta}$ on the xz plane makes with the z-axis. Then the wave equations become

$$[\beta^2(1-\sin^2\theta\cos^2\delta) - k_x^2]E_x - \\ -\beta^2 \sin\theta \sin\delta \cos\delta\, E_y - \\ -\beta^2 \sin\theta \cos\theta \cos^2\delta\, E_z = 0 \quad \ldots\ldots(27)$$

$$-\beta^2 \sin\theta \sin\delta \cos\delta\, E_x + [\beta^2\cos^2\delta - k_y^2]E_y - \\ -\beta^2 \cos\theta \sin\delta \cos\delta\, E_z = 0 \quad \ldots\ldots(28)$$

$$-\beta^2 \sin\theta \cos\theta \cos^2\delta\, E_x - \beta^2 \cos\theta \sin\delta \cos\delta\, E_y + \\ +[\beta^2(1-\cos^2\theta\cos^2\delta) - k_z^2]E_z = 0 \quad \ldots\ldots(29)$$

where

$$k_x^2 = \omega^2 \mu \varepsilon_x, \quad k_y^2 = \omega^2 \mu \varepsilon_y \quad \text{and} \quad k_z^2 = \omega^2 \mu \varepsilon_z$$

Elimination of E_y from eqns. (27) and (29) gives

$$\frac{E_x}{E_z} = \frac{(\beta^2 - k_z^2)}{(\beta^2 - k_x^2)} \tan\theta \quad \ldots\ldots(30)$$

Similarly from eqns. (27) and (28) again elimination of E_y results in

$$\frac{E_x}{E_z} = \frac{\beta^2 \sin\theta \cos\theta \cos^2\delta(\beta^2 - k_y^2)}{[\beta^4 \cos^2\theta \cos^2\delta - \beta^2 k_x^2 \cos^2\delta - \beta^2 k_y^2 \times (1 - \sin^2\theta\cos^2\delta) + k_x^2, k^2]} \quad \ldots\ldots(31)$$

Equation (31) is general in the sense that it applies to cases where k_x is not necessarily equal to k_y. In the case of uniaxial crystals, however, where this equality holds, the equation may be simplified, provided $\beta^2 \neq k_y^2$ to

$$\frac{E_x}{E_z} = \frac{\beta^2 \sin\theta \cos\theta \cos^2\delta}{\beta^2 \cos^2\theta \cos^2\delta - k_x^2} \quad \ldots\ldots(32)$$

If $\beta = k_y = k_x$, then eqn. (30) indicates that E_z is zero for a finite value of E_x and this corresponds to the ordinary wave solution in optics.

For $\beta \neq k_y = k_x$, equating eqns. (30) and (32) the phase constant β can be found from the known values of ε_x, ε_z, θ and δ. Thus

$$\beta^2 = \frac{k_x^2 k_z^2}{k_x^2(1-\cos^2\theta\cos^2\delta) + k_z^2\cos^2\theta\cos^2\delta} \quad \ldots\ldots(33)$$

Since the term $\cos\theta \cos\delta$ represents the cosine of the angle between the direction of propagation and the optic-axis, say, ϕ, this gives more simply

$$\beta^2 = \frac{k_x^2 k_z^2}{k_x^2 \sin^2\phi + k_z^2 \cos^2\phi} \quad \ldots\ldots(34)$$

in accordance with the value given in standard texts[7] for the square of the phase constant of the extraordinary wave.

Manuscript first received by the Institution on 10th October 1967 and in final form on 1st April 1968. (Paper No. 1193/CC9.)

Letter to the Editor

Using References to Retrieve Current Articles

SIR,

Under the section, Institution Notices ('Abstracts of Journal Papers')[1] you recently listed the various services which index this journal. May I add to your list the *Science Citation Index* and *ASCA*, which are other publications of the Institute for Scientific Information?

In the *Science Citation Index* (*SCI*), articles are indexed in two basic ways. The 'Source Index' section is an author-ordered index of the articles culled from about 1800 journals during the current year. Most of these articles carry references to the prior art (books, articles, theses, monographs, etc.) which today's authors consider to be worth citing. The 'Citation Index' section is a chronologically ordered list of this prior art; beneath each cited item is a list of the current articles from the Source Index which cite it; thus if a current article, indexed once in the Source Index, has 15 references, it will be indexed in 15 places in the Citation Index.

To demonstrate the application of this system in the 'electronics' discipline, an example follows using a recent paper from *The Radio and Electronic Engineer* as the starting point for a search.

It is assumed that a reader, having perused the paper 'System engineering for reliability and ease of maintenance' by K. F. Rankin[2] (February 1968 issue), wishes to know more about recent developments in the subject 'Equipment Reliability'.

He may look up one or more of the references in the Rankin paper in the *SCI*, and expect to find a list of current articles which have cited it.

For instance, if the mathematical aspect is of interest, Rankin's reference to the book by Lloyd and Lipow, 'Reliability: Management, Methods, and Mathematics',[3] would be a good starting point.

Referring to the 1966 *SCI*, we find that several authors cited this book in 1966, for instance R. E. Barlow in an article in the journal *Technometrics*, entitled 'Reliability growth during a development testing programme'.[4]

Similarly in 1967 the book by Lloyd and Lipow has been cited several times, for instance by L. L. Levy in an article entitled 'A Monte Carlo technique for obtaining system reliability confidence limits from components test data', published in *I.E.E.E. Transactions on Reliability*.[5]

Other slightly less obvious search strategies may be used; for instance the two articles by P. Cox referred to by Rankin were not cited in 1966. However upon looking up that author, it is observed that another article by Cox[6] in the *A.T.E. Journal* has been cited—by Groocock,[7] in this journal (*The Radio and Electronic Engineer*). Knowing from the Rankin references that Cox has written at least two articles about the subject of interest, it is reasonable to suppose that this article also may be about it, in which case the Groocock article may be as well. Upon referring to the Source Index for more detailed information about this Groocock article, we find that its title is 'Transistors—reliability, life and the relevance of circuit design'.

Other articles may be found by identifying an author with the subject. For instance, the author R. E. Barlow wrote an article in 1966[4] citing the book by Lloyd and Lipow as we already know. Seven other articles by Barlow are listed adjacently in the 1966 Source Index, including for example—'Statistical estimation procedures for burn-in process', published in the *Annals of Mathematical Statistics*.[8]

The inter-relationship of the above-mentioned articles is shown in Fig. 1. The search may be continued for as long as is considered justifiable.

For those interested, the *Science Citation Index* and its applications has been described in greater detail elsewhere.[9,10]

The above example is essentially a 'catching-up' operation, but a similar technique may be used to 'keep-up' with a subject through an associated SDI service. It will be recalled that SDI (Selective Dissemination for Information) was discussed some time ago in this Journal by Lord Mountbatten.[11] As carried out by a particular SDI service (called ASCA,[12] which also covers this Journal), a man receives a weekly list of just-published articles of interest, according to the instructions contained in his 'Profile'. Various kinds of terms may be used in a profile, including words in title, cited article, cited author, source author, or author's organization.

If a man is interested in keeping up-to-date on the subject 'Reliability', he might include in his profile the term—

'Lloyd, D. K., 1962. Reliability: Management, Methods, and Mathematics.'

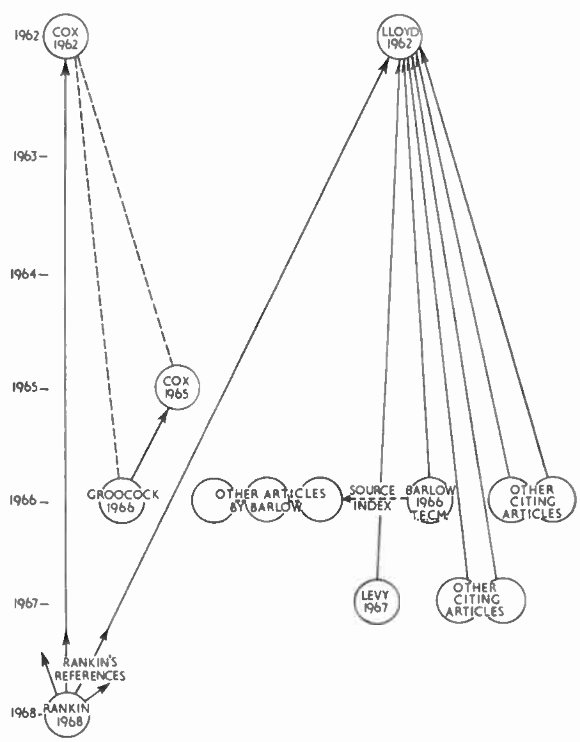

Fig. 7.

He will be informed about any current article, published in any of 1800 prime scientific journals, which cites the book by Lloyd.

Although only a few articles about the subject 'Reliability' have been mentioned, it should be observed that they were published in journals covering several disciplines. The ability to find such articles is one advantage of a multidisciplinary journal coverage. Scientific knowledge, particularly in 'Electronics' tends to cut across the conventional disciplinary divisions.

Of all the symbols in a reference, the title is usually the most helpful to a reader when he has to decide to whic cited item he should refer. In the references of papers published in *The Radio and Electronic Engineer*, full titles are normally given. It is unfortunate that this practice is not followed in all journals.

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5th April 1968

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STANDARD FREQUENCY TRANSMISSIONS

(Communication from the National Physical Laboratory)

Deviations, in parts in 10^{10}, from nominal frequency for May 1968

May 1968	24-hour mean centred on 0300 U.T.			May 1968	24-hour mean centred on 0300 U.T.		
	GBR 16 kHz	MSF 60 kHz	Droitwich 200 kHz		GBR 16 kHz	MSF 60 kHz	Droitwich 200 kHz
1	− 300·0	+ 0·1	0	**17**	− 299·9	+ 0·1	+ 0·2
2	− 299·9	+ 0·1	+ 0·1	**18**	− 299·9	+ 0·2	+ 0·2
3	− 299·9	+ 0·1	+ 0·1	**19**	− 300·1	0	+ 0·2
4	− 299·9	+ 0·1	+ 0·1	**20**	− 299·9	0	+ 0·1
5	− 299·9	+ 0·2	+ 0·1	**21**	− 300·0	0	− 0·1
6	− 299·9	+ 0·1	0	**22**	− 299·9	+ 0·1	− 0·1
7	− 299·8	+ 0·2	+ 0·1	**23**	− 299·9	0	− 0·1
8	− 299·9	+ 0·1	0	**24**	− 299·8	+ 0·1	− 0·1
9	− 299·8	+ 0·2	+ 0·1	**25**	− 300·0	+ 0·1	0
10	− 299·9	0	+ 0·1	**26**	− 300·0	0	− 0·1
11	− 299·8	0	+ 0·1	**27**	− 299·9	+ 0·1	− 0·1
12	− 299·9	+ 0·2	+ 0·2	**28**	− 299·9	+ 0·1	0
13	− 299·9	0	+ 0·1	**29**	− 299·9	+ 0·1	0
14	− 299·9	+ 0·2	+ 0·2	**30**	− 299·9	+ 0·1	0
15	− 299·8	0	+ 0·2	**31**	− 299·8	+ 0·1	0
16	− 300·0	+ 0·1	+ 0·2				

Nominal frequency corresponds to a value of 9 192 631 770·0 Hz for the caesium F,m (4,0)–F,m (3,0) transition at zero field.

Notes: (1) **All measurements were made in terms of H.P. Caesium Standard No. 134 which agrees with the NPL Caesium Standard to 1 part in 10^{11}.**

(2) Due to work at the transmitter, MSF 60 kHz transmissions were suspended during the hours 0900–1700 UT over the period May 6th to May 17th 1968. MSF readings given from May 1st to May 18th are centred on 1200 UT.

Instruments, Electronics and Automation Exhibition

The International Instruments, Electronics and Automation Exhibition of 1968 was staged at Olympia, London, from 13th to 18th May. It represented a tremendous export effort by British industries and attracted a record number of overseas visitors representing 80 countries—9431 out of a total attendance of over 112 000. About 1000 firms from 15 countries exhibited their products which ranged from simple connectors to complex factory automation systems.

The Government Establishments

Among the exhibits by research establishments of the British Government was a novel range of instruments and techniques demonstrated by the Research Councils of the Department of Education and Science.

The two component ship's log developed and used by the National Institute of Oceanography consists of a measuring head mounted on a spar projecting ½ metre below the ship's bottom. The ellipsoidal housing of the head contains a horizontal coil which generates a vertical magnetic field when energized by square wave switched d.c. Sea water flowing through the magnetic field generates a small voltage at right angles to both flow and magnetic field; this is picked up by pairs of electrodes, one fore and aft, one athwartships, mounted on the outside of the ellipsoidal housing. The voltage picked up by the athwartships electrodes will be proportional to the leeway of the ship, that picked up by the fore and aft pair proportional to the forward speed.

A tractor automatic control system is being developed by the National Institute for Agricultural Engineering. For work in which there are pronounced ground or crop features, guidance can be derived from range measurements. For operations on the ground (such as ploughing) vertical range measurements are made by an ultrasonic generator and detector which are scanned together across the front of the tractor. This enables the important feature (such as the ploughing furrow wall) to be located and followed. For the harvesting of a standing crop, the distance of the uncut vertical face can be measured horizontally by the same ultrasonic apparatus without scanning and the steering controlled to keep this constant. The signals from the ultrasonic pulse echo ranging system are processed to give 'left' or 'right' instructions, depending on the errors detected, and these can be used to actuate the hydraulic steering of the tractor.

When turning headlands the first stage of a turn is for the tractor to identify the perimeter of the field as it is approached and to measure automatically its distance from it in order to withdraw its implement, slow down, and start turning at the correct point.

An optical system was demonstrated in which the perimeter of the field is defined by a two-bar reflective fence. This is scanned in a vertical plane by a parallel beam of pulsed infra-red light, derived from a gallium arsenide lamp. Measurement of the time interval between pairs of reflections enables the distance to the fence to be measured automatically. The time interval increases for the shorter distance. A later version will employ a single-bar fence and two non-intersecting beams of light to simplify the field intersection.

The Ministry of Technology exhibited a wide range of instrumentation from eight of its research stations.

An infra-red thermal imaging technique which enables temperature distributions to be translated into a corresponding visual pattern was demonstrated by the Royal Radar Establishment. This technique has several potential industrial applications, namely monitoring temperature distributions in paper and textiles during their manufacture checking electronic circuits and power lines by observing their heat emission; testing for excessive friction and detecting flaws in various materials when the faults rise to changes in thermal conductivity.

A new acousto-electric oscillator operating in the v.h.f./u.h.f. range has also been developed. Amplification of ultrasonic waves by the motion of electrons occurs and causes the platelet to resonate at a high harmonic of its fundamental frequency. The oscillator has several important advantages over oscillators using passive acoustic resonators.

Other exhibits included a miniature Doppler radar and a phase sensitive detector.

Bio-medical Electronics

Medicine and surgery are now reaping significant benefits from the progress in industrial electronics and automation, and a number of industrial systems were shown operating in their medical context. An automated biochemical analytical machine (auto-analyser) was shown by Baird and Tatlock which can analyse samples of body fluids at a rate of several hundred per hour and requires little skill on the part of the operator.

The expanding use of very sensitive electronic instruments to register brain, heart and muscle reactions in patients gains from a new instrument which overcomes the noise in circuits designed to measure extremely small parameters. An instrument of this type, a new digital voltmeter, manufactured by Fenlow Electronics, reduces series-mode interference by a factor of about 10 000; thus the almost imperceptible electrical impulses transferred by the nerves can be measured with great accuracy.

A new automatic single patient dialysis unit—or artificial kidney machine—which is designed for either hospital, or home, use, was shown by the Cambridge Instrument Group. Built-in safety systems include dialysate pressure monitoring, blood leak detection, venous pressure monitoring and aural and visual indication of every function of the machine.

Another medical instrument was a multi-channel physiological recorder which provides simultaneous monitoring and recording of up to 6 or 12 parameters. It is able to transmit its information to remote monitors.

Overseas Products

For the first time, Canada participated in the I.E.A. Exhibition. The theme of the Canadian stand was 'Measurement and Control, Systems, Instruments and

Devices'. Products ranged from units for thickness control in steel rolling mills; flaw detection in paper manufacture, monitoring, counting and digital indication in petro-chemical processing, to aerospace data computers, radio equipment for telemetry, television and general communications.

The Scandinavian countries also took part through their agents in the United Kingdom. A system designed for high-speed drafting to a high degree of accuracy, developed by Konsberg Vapenfabrikk of Norway, and to be marketed in the United Kingdom by D-Mac Ltd. of Glasgow, was shown. The system has applications in a wide field including electronics, shipbuilding and cartography.

Oy Fima, the joint export organization for several Finnish factories, exhibited equipment for remote control and automation of different processes. Equipment for microwave, v.h.f. and u.h.f. radio links, carrier systems and test equipment was shown.

Two systems, a 'Unalog' control circuit simulator and a 'Unalog' calculator were shown by the East German foreign trade organization Feinmechanik-Optik. The control circuit simulator with indicator board uses standard amplifiers, storage resistance and transmission elements. The system is designed as a low-pressure analogue and logic system for control and automation purposes. It can be used for the simulation of all kinds of analogue controllers as well as control circuits up to fourth order and for determining the dynamics of closed-control loops. The simulator is also suitable for teaching and demonstration purposes and the apparatus is being used in universities as an aid to experimental exercises.

On the Irish Export Board stand products of several electronics firms in Ireland were shown. These included carrier frequency equipment; telemetering systems; memory cores, memory arrays and stacks; gas analysis and detection apparatus; and electronic components, e.g. capacitors, transistors, etc.

The United States had the largest single stand in the exhibition. The instruments, test and measuring equipment shown has applications in industries as widespread as steel, glass, plastics, chemicals and textiles, as well as in services like laboratories and hospitals. Other categories covered included components, data processing equipment and manufacturing machinery and tools for the electronics industry.

For computer application a new laser memory was shown. This can not only store massive amounts of information—645 million items of digital data on a square inch of tape—but can also accept information at a speed of 12 million binary digits per second. A laser with a finely focused beam 'burns' very minute holes in the tape surface. These holes can be detected and read by a second laser beam.

Components for the electronics industry including miniature valves, filters, amplifiers, semiconductor diodes and thyristors, items for precision magnetic systems, were also shown. Equipment concerned with the production of printed circuits and integrated circuits, pre-production and educational exhibits were included. One other exhibit was *Explorer II*, a laboratory or small-production furnace for thick film circuits.

A range of digital voltmeters, multimeters, curve tracers and other small instruments was shown by Fairchild Instrumentation as well as an automatic test system for integrated circuits and modules. The system's test sequence is programmed on magnetic disks with the option of paper tape or direct computer control. Selection of test sequence is done on a simple keyboard with read-out via GO/NO-GO direct digital display or data logging.

British Industries

For the first time a working demonstration, of how a specially-programmed computer can solve complex electronic design problems, was arranged by Racal Research Ltd. Visitors were able to have their circuit design and analysis programs processed via data link to the Racal computer at Tewkesbury within minutes.

The English Electric Group was the largest single exhibitor at the show with exhibits drawn from its six divisions. A wide range of activities from industrial and military automation, closed circuit television, microcircuits to electronic valves used in astronomy and surgery was demonstrated.

The Marconi Company featured a working demonstration of computer graphics showing how the engineer of the future will be able to reduce dramatically the time he takes to design equipment, buildings, etc. The way computers can assist operators to control or supervise complex networks such as rail routes and electricity distribution systems was also demonstrated.

For the design automation demonstration, a complete *Myriad II* computer was installed together with a new graphical/tabular display system. Several alternative input systems, including a light pen, was used to feed information into the computer via the display.

A closed-circuit television exhibit by Marconi showed two vital industrial applications of television, in training employees and in monitoring industrial processes. The uses of television for such applications as helicopter surveillance, missile guidance and other military applications were also illustrated. Marconi's Line Communications Division showed a model of their message switching system, MARS, which provides computer control of a telegraph centre, routing data and telegraph traffic automatically.

A completely new technique for analogue to digital conversion, shown by a display featuring data acquisition, data transmission via a telemetry link and output to a data logging facility, was shown by Smiths Industries. Data acquisition from a number of different signal generators was demonstrated. In the case of those generating d.c. analogue signals, these were fed into a voltage to frequency converter. The output of the converter is a square-wave signal, the frequency of which is proportional to the input voltage. This, and signals already in digital form, are fed into the telemetry transmitter unit. The output from the transmitter is fed into a receiver via a two-wire transmission line, and finally into a data logger featuring both print-out and tape-punch facilities.

The range of remote control and telemetry equipment on show included an example of the G.E.C.–A.E.I. Electronics Mosaic Diagram System incorporating a wide selection of mimic components, all based on the 1-inch mosaic tile, show Teleducer, Teleshift and Telecode systems which transmit over a pair of wires indications of remote measurement, interogatory and metering signals and quantitive information were also demonstrated.

In the radio-communications field, this organization showed the latest Lincompex development that greatly increases the efficiency of existing international h.f. radio links, h.f. synthesized receiver and h.f. frequency synthesizer. V.h.f. equipment included the full range of personal, mobile and portable radio telephones.

The Ferranti Electronics Department exhibited u.h.f. power transistors, strain gauges, step-recovery diodes, plastic transistors, silicon solar cells and photocells. The latter include blue-sensitive, radiation-resistant, n-on-p-type solar cells specially developed for electric power generation on space vehicles, and high-speed miniature photocells for instrumentation purposes.

A number of demonstrations were presented to illustrate the activities of the Automation Division. Illustrated by means of a c.r.t. display unit was the *Argus* computer in control applications taking examples from the power generation, chemical, steel and petro-chemical industries, and the distribution of utilities.

A software package, CONSUL (CONtrol SUb-routine Language), developed to meet a growing need for a fast and flexible method of programming for on-line control systems, was presented. Two c.r.t. displays, linked to a CONSUL control panel, showed how CONSUL can be used as a powerful tool in process engineering.

The Ferranti terminal equipment for B.O.A.C.'s BOADICEA system was also demonstrated. The progress of a seat reservation on a scheduled flight was shown from the initial enquiry to a confirmed booking. Although this facility is only a small part of B.O.A.C.'s plan for the use of a central computer complex with computer-controlled terminal displays, it is expected to provide a great improvement in this sphere alone. Further planned applications for BOADICEA include cargo control, aircraft loading plans and the maintenance of records.

The Plessey Components Group and Plessey Dynamics Group featured a wide range of products. Of great interest was the ranges of opto-electronic devices; silicon integrated circuits and s.i.c. customer design service; thin film circuits; memory stacks and planes; and fluidic devices.

Important advances in opto-electronics were in working exhibits of two new products from the Chemical and Metallurgical Division. The OPT 1 is the first of a range of integrating light detectors combining planar photo-diodes and MOST circuitry on a single silicon chip, and a much higher sensitivity than is usual with photoelectric devices is claimed. The OPT 5 image detector is an extension of this technology, and has a 10×10 array of light sensitive areas. Scanning circuitry using digital techniques is formed around the periphery of the chip, giving a video waveform output.

Another product shown was the type TEO 1 gallium arsenide diode, which utilizes the Gunn effect for microwave generation. The TEO 1 oscillator is encapsulated in a standard S4 varactor pill package and has an output of typically 5 mW c.w. in the range 7–12 GHz; 20 mW c.w. is obtainable if it is used in a limited frequency range.

The latest Mullard developments for instrumentation and control ranged from a vibrating capacitor that can measure currents as low as 10^{-17} A to inexpensive replacements for standard cells.

Used for the measurement of extremely small voltages and currents is the vibrating capacitor type XL7900. This novel capacitor contains a membrane that can vibrate at high frequencies, and offers a new means of measuring small voltages and currents; electrometers using it have measured currents as small as 8×10^{-17} A, or 500 electrons per second! The XL7900 contains four metal plates in parallel, the two outer being fixed. The two inner plates are mechanically linked so that the distance between them is constant; they can, however, vibrate as a rigid pair with respect to the outer plates. This movement is achieved by applying an alternating voltage between one of the vibratory plates and a fixed plate. The d.c. supply is connected to the other two plates, one fixed and one vibratory. Consequently, when the middle plates vibrate, the capacitance across the d.c. supply changes and modulates the direct voltage at the frequency of the vibration. The alternating voltage so produced is easily amplified to produce a signal proportional to the current.

Developments, both in solid-state and thermionic devices, for the telecommunications, radar and industrial heating industries shown included the following: a range of electronically-tuned Gunn oscillators, capable of providing continuous outputs of 5, 10, and 15 mW at X-band; a broadband pre-focused travelling wave tube only 29·5 cm long; a range of ceramic capacitors for use in high-power transmitters and microwave heating equipment; new magnetrons for use in lightweight, compact, airborne radar equipment and high-power, ground radar installations.

Magnetic recording techniques have been developed by Recording Designs for use in torpedoes to determine the various performance parameters. The extension of similar devices to a 'civil' torpedo for oceanographic measurements was also shown. In both applications the torpedo is recovered and the multi-channel recordings are removed for playback and analysis.

The Exhibition was sponsored by five trade associations: the British Electrical and Allied Manufacturers' Association, the British Industrial Measuring and Control Apparatus Manufacturing Association, the Electronic Engineering Association, the Radio and Electronic Component Manufacturers Federation and the Scientific Instrument Manufacturers Association.

U.D.C. 621.372.832.8

Adjustment of the 4-port Single-junction Circulator

By

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AND

C. R. BUFFLER, Ph.D.‡

Summary: The 4-port junction circulator differs from the 3-port junction in that the circulation adjustment cannot be reduced to an admittance matching procedure. To adjust the 4-port junction circulator requires setting three independent variables. A unique procedure is given for setting the three independent variables one at a time in terms of the scattering coefficients of the junction. The procedure is demonstrated on an experimental waveguide junction.

1. Introduction

The 4-port junction circulator consists of a symmetrical distribution of ferrite material at the junction of four transmission lines. Waveguide and stripline versions have been described.[1-4] A schematic of the circulator is shown in Fig. 1. In this device power entering port 1 emerges from port 2, and so on in a cyclic manner. When the junction is matched, it is found to exhibit some of the properties of a transmission cavity resonator between ports 1 and 2. Also, a definite standing-wave exists within the junction with nulls at ports 3 and 4.

An important difference between the 3-port and 4-port junction circulators is that the 4-port circulator cannot be adjusted with external tuning elements only. Unlike the 3-port junction, a 4-port junction can be matched without being a circulator. Hence, the adjustment procedure for the 4-port circulator cannot be reduced solely to an admittance problem. To adjust the 4-port junction requires three independent variables.[5] This makes the tuning procedure quite difficult. The approach to date has usually been one of trial and error.

The 4-port junction circulator was first described by the coefficients and the eigenvalues of the scattering matrix when the junction was operating as a circulator.[5] A phenomenological description in terms of the modes of the junction has also been given in the case of the stripline circulator.[6] However, although this is well understood, the adjustment of the modes has not been related to measured quantities at the external terminals of the junction. It is the purpose of this paper to do this by extending the matrix approach to obtain the scattering coefficients and the eigenvalues associated with each mode adjustment.

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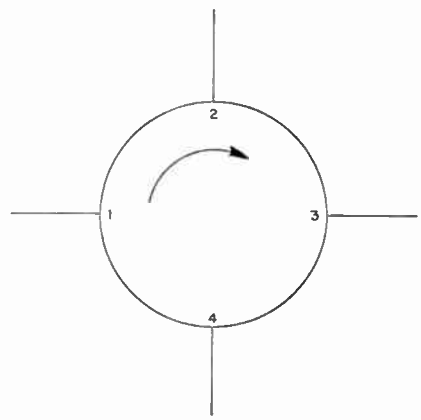


Fig. 1. Schematic of 4-port junction circulator.

This will allow the junction to be matched in three very simple steps.

The experimental work described was carried out on a waveguide junction.

2. Scattering Matrix of 4-port Junction Circulator

The scattering matrix of the 4-port circulator relates the amplitude of the out-going waves to the amplitudes of the incident waves of the junction. The elements along the main diagonal of the scattering matrix are reflection coefficients, and the other elements are transmission coefficients. The principal of the conservation of energy requires that the scattering matrix be unitary. For a 4-port junction the scattering matrix has the form

$$S = \begin{vmatrix} S_{11} & S_{12} & S_{13} & S_{14} \\ S_{21} & S_{22} & S_{23} & S_{24} \\ S_{31} & S_{32} & S_{33} & S_{34} \\ S_{41} & S_{42} & S_{43} & S_{44} \end{vmatrix}$$

For a matched circulator the elements on the main diagonal must be zero. The scattering coefficients of a lossless 4-port junction are given as a function of the eigenvalues, s_0, s_1, s_{-1} and s_2 by the following equations:[5]

$$S_{11} = S_{22} = S_{33} = S_{44} = \frac{s_0 + s_1 + s_{-1} + s_2}{4}$$

$$S_{12} = S_{23} = S_{34} = S_{41} = \frac{s_0 + js_1 - js_{-1} - s_2}{4}$$

$$S_{13} = S_{31} = S_{24} = S_{42} = \frac{s_0 - s_1 - s_{-1} + s_2}{4}$$

$$S_{14} = S_{43} = S_{32} = S_{21} = \frac{s_0 - js_1 + js_{-1} - s_2}{4}$$

The field distributions corresponding to the eigenvectors are known as the modes of the junction. These are introduced in the next Section in connection with the phenomenological discussion.

The eigenvalues of the junction operating as a circulator can be obtained from the characteristic equation using the scattering matrix of an ideal circulator, $S_{12} = 1$, $S_{11} = S_{13} = S_{14} = 0$. The eigenvalues of a matched 4-port circulator are shown in Fig. 2. This requires setting the phases of three of the eigenvalues with respect to the fourth, and so three independent variables are required. A shortcoming of this analysis, however, is that it does not give the steps required to obtain this eigenvalue arrangement.

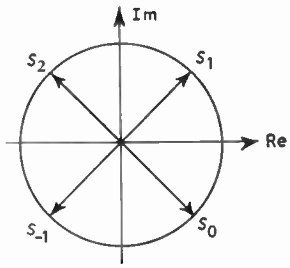
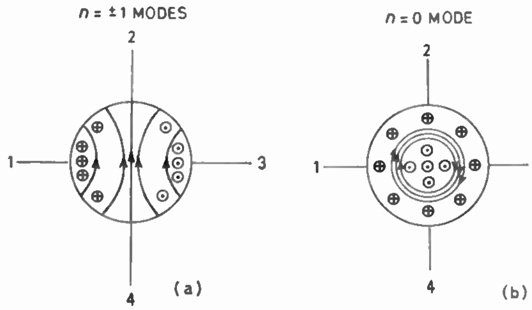
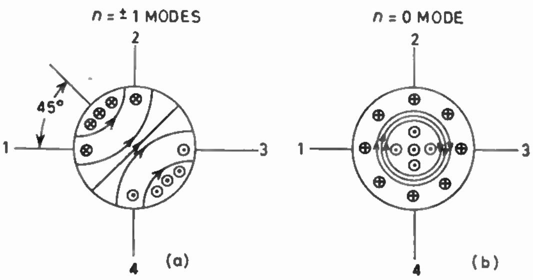


Fig. 2. Eigenvalues of a matched 4-port junction circulator (after Auld[5]).



(a) $n = \pm 1$.   (b) $n = 0$.

Fig. 3. Field patterns for unmagnetized disk.



(a) $n = \pm 1$.   (b) $n = 0$.

Fig. 4. Four-port circulator synthesized from $n = \pm 1$ and $n = 0$ modes (after Fay and Comstock[6]).

3. Theory of the 4-port Junction Circulator

In terms of the phenomenological description, the junction geometry has to support a pair of degenerate resonant modes to form a standing-wave pattern within the junction at the operating frequency. This is the same condition as for the 3-port junction circulator.[7] An additional mode is also required which is resonant at the same frequency. The standing-wave pattern for the degenerate pair of $n = \pm 1$ modes is shown in Fig. 3(a), for which the electric field configuration is seen to form a figure of eight pattern. The $n = 0$ mode pattern is shown in Fig. 3(b). The degenerate pair of $n = \pm 1$ modes are then split by magnetizing the junction until the standing-wave pattern is rotated through 45°. The rotated standing-wave pattern is shown in Fig. 4(a), and the $n = 0$ field pattern is also reproduced in Fig. 4(b). If the magnitudes of the E-fields for the two patterns are equal at the 4-ports, then we have circulation between ports 1 and 2 and ports 3 and 4 are isolated.

Although the mode patterns described above were first given in connection with the stripline junction they are assumed to apply phenomenologically to the waveguide version also. It has been shown by Volman[8] that the electric field distribution of the waveguide 3-port junction has a similar figure of eight pattern to the stripline version. The scattering matrices of the two versions are of course the same. Hence, no distinction is made in this paper between the two versions.

4. Circulation Adjustments

In this Section the mode adjustment described previously is related to the coefficient of the scattering matrix. The first circulation adjustment is made by adjusting the diameter of the ferrite disk so that the degenerate $n = \pm 1$ modes are resonant at the operating frequency. When the junction is magnetized the operating frequency then lies between the two split modes. This condition can be determined from the electric field pattern in the ferrite disk which requires that $S_{13} = 1$ and $S_{11} = S_{12} = S_{14} = 0$. The corresponding eigenvalues can be obtained from the relations in Section 2. The second circulation condition is satisfied by making the $n = 0$ mode resonate at the same frequency as the $n = \pm 1$ modes. Since the $n = 0$ mode is initially tuned to a higher frequency some means is required to tune this mode to the frequency of the $n = \pm 1$ modes without disturbing the latter modes. One way of doing this is to introduce a thin pin through the centre of the junction as shown in Fig. 5 which capacitively loads the $n = 0$ mode but leaves the $n = \pm 1$ modes undisturbed.[3,5,6] This modifies the eigenvalues of the junction with the result that the scattering coefficients become $S_{11} = S_{12} = S_{13} = S_{14} = 0{\cdot}5$. This last condition is obtained from the phase relation of the eigenvalues associated with the first and third circulation adjustment. The third circulation condition is now immediately satisfied by symmetrically removing the degeneracy between the $n = \pm 1$ modes until the standing wave pattern is rotated through 45°. This last setting removes the degeneracy between the eigenvalues s_1, s_{-1} and results in $S_{12} = 1$, $S_{11} = S_{13} = S_{14} = 0$.

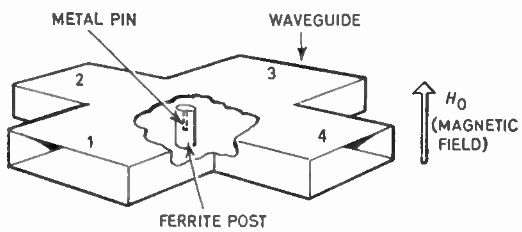


Fig. 5. Four-port waveguide single junction circulator.

If the splitting is not symmetrical the eigenvalues will not be in the position required for circulation. A small adjustment of the ferrite diameter and metal pin may then be necessary. However an essential feature of the circulator adjustment is that the splitting is very small and essentially symmetrical to first-order. The experimental work described shows that no such adjustment was necessary. It should also be noted that the $n = 0$ mode cannot be split by the magnetic field.

5. Adjustment of $n = \pm 1$ Modes

The first circulation adjustment is met by setting the diameter of the ferrite disk so that the degenerate $n = \pm 1$ modes are resonant at the operating frequency. This condition can be obtained experimentally by noting that the electric field pattern in Fig. 3(a) forms a figure of eight in the vicinity of the disk. This means that when the junction resonance is satisfied, ports 2 and 4 are decoupled, and transmission should occur between ports 1 and 3. In other words, $S_{13} = 1$ and $S_{11} = S_{12} = S_{14} = 0$. The eigenvalues corresponding to this situation are

$$s_2 = s_0$$
$$s_1 = s_{-1} = -s_0$$

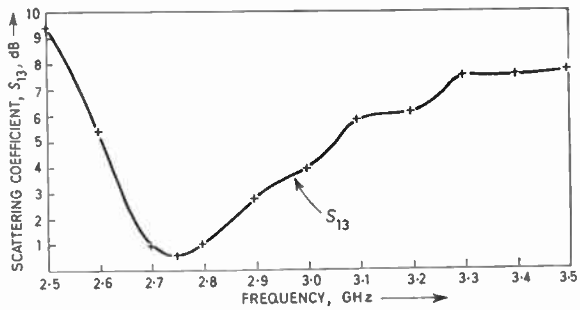


Fig. 6. Scattering coefficient, S_{13}, versus frequency (diameter of garnet ring = 1·96 cm).

For experimental convenience the diameter of the ferrite disk was kept constant and the frequency was varied instead of the more usual situation in which the frequency is fixed and the ferrite diameter is varied. The transmission coefficient S_{13} is shown in Fig. 6 as a function of frequency for the waveguide junction.

The transmission Q associated with this last figure is fairly large, which is not desirable for a broadband junction. In one experiment, the transmission Q was substantially reduced by replacing the garnet rod by two garnet disks.

The material used in the experimental work was a modified yttrium iron garnet with $M_s/\mu_0 = 55000\,\text{AT/m}$ and $\varepsilon_r = 14{\cdot}4$.

6. Adjustment of $n = 0$ Mode

Once the $n = \pm 1$ modes have been adjusted, the $n = 0$ mode is tuned by adjusting the length of the metal pin through the centre of the junction. When the isotropic resonances of the junction are satisfied then we have

$$s_2 = -s_0$$

and

$$s_1 = s_{-1} = s_0$$

With the help of the above relations, the scattering coefficients are given by $S_{11} = S_{12} = S_{13} = S_{14} = 0{\cdot}5$. The isotropic resonances are satisfied when this condition is met. The variation of the scattering

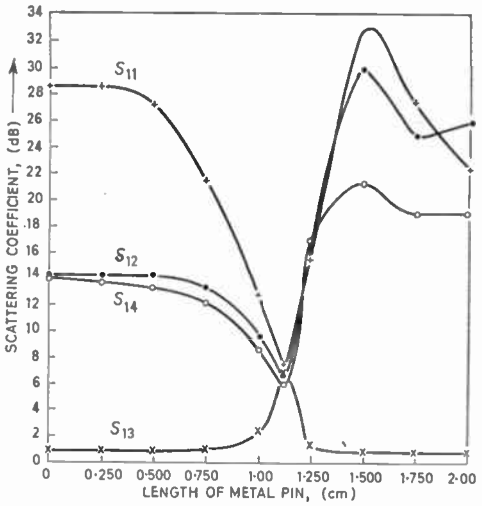


Fig. 7. Scattering coefficients versus length of metal pin (frequency = 2·74 GHz; diameter of garnet ring = 1·96 cm).

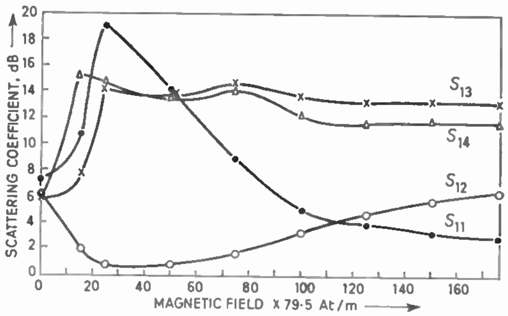


Fig. 8. Scattering coefficients versus magnetic field at 2·74 GHz (length of pin = 1·1 cm; diameter of garnet ring = 1·96 cm).

coefficients versus the length of the metal pin is shown in Fig. 7.

It can be seen that the $n = 0$ mode is extremely sensitive to the length of the metal pin. A more appropriate tuning arrangement might consist of varying the dielectric constant of a dielectric rod at the centre of the garnet disk.

The final adjustment is now met by removing the degeneracy between the $n = \pm 1$ modes.

7. Splitting of the Degenerate $n = \pm 1$ Modes

Once the isotropic resonances have been adjusted we can now magnetize the junction, thereby removing the degeneracy of the $n = \pm 1$ modes to rotate the standing-wave pattern formed by the latter modes through 45°.

The variation of the scattering coefficients with magnetic field at 2·74 GHz is shown in Fig. 8. The junction is directly matched with a magnetic field of 2780 AT/m. We then have $S_{12} = 1$ and $S_{11} = S_{13} = S_{14} = 0$. The eigenvalues are shown in Fig. 2. In obtaining the last result no intermediate adjustments were made besides those described in Sections 5 and 6.

8. Conclusions

In this paper an adjustment procedure has been described which allows the three independent variables of the 4-port junction to be set independently. An experimental junction was developed using the procedure outlined. The approach described should be helpful in optimizing each of the independent variables to make the junction capable of handling large bandwidths.

9. Acknowledgment

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An Invariant Alignability Factor and Its Significance

By

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Summary: The sensitivities of the port immittance to small immittance and large reactive changes of the self-immittance at the opposite port of a linear active two-port network are investigated. A new 'invariant alignability factor', δ, is defined as the inverse of the maximum modulus of the sensitivity. δ is an invariant for interchange of the ports, arbitrary lossless terminations and immittance substitution. It precisely expresses the maximum interaction between the two ports and is related to earlier stability and performance factors. δ can be evaluated by a simple experimental method.

The power gain margin of any active linear two-port network is δ nepers, when δ is very small compared with unity. The upper limit of its maximum power gain for a given performance factor is inversely proportional to the square of δ.

List of Symbols

g_1	loop gain of two-port network, $\frac{p_{12}p_{21}}{p_1 p_2}$ (with source and load terminations)
k	Stern's stability factor of two-port network $\left(= \frac{2\rho_1\rho_2}{L+M}\right)$
K	sensitivity of the total port immittance to small immittance changes in the total self-immitance at the opposite port
K'	sensitivity as above but for large reactive changes
K_i and K_i'	'inherent sensitivity' or sensitivity for small immittance changes and large reactive changes respectively, when the real parts of source and load immittances vanish
L	$= \lvert p_{12}p_{21}\rvert$
M	$= \mathrm{Re}\,(p_{12}p_{21})$
η	performance factor of two-port network $\left(\text{equals the inverse of } \lvert g_1\rvert_{\max} \text{ or } \frac{\rho_1\rho_2}{L}\right)$
N	$= \mathrm{Im}\,(p_{12}p_{21})$
$\begin{bmatrix} p_{11} & p_{12} \\ p_{21} & p_{22} \end{bmatrix}$	two-port matrix of network without source and load immittances, where $p = h, z, y$ or g
p_S and p_L	source and load immittances of two-port network
p_1 and p_2	total self-immittance at input and output ports ($= p_{11}+p_S$ and $p_{22}+p_L$ respectively)
p_{p1} and p_{p2}	total port immittance at input and output ports $\left(= p_1 - \frac{p_{12}p_{21}}{p_2} \text{ and } p_2 - \frac{p_{12}p_{21}}{p_1} \text{ respectively}\right)$
s	invariant stability factor of two-port network ($= \eta+\sqrt{\eta^2-1}$ or the inverse of the modulus of loop gain with conjugate matched terminations, if $s \geqslant 1$)
Δp_{p1}	a change in p_{p1} due to a small immittance change or a large reactive change in p_2
Δp_{p2}	a change in p_{p2} due to a small immittance change or a large reactive change in p_1
δ	'invariant alignability factor' of two-port network $\left(= \frac{\eta-1}{2} \text{ or the inverse of the maximum modulus of the sensitivity}\right)$
δ_i	'invariant inherent alignability factor' of two-port network $\left(= \frac{\eta_i-1}{2} \text{ or the inverse of the maximum modulus of the inherent sensitivity}\right)$

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η invariant factor of two-port network $\left(=\dfrac{2\rho_1\rho_2-M}{L}\right)$

η_i inherent invariant factor of two-port network $\left(=\dfrac{2\rho_{11}\rho_{22}-M}{L}\right)$

θ = arg $(p_{12}p_{21})$

Π a normalizing factor $\left(=\dfrac{L+M}{2L}\right)$

ρ = Re (p)

σ = Im (p)

1. Introduction

A given two-port network can be terminated at its ports by 'elements'—namely resistance and reactance—in series or in shunt combinations giving rise to four 'matrix environments'. Figure 1 illustrates these environments. It is then convenient to characterize the network by that set of matrix parameters p (equals h, z, y or g) which corresponds with the environment of the network. For example, if elements are in shunt at both the input and output ports as in Fig. 1(c), it is convenient to express the network by its y- or admittance parameters.

Characterize the linear two-port network with passive source immittance, p_S, and passive load immittance, p_L, by its general p-matrix, where

$$[p]=\begin{bmatrix}p_{11}+p_S & p_{12}\\ p_{21} & p_{22}+p_L\end{bmatrix}=\begin{bmatrix}p_1 & p_{12}\\ p_{21} & p_2\end{bmatrix} \quad ...(1)$$

In eqn. (1), p_1, etc., may be sets of h-, z-, y- or g-matrix parameters.

Let

$$p_1=\rho_1+j\sigma_1 \quad \text{etc.} \quad ......(2)$$

$$p_{12}p_{21}=M+jN;\quad |p_{12}p_{21}|=L \quad ......(3)$$

The total port immittance, p_{p1} and p_{p2} at ports 1 and 2 respectively are given by

$$p_{p1}=p_1-\frac{p_{12}p_{21}}{p_2};\quad p_{p2}=p_2-\frac{p_{12}p_{21}}{p_1} \quad ...(4)$$

The small change in the total input port immittance due to a small change in the total output self-immittance, say, due to a change in the load immittance, is given by

$$\Delta p_{p1}=\frac{p_{12}p_{21}}{p_2^2}\Delta p_2 \quad ......(5)$$

Equation (5) can be rearranged[1] as

$$\left|\frac{\Delta p_{p1}}{p_1}\right|\leqslant\frac{1}{n}\left|\frac{\Delta p_2}{p_2}\right| \quad ......(6)$$

where n is the performance factor[1-5] of the active two-port network and is also the inverse of the maximum modulus of loop gain.

$$n=\frac{1}{|g_l|_{max}}=\frac{\rho_1\rho_2}{L} \quad ......(7)$$

Similarly,

$$\left|\frac{\Delta p_{p2}}{p_2}\right|\leqslant\frac{1}{n}\left|\frac{\Delta p_1}{p_1}\right| \quad ......(8)$$

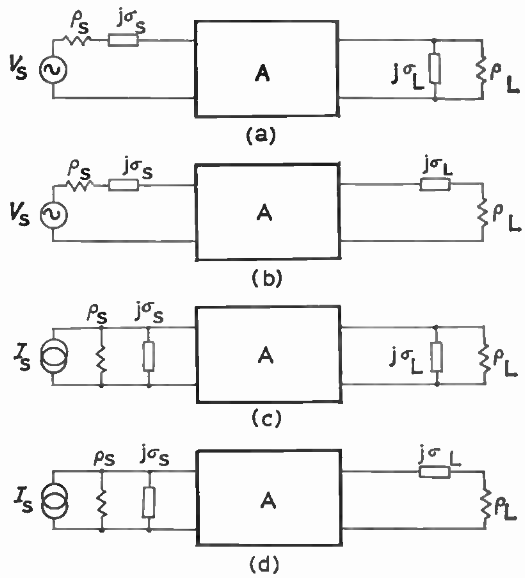

Fig. 1. Possible matrix environments for a two-port active network, A.

(a) h-environment with series input and shunt output terminations (source impedance, $\rho_S+j\sigma_S$; load admittance, $\rho_L+j\sigma_L$).
(b) z-environment with series input and series output terminations (source impedance, $\rho_S+j\sigma_S$; load impedance $\rho_L+j\sigma_L$).
(c) y-environment with shunt input and shunt output terminations (source admittance, $\rho_S+j\sigma_S$; load admittance, $\rho_L+j\sigma_L$).
(d) g-environment with shunt input and series output terminations (source admittance, $\rho_S+j\sigma_S$; load impedance, $\rho_L+j\sigma_L$).

In eqns. (6) and (8), the normalizations of Δp_{p1} and Δp_{p2} are not with respect to p_{p1} and p_{p2} respectively. Therefore, the inverse of n is not an exact measure of the maximum 'sensitivity'[6] of the total port immittance to changes in the total self immittance at the opposite port.

The question arises as to whether the maximum possible modulus of the sensitivity can be determined and if so, whether it is an invariant in matrix environments,[2] under arbitrary lossless terminations and for interchange of input and output ports. An affirmative answer is provided in the following sections. A new 'invariant alignability factor', δ, is obtained and is related to earlier stability based factors, including the performance factor. Some applications of δ to mismatched amplifier stages are considered.

2. Sensitivity to Small Immittance Changes

Define sensitivity, K_{12} of total input port immittance for small immittance changes in the total output self-immittance by

$$K_{12} = \frac{\partial(\ln p_{p1})}{\partial(\ln p_2)} = \frac{\Delta p_{p1}}{p_{p1}} \cdot \frac{p_2}{\Delta p_2} \quad \text{......(9)}$$

Similarly define sensitivity, K_{21}, of total output port immittance for small immittance changes in the total input self-immittance.

From eqns. (4), (5) and (9) and above sentence,

$$K_{12} = K_{21} = K = \frac{p_{12}p_{21}}{p_1 p_2 - p_{12}p_{21}} \quad \text{......(10)}$$

Thus the sensitivity, K, is an invariant for interchange of input and output ports.

From eqn. (10), the maximum modulus can be written as

$$|K|_{\max} = \frac{L}{|p_1 p_2 - p_{12}p_{21}|_{\min}} \quad \text{......(11)}$$

since changes in p_S, p_L alter p_1, p_2 but not the product $p_{12}p_{21}$.

From eqn. (4),

$$|p_1 p_2 - p_{12}p_{21}|_{\min} \geqslant |p_2|_{\min}|p_{p1}|_{\min}$$

or

$$\geqslant |p_1|_{\min}|p_{p2}|_{\min} \quad \text{......(12)}$$

The minimum of $|p_2|$ and the minimum of $|p_{p1}|$ are usually obtained for different values of σ_2, the imaginary part of p_2. For the special case, where these values coincide (see Section 2.1), the equality sign holds in statement (12). Similarly for the minimum of $|p_1|$ and the minimum of $|p_{p2}|$.

The minimum of $|p_2|$ equals ρ_2 and occurs when σ_2 vanishes. The minimum modulus of p_{p1} is investigated in the Appendix. Thus

$$|p_{p1}|_{\min} = \frac{2\rho_1\rho_2 - M - L}{2\rho_2} \quad \text{......(13)}$$

and occurs when

$$\sigma_1 = \frac{N}{2\rho_2} \quad \text{and} \quad \sigma_2 = \frac{N}{L+M}\rho_2 \quad \text{......(14)}$$

From eqns. (11) to (13) and the above sentence,

$$|K|_{\max} \leqslant \frac{1}{\delta}, \quad \text{where } \delta = \frac{\eta - 1}{2} \quad \text{......(15)}$$

and the invariant factor,[3,4,7] η, is given by

$$\eta = \frac{2\rho_1\rho_2 - M}{L} \quad \text{......(16)}$$

η and hence δ is an invariant in matrix environments[2] or for immittance substitution,[7] under arbitrary lossless terminations and for interchange of input and output ports. Therefore, the factor, δ, is defined as the 'invariant alignability factor'.

From eqns. (9), (10) and statement (15),

$$\left|\frac{\Delta p_{p1}}{p_{p1}}\right| \leqslant \frac{1}{\delta}\left|\frac{\Delta p_2}{p_2}\right| \quad \text{......(17)}$$

$$\left|\frac{\Delta p_{p2}}{p_{p2}}\right| \leqslant \frac{1}{\delta}\left|\frac{\Delta p_1}{p_1}\right| \quad \text{......(18)}$$

Equations (17) and (18) are true sensitivity statements unlike eqns. (6) and (8). It is of interest to find out when the equality sign holds for statements (17) and (18).

2.1. *Sensitivity Modulus Maximum when $p_{12}p_{21}$ is Real and Positive*

Let the square of the modulus of the determinant, $p_1p_2 - p_{12}p_{21}$, be denoted by D. Thus,

$$D = (\rho_1\rho_2 - \sigma_1\sigma_2 - M)^2 + (\rho_1\sigma_2 + \rho_2\sigma_1 - N)^2 \quad (19)$$

D is a maximum or minimum[2,8] with respect to the variables σ_1 and σ_2 only when

$$\frac{\sigma_1}{\rho_1} = \frac{\sigma_2}{\rho_2} = \lambda \quad \text{......(20)}$$

and

$$\lambda^3 + \lambda\left(1 + \frac{M}{\rho_1\rho_2}\right) - \frac{N}{\rho_1\rho_2} = 0 \quad \text{......(21)}$$

If $p_{12}p_{21}$ is real and positive, $M = L$ and $N = 0$. Then, the only real root, λ_0, equals zero and from eqn. (20), σ_1 and σ_2 both vanish. Hence eqn. (19) gives

$$D_{\min} = (\rho_1\rho_2 - M)^2 \quad \text{......(22)}$$

From eqns. (10), (19) and (22)

$$|K|_{\max} = \frac{L}{\sqrt{D_{\min}}} = \frac{1}{\delta} \quad \text{......(23)}$$

because $M = L$.

The terminations σ_1 and σ_2 for this case can also be obtained from eqn. (14) by substituting zero for N. For this situation, the upper bound for the modulus of sensitivity equals the inverse of the alignability factor, δ.

So far, we have considered the sensitivity for small immittance changes only. It is now proposed to investigate the sensitivity for large reactive changes.

3. Sensitivity to Large Reactive Changes

Define an input sensitivity, K'_{12}, for large reactive changes at output port by

$$K'_{12} = \frac{p'_{p1}-p_{p1}}{p_{p1}}\frac{p_2}{p'_2-p_2} = \frac{\Delta p_{p1}}{p_{p1}}\frac{p_2}{\Delta p_2} \quad \text{......(24)}$$

Here, the input port immittance has an initial value p_{p1} and a final value p'_{p1} due to a large reactive change in the output self-immittance from p_2 to p'_2. Thus eqns. (9) and (24) are similar in form but not in detail; changes in eqn. (9) refer to small immittances, whereas in eqn. (24), large reactive changes are considered.

Similarly an output sensitivity, K'_{21}, for large reactive changes at input port can be defined.

From eqns. (4) and (24)

$$|K'_{12}|_{\max} = \frac{L}{|p'_2 p_{p1}|_{\min}} \quad \text{......(25)}$$

$|p_{p1}|_{\min}$ is given by eqn. (13) and occurs when condition (14) is satisfied. $|p'_2|_{\min}$ occurs when the final value of the imaginary part of p_2, i.e. σ'_2, vanishes. Thus, as in eqn. (14), the initial value of σ_2 can be $N\rho_2/(L+M)$ and its final value σ'_2 can be zero so that

$$|K'_{12}|_{\max} = \frac{2L}{2\rho_1\rho_2-M-L} = \frac{1}{\delta} \quad \text{......(26)}$$

Since δ is reciprocal, it follows that

$$|K'_{12}|_{\max} = |K'_{21}|_{\max} = |K'|_{\max}, \quad \text{say.}$$

Thus, the maximum modulus of the sensitivity of the total port immittance to large reactive changes of the total self-immittance at the opposite port equals the inverse of the alignability factor, δ.

4. Invariant Alignability Factor

The invariant alignability factor, δ, has been defined for small immittance changes by eqn. (9) and statement (15) while for large reactive changes it has been defined by eqns. (24) and (26). A particular alignability factor, called the 'invariant inherent alignability factor', δ_i, results when the real parts of the source and load immittances vanish, i.e. with ρ_S and ρ_L both equal to zero.

$$\delta_i = \frac{\eta_i - 1}{2} \quad \text{......(27)}$$

where the inherent invariant factor,

$$\eta_i = \frac{2\rho_{11}\rho_{22}-M}{L} \quad \text{......(28)}$$

The new alignability factor, δ, can be compared with Gibbons' alignability factor,[9]

$$\delta_G = \frac{\Delta p_{in}}{p_{in}}\frac{p_L}{\Delta p_L} \quad \text{......(29)}$$

where

$$p_{in} = p_{11} - \frac{p_{12}p_{21}}{p_2} \quad \text{......(30)}$$

so that

$$\delta_G = \frac{p_{12}p_{21}}{p_{11}p_2 - p_{12}p_{21}}\frac{p_L}{p_2} \quad \text{......(31)}$$

Gibbons' alignability factor, δ_G, is a particular sensitivity measure and is generally complex in value. It does not express the upper bound of sensitivity of port immittance for small self-immittance changes or reactive changes at the opposite port. δ_G is a variant for interchange of ports, arbitrary lossless terminations and for immittance substitution.

The invariant alignability factor, δ, as already shown, is related to the invariant factor, η, which is a stability-based factor. Other stability-based factors are the performance factor, n, of eqn. (7), Stern's stability factor,[8,10] k and the invariant stability factor,[3,4] s, both of which are defined in the list of symbols. For convenience and ready reference all these factors are shown in Table 1. Similar relationships exist between the various inherent factors, i.e. when $\rho_S = \rho_L = 0$.

The two-port network with given ρ_S and ρ_L is absolutely stable or stable for all reactive or further passive terminations, provided

$$s \geqslant 1 \text{ or } \eta \geqslant 1 \text{ or } \delta \geqslant 0 \text{ or } n \geqslant \Pi \text{ or } k \geqslant 1 \quad \text{......(32)}$$

where

$$\Pi = \frac{L+M}{2L} \quad \text{......(33)}$$

The value of the invariant alignability factor, δ, for a given two-port network can be obtained by a simple experiment. The circle of p_{p1} can be drawn[11-13] from three measurements of p_{p1} for three convenient values of reactive load termination. The centre of this circle is (a_1, b_1) and its radius is c_1 where[13]

$$a_1 = \frac{\eta L}{2\rho_2}$$

$$b_1 = \frac{2\sigma_1\rho_2 - N}{2\rho_2}$$

and

$$c_1 = \frac{L}{2\rho_2} \quad \text{......(34)}$$

so that

$$\delta = \frac{\eta-1}{2} = \frac{a_1-c_1}{2c_1} \quad \text{......(35)}$$

Table 1

Interrelations between k, n, η, s and δ. Here $p_{12}p_{21} = M+jN$; $|p_{12}p_{21}| = L$; $\Pi = \dfrac{L+M}{2L}$.

From→ To ↓	k	n	η	s	δ
k	k	$\dfrac{n}{\Pi}$	$\dfrac{\eta L + M}{L+M}$	$1 + \dfrac{(s-1)^2}{4s\Pi}$	$\dfrac{\delta}{\Pi} + 1$
n	$k\Pi$	n	$\dfrac{\eta L + M}{2L}$	$\Pi + \dfrac{(s-1)^2}{4s}$	$\delta + \Pi$
η	$k + (k-1)\dfrac{M}{L}$	$2n - \dfrac{M}{L}$	η	$\dfrac{s^2+1}{2s}$	$2\delta + 1$
s	$k + (k-1)\dfrac{M}{L} + \sqrt{\left[k + (k-1)\dfrac{M}{L}\right]^2 - 1}$	$2n - \dfrac{M}{L} + \sqrt{\left(2n - \dfrac{M}{L}\right)^2 - 1}$	$\eta + \sqrt{\eta^2 - 1}$	s	$2\delta + 1 + 2\sqrt{\delta(\delta+1)}$
δ	$(k-1)\Pi$	$n - \Pi$	$\dfrac{\eta - 1}{2}$	$\dfrac{(s-1)^2}{4s}$	δ

5. Applications of δ to Mismatched Active Two-port Networks

5.1. *Maximum Power Gain as δ Tends to Zero*

The maximum power gain of an active two-port network was evaluated[2] for k tending to unity or the present δ tending to zero, with the stipulation that $\Pi > 1/9$. This ensures that the cubic eqn. (21) has only one real root for all values of $k \geqslant 1$ (or $\delta \geqslant 0$). It is now proposed to evaluate the maximum power gain of the active two-port network as δ tends to zero, when $\Pi < 1/9$, i.e. when there are three real roots for the cubic equation.

Equation (21) can be factorized[2] for $k = 1$ as

$$\left(\lambda - \frac{N}{L+M}\right)\left(\lambda^2 + \frac{N}{L+M}\lambda + 2\right) = 0 \quad \text{...(36)}$$

Let

$$\arg(p_{12}p_{21}) = \theta \quad \text{......(37)}$$

so that

$$\frac{N}{L} = \sin\theta \quad \text{and} \quad \frac{M}{L} = \cos\theta \quad \text{......(38)}$$

The three roots of eqn. (36) are

$$\lambda = \lambda_1 = \frac{\sin\theta}{1+\cos\theta} \quad \text{......(39)}$$

$$\lambda = \lambda_2 = -\frac{\sin\theta}{2(1+\cos\theta)} + \sqrt{-\frac{(7+9\cos\theta)}{4(1+\cos\theta)}} \quad \text{......(40)}$$

$$\lambda = \lambda_3 = -\frac{\sin\theta}{2(1+\cos\theta)} - \sqrt{-\frac{(7+9\cos\theta)}{4(1+\cos\theta)}} \quad \text{......(41)}$$

If $7+9\cos\theta < 0$ or $\Pi[= (1+\cos\theta)/2]$ is $< 1/9$, all the three roots listed above are real. For Stern's $k = 1$ or present $\delta = 0$,

$$\rho_1\rho_2 = \frac{L+M}{2} \quad \text{......(42)}$$

From eqns. (19) to (21) and (42),

$$\frac{4D_0/L^2}{(1+\cos\theta)(1+3\cos\theta)} = \lambda^2 - \frac{6\sin\theta}{1+3\cos\theta}\lambda + \frac{(1-\cos\theta)(5+3\cos\theta)}{(1+\cos\theta)(1+3\cos\theta)} \quad \text{......(43)}$$

D_0 or optimum D can equal zero, only when eqn. (43) equals zero, or

$$\left[\lambda - \frac{\sin\theta}{1+\cos\theta}\right]\left[\lambda - \frac{(5+3\cos\theta)(1-\cos\theta)}{(1+3\cos\theta)\sin\theta}\right] = 0 \quad (44)$$

i.e. when eqn. (39) is satisfied or when

$$\lambda = \lambda_4 = \frac{(5+3\cos\theta)(1-\cos\theta)}{(1+3\cos\theta)\sin\theta} \quad \text{......(45)}$$

λ_4 can equal λ_2 of eqn. (40) or λ_3 of eqn. (41) only if

$$\left[\frac{(5+3\cos\theta)(1-\cos\theta)}{(1+3\cos\theta)\sin\theta} + \frac{\sin\theta}{2(1+\cos\theta)}\right]^2 = -\frac{(7+9\cos\theta)}{4(1+\cos\theta)} \quad \text{......(46)}$$

or when

$$\frac{(\sin^2\theta)(11+9\cos\theta)^2}{(1+\cos\theta)^2(1+3\cos\theta)^2} + \frac{(7+9\cos\theta)}{(1+\cos\theta)} = 0 \quad \text{......(47)}$$

If $\Pi = 0$, corresponding to $\cos\theta = -1$, k cannot tend to unity or δ cannot tend to zero. Similarly, if $\Pi < 1/9$, corresponding to $\cos\theta < -7/9$, there are three real roots of cubic eqn. (21). Hence, for this discussion, neither $(1+\cos\theta)$ nor $(1+3\cos\theta)$ equals zero. Equation (47) now reduces to

$$128(1+\cos\theta)^2 = 0 \qquad \text{......(48)}$$

a condition that cannot hold because δ cannot tend to zero for this situation. Therefore $D_0 = 0$, only for the real root λ_1 of eqn. (39). This is true for all values of Π (other than zero).

The maximum power gain, as k tends to unity or δ tends to zero, is therefore the same as obtained earlier[2] for $\Pi > 1/9$, namely,

$$\underset{k\to 1}{g_{\max k}} \simeq \frac{4k}{(k-1)^2}\left|\frac{p_{21}}{p_{12}}\right|\left\{1-\sqrt{\frac{k_i}{k}}\right\}^2 \qquad \text{......(49)}$$

The terminations required to realize this maximum power gain are[2]

$$\rho_S = \rho_{11}\left(\sqrt{\frac{k}{k_i}}-1\right); \quad \rho_L = \rho_{22}\left(\sqrt{\frac{k}{k_i}}-1\right) \qquad (50)$$

and

$$\sigma_S = \rho_1\frac{(1+k)\sin\theta}{2k(1+\cos\theta)} - \sigma_{11} \qquad \text{......(51)}$$

$$\sigma_L = \rho_2\frac{(1+k)\sin\theta}{2k(1+\cos\theta)} - \sigma_{22} \qquad \text{......(52)}$$

Equation (49) can be rewritten as

$$\underset{\delta\to 0}{g_{\max\delta}} \simeq \frac{4\Pi^2}{\delta^2}\left|\frac{p_{21}}{p_{12}}\right|\left\{1-\sqrt{\frac{\delta_i+\Pi}{\delta+\Pi}}\right\}^2 \qquad \text{......(53)}$$

This follows from the interrelations of Table 1, namely,

$$\delta = (k-1)\Pi = n-\Pi \qquad \text{......(54)}$$

Equations (49) and (53) show that the maximum power gain as k tends to unity or δ tends to zero, is dependent upon the matrix environment selected. This is because k_i and Π vary[2] from environment to environment.

5.2. *Gain Margin as δ Tends to Zero*

As in control systems design, it is useful to consider a gain margin as an indication of the closeness with which the active two-port approaches the condition of instability. The power gain margin is accordingly defined[2]† as the inverse of the real part of the internal loop gain at the frequency of maximum power gain as k tends to unity or δ tends to zero. It is expressed in nepers or decibels.

† This definition is different from the usual one of control systems.

This definition is necessary because the internal loop gain is generally not purely real for the terminations associated with the maximum power gain. It was previously shown[2] that

$$\underset{\substack{k\to 1\\ \Pi>1/9}}{\text{power gain margin}} \simeq (k-1)\Pi \quad \text{nepers}$$
$$\text{or} \simeq 8{\cdot}686(k-1)\Pi \quad \text{dB} \qquad \text{......(55)}$$

The root of the cubic eqn. (21), used to evaluate the loop gain, is λ_1 of eqn. (39). It was established in Sect. 5.1 that the other roots, λ_2, λ_3, even if real, cannot make D tend to zero or make the loop gain tend to unity. Thus eqn. (55) is true for all values of Π and can be rewritten

$$\underset{\delta\to 0}{\text{power gain margin}} \simeq \delta \quad \text{nepers}$$
$$\text{or} \simeq 8{\cdot}686\delta \quad \text{dB} \qquad \text{......(56)}$$

This follows from relation (54).

For a given k close to unity, gain margin is not unique according to eqn. (55) because Π varies[2] from one environment to another. But, for a given δ close to zero, the power gain margin, according to eqn. (56) is an invariant, since δ is invariant in matrix environments, for arbitrary lossless terminations and for interchange of ports. The new alignability factor, δ, controls directly the power gain margin of any active two-port network.

5.3. *Upper Limit for Maximum Power Gain*

The maximum power gain for a given performance factor, n, is given[14] by

$$\underset{n\geqslant\Pi}{g_{\max n}} \simeq 4F\left|\frac{p_{21}}{p_{12}}\right|\left\{1-\sqrt{\frac{n_i}{n}}\right\}^2 \qquad \text{......(57)}$$

where

$$F = \frac{nL^2}{D_{\min}} \qquad \text{......(58)}$$

and

$$D_{\min} = |p_1 p_2 - p_{12}p_{21}|^2_{\min} \qquad \text{......(59)}$$

From eqns. (11), (59) and statement (15),

$$D_{\min} \geqslant \delta^2 L^2 \qquad \text{......(60)}$$

Hence eqns. (57) to (60) yield

$$\underset{n\geqslant\Pi}{g_{\max n}} \leqslant \frac{4n}{\delta^2}\left|\frac{p_{21}}{p_{12}}\right|\left\{1-\sqrt{\frac{n_i}{n}}\right\}^2 \qquad \text{......(61)}$$

$$\text{or} \leqslant \left(\frac{n}{\delta}\right)^2\left[\frac{4}{n}\left|\frac{p_{21}}{p_{12}}\right|\left(1-\sqrt{\frac{n_i}{n}}\right)^2\right] \qquad \text{......(62)}$$

where the terms within the square bracket of statement (62) is the approximate gain expression[1,2,14] derived earlier, namely,

$$\underset{n\gg 1}{g_{\max n}} \simeq \frac{4}{n}\left|\frac{p_{21}}{p_{12}}\right|\left\{1-\sqrt{\frac{n_i}{n}}\right\}^2 \qquad \text{......(63)}$$

When $\Pi = 1$ or $M = L$, Sect. 2.1 shows that equality sign holds in statements (61) and (62) also.

δ equals $n-\Pi$, where Π ranges between 0 and 1, so that the upper limit of gain in statement (62) is higher than the gain of eqn. (63) for a given value of n.

The difference between upper limit of maximum power gain, exact and approximate maximum power gains is best illustrated by an example. Consider the common-emitter amplifier stage in y-matrix environment using transistor type AF 117 that was discussed previously.[14] At an emitter current of 2 mA, collector voltage of -6 V and an ambient temperature of 298°K, the device admittance parameters, including biasing resistances, at a frequency of 455 kHz are[14]

$$y_{11} = (510+j410)10^{-6}\ \Omega^{-1}$$
$$y_{12} = -j8\times 10^{-6}\ \Omega^{-1}$$
$$y_{21} = 70\times 10^{-3}\ \Omega^{-1}$$
$$y_{22} = (2\cdot 5+j14)10^{-6}\ \Omega^{-1}$$

Hence

$$y_{12}y_{21} = 0\cdot 560\times 10^{-6}\underline{/-90^\circ}\ \Omega^{-2}$$
$$n_i = 2\cdot 28\times 10^{-3}$$

and

$$\Pi = 0\cdot 5$$

Table 2 lists these various power gains for performance factor, n, ranging between 1 and 10. Exact power gains were obtained[14] with the help of a computer. Exact power gain lies between the upper limiting power gain and approximate power gain.

6. Conclusions

A new invariant alignability factor, δ, and a new invariant inherent alignability factor, δ_i, have been obtained from the sensitivity of port immittance to small immittance changes and large reactive changes of self-immittance at the opposite port. δ is the inverse of the maximum modulus of the sensitivity for given ρ_S and ρ_L, while δ_i is a particular value of δ when $\rho_S = \rho_L = 0$. The new alignability factor is an invariant for interchange of input and output ports, arbitrary lossless terminations and for immittance substitution. It is related to earlier stability-based factors, including the performance factor. For absolute stability, the value of δ should exceed 0. When $\delta \gg \Pi$, the new alignability factor approximately equals the performance factor, n, and is therefore approximately the inverse of the maximum modulus of loop gain. It is shown possible to evaluate δ from three simple measurements. The power gain margin of an amplifier stage equals δ nepers or $8\cdot 686\ \delta$ dB, when $\delta \ll 1$, while its maximum possible power gain for a given performance factor is inversely proportional to δ^2.

Table 2

Upper limiting, exact and approximate maximum power gains of a test amplifier stage for values of n ranging between 1 and 10

n	g_1 dB	g_2 dB	g_3 dB	$g_1 - g_2$ dB	$g_2 - g_3$ dB
1·0	51·02	48·80	45·02	2·22	3·78
1·5	46·87	45·05	43·35	1·82	1·70
2·0	44·63	43·14	42·13	1·49	1·01
2·5	43·13	41·80	41·19	1·33	0·61
3·0	42·01	40·88	40·43	1·13	0·45
3·5	41·12	40·13	39·78	0·99	0·35
4·0	40·37	39·47	39·21	0·90	0·26
5·0	39·18	38·43	38·26	0·75	0·17
6·0	38·24	37·60	37·49	0·64	0·11
7·0	37·48	36·92	36·83	0·56	0·09
8·0	36·82	36·33	36·26	0·49	0·07
9·0	36·26	35·81	35·76	0·45	0·05
10·0	35·75	35·35	35·31	0·40	0·04

Note:

$$g_1 = \frac{4n}{\delta^2}\left|\frac{p_{21}}{p_{12}}\right|\left\{1-\sqrt{\frac{n_i}{n}}\right\}^2$$

$$g_2 = 4F\left|\frac{p_{21}}{p_{12}}\right|\left\{1-\sqrt{\frac{n_i}{n}}\right\}^2$$

$$g_3 = \frac{4}{n}\left|\frac{p_{21}}{p_{12}}\right|\left\{1-\sqrt{\frac{n_i}{n}}\right\}^2$$

7. Acknowledgments

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9. Appendix

9.1. *Minimum Modulus of* p_{p1}

From eqns. (2) to (4)

$$\rho_{p1} = \mathrm{Re}\,(p_{p1}) = \rho_1 - \frac{M\rho_2 + N\sigma_2}{\rho_2^2+\sigma_2^2} \quad \ldots\ldots(64)$$

and

$$\sigma_{p1} = \mathrm{Im}\,(p_{p1}) = \sigma_1 - \frac{N\rho_2 - M\sigma_2}{\rho_2^2+\sigma_2^2} \quad \ldots\ldots(65)$$

From eqns. (64) and (65)

$$|p_{p1}|^2 = \left(\rho_1 - \frac{M\rho_2+N\sigma_2}{\rho_2^2+\sigma_2^2}\right)^2 + \left(\sigma_1 - \frac{N\rho_2-M\sigma_2}{\rho_2^2+\sigma_2^2}\right)^2 \quad \ldots\ldots(66)$$

To obtain the maximum or the minimum modulus of p_{p1}, partially differentiate eqn. (66) with respect to the variables σ_1 and σ_2 and equate each of these differential coefficients to zero. This gives

$$\sigma_1 = \frac{N\rho_2 - M\sigma_2}{\rho_2^2+\sigma_2^2} \quad \ldots\ldots(67)$$

and

$$\left[\rho_1 - \frac{M\rho_2+N\sigma_2}{\rho_2^2+\sigma_2^2}\right] \times \left[\frac{2\sigma_2(M\rho_2+N\sigma_2) - N(\rho_2^2+\sigma_2^2)}{(\rho_2^2+\sigma_2^2)^2}\right] = 0 \quad \ldots\ldots(68)$$

In eqn. (68)

$$\rho_1 \neq \frac{M\rho_2+N\sigma_2}{\rho_2^2+\sigma_2^2} \quad \ldots\ldots(69)$$

Otherwise eqns. (64) and (65) will yield the value zero for p_{p1}. The inequality sign in statement (69) becomes an equality sign only when η of eqn. (16) is unity and as a result the minimum of p_{p1} is zero. From eqn. (68)

$$\left(\frac{\sigma_2}{\rho_2}\right)^2 + \frac{2M}{N}\left(\frac{\sigma_2}{\rho_2}\right) - 1 = 0 \quad \ldots\ldots(70)$$

yielding

$$\sigma_2 = \left(\frac{-M \pm L}{N}\right)\rho_2 \quad \ldots\ldots(71)$$

Take positive sign in eqn. (71) for the minimum modulus of p_{p1}. Equation (67) then becomes

$$\sigma_1 = \frac{N}{2\rho_2} \quad \ldots\ldots(72)$$

Equation (71) with positive sign and eqns. (64), (65) and (72) lead to the minimum modulus of p_{p1} as in eqn. (13).

It may be verified that the negative sign in eqn. (71) gives the maximum modulus of p_{p1}.

$$|p_{p1}|_{\max} = \frac{2\rho_1\rho_2 - M + L}{2\rho_2} \quad \ldots\ldots(73)$$

for

$$\sigma_1 = \frac{N}{2\rho_2} \quad \text{and} \quad \sigma_2 = \left(\frac{-L-M}{N}\right)\rho_2 \quad \ldots(74)$$

The second differential coefficient of $|p_{p1}|^2$ with respect to σ_2 confirms that eqns. (13) and (73) give the minimum and the maximum modulus of p_{p1} respectively.

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Generation of Random Digital Numbers with Specified Probability Distributions

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Summary: Current interest in digital simulation models has given rise to the need to produce random digital numbers. Several methods of generating pseudo-random sequences of digital numbers are discussed. These methods are used as a basis for generating truly random digital numbers whose probability of occurrence can be set by programming a simple patch-board.

1. Introduction

The generation of random digital signals and numbers is a subject which up to now has received relatively little attention compared with the generation of random analogue signals. This is probably because analogue computers have for a long time been widely employed for simulating and solving problems of a continuous nature; hence the need to generate analogue signals and voltages with specified probability distributions. However, as the use of digital and hybrid simulation of problems has increased, so the need for equipment to produce digital signals has become apparent.

A method is described by which it is possible to generate, in parallel binary form, random digital numbers with a probability of occurrence or probability distribution function which can be set on a programmed patch-board. The method relies on first producing a pseudo-random sequence of digital numbers, the frequencies of occurrence of which are contained in the sequence in the ratio of the probability required for each number. Random digital numbers are then obtained by sampling this pseudo-random sequence; various methods of doing this are discussed.

Among the possible uses of such numbers is the testing of digital communication equipment and associated data circuits. The author has in fact used them as input data for a digital economic inventory simulator.

2. Methods of Generating Digital Numbers

2.1. *Use of Feedback Shift Registers which Produce Pseudo-random Binary Sequences*

It is well known that feedback shift registers can, by a suitable choice of the feedback connections, be made to generate a sequence of binary digits of any desired length.[1, 2] An investigation of the parallel binary numbers produced across several stages of a shift register revealed that the only useful distribution that can be obtained is a uniform one when a sequence of maximum length is being generated.

There is another possible method of generating a pseudo-random sequence of binary digital numbers using feedback shift registers. This is to arrange the required numbers in the distribution in a column, such that it contains the numbers in the ratio of the probabilities required, and then to generate each individual column of digits on a separate shift register. Combining the individual columns of digits in the correct phase will then result in the production of a pseudo-random sequence of binary digital numbers. For example, a sequence of four-bit numbers would require four feedback shift registers, each with different feedback connections.

2.2. *Quantizing of Analogue Signals*

In generating digital numbers by quantizing analogue signals, the limitation arises that only a few well-known distributions can be generated at any reasonable speed. The generated distributions are determined by the characteristics of the analogue signals and by the method of sampling. Analogue signals of many different characteristics can be set up using an analogue function generator, but this is essentially a slow process, and hence is unsuitable for high-speed digital number generation. Two examples of digital number distributions which can be obtained easily are a Gaussian distribution, by sampling a Gaussian noise signal, and a uniform distribution, by sampling a triangular or saw-tooth waveform.

The construction of complicated or unusual distributions by this method becomes exceedingly difficult if not impossible.

3. New Method of Generating Random Digital Numbers with Specified Probability Distribution

The basic principle of the new method presented in this paper is as follows. To start with, all the possible numbers which can be generated are con-

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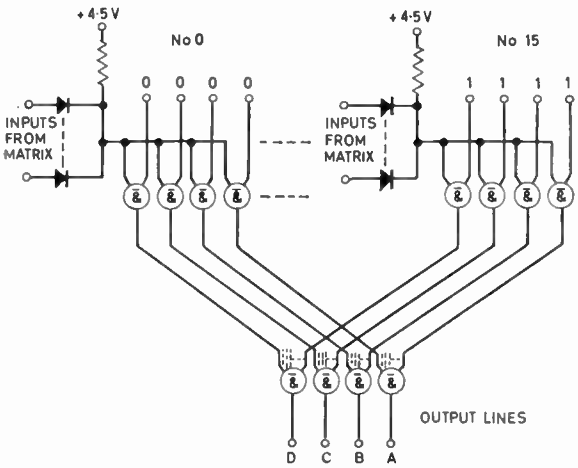


Fig. 1. Number gating unit.

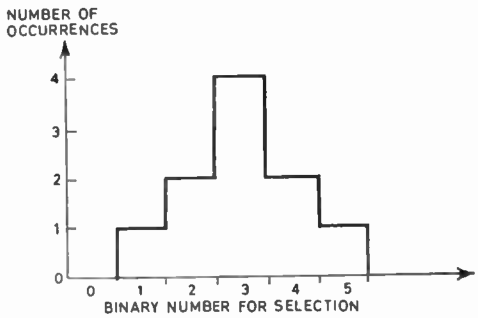


Fig. 2. Number histogram.

tained in simple stores, i.e. wired-in. Each number can then be gated to common output lines by applying a d.c. gating pulse to the gates as shown in Fig. 1. If the numbers are gated in successive time intervals a sequence of digital numbers will appear on the output lines. This sequence can be programmed to have any length N, where N is the number of gating pulses. Suppose, for example, that a sequence of digital numbers is required which has a probability density histogram of the form shown in Fig. 2. The frequency of occurrence of the numbers is as follows:

Number	1	2	3	4	5
Occurrence	1	2	4	2	1

It can be seen that the total occurrence of all numbers in the histogram is ten and that this gives rise to a sequence of length $N = 10$. The ten numbers are now arranged into a random sequence, with the aid of tables of random numbers, or by some other method. A typical sequence might be 2, 4, 1, 3, 3, 5, 4, 3, 2, 3. To generate this sequence, ten gating pulses are required, which are connected to the number gates as follows:

Gating pulse	1	2	3	4	5	6	7	8	9	10
Triggered number	2	4	1	3	3	5	4	3	2	3

The gating pulses are made to occur cyclically thus causing the set sequence of digital numbers to repeat continually.

If the clock rate of the gating pulses is f Hz, then the pseudo-random sequence of digital numbers programmed will appear on the output lines, repeating itself every N/f seconds. Random digital numbers can be obtained merely by sampling the programmed sequence, thus obtaining numbers with a probability of occurrence depending on the relative number of times they occur in the programmed pseudo-random sequence.

The sampling frequency and method of sampling are important in determining the characteristics of the random numbers. Random sampling is necessary if the numbers are to occur at random times. If, however, the numbers are required to change at fixed periodic time intervals, they can be sampled at a fixed frequency f_1, where $f_1 < f$ and f_1 and f have no common factor. If higher speeds are required, the numbers can be sampled at random times and this random sequence then sampled at the desired fixed frequency. The latter method is to be preferred at all speeds if sampling is carried out at a fixed frequency.

4. System Implementation

4.1. *Generation of Gating Pulses*

The first requirement of a method of producing gating pulses is that it must allow the correct number, N, of gating lines to be selected. One method of generating gating pulses is to employ a ring counter, containing only one digit and clocked with a train of square waves. The disadvantage of this method is that each gating pulse requires a separate element.

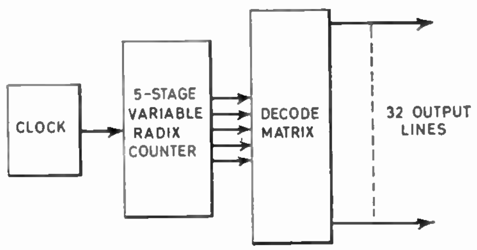


Fig. 3. Gating-pulse generation.

The method devised uses fewer elements but sacrifices some of the simplicity of the ring counter method. The basic arrangement is shown in Fig. 3. Here a clock drives a five-bit variable-radix binary counter. The five outputs of the counter are decoded into 32 outputs by a diode decode matrix.

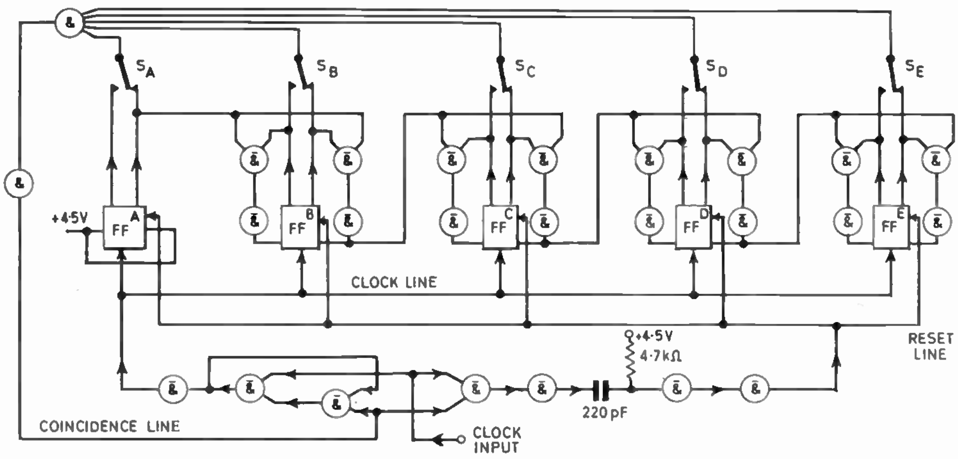


Fig. 4. Variable-radix counter.

4.2. *Variable-radix Counter*

The counter which is shown in Fig. 4 is a five-stage synchronous one which will count in any radix up to 31. The basic counter circuit consists of five bistable elements A, B, C, D and E, and counts in binary code. If it is required to count up to a given number N, the number $N-1$ is set up in binary code on the five switches S_A, S_B, S_C, S_D and S_E. Clock pulses are allowed to enter the clock line and counting proceeds until the number set on the switches is reached, when a coincidence pulse is generated to prevent the next clock pulse from entering the clock line. The clock pulse is then routed into a monostable circuit which generates a reset pulse to reset all the counter stages to zero. As the counter is reset, so the coincidence condition disappears and clock pulses are allowed to enter the clock line again. This cycle of events then repeats. The reason that the number $N-1$ has to be set to count to N is because the counter starts counting from 0 to N, i.e. $N+1$ counts. The correct setting for the count could be made if the bistables were reset to 00001 instead of 00000.

4.3. *Decode Matrix*

A conventional matrix for decoding five output lines into 32 requires 160 diodes; however, by using a dual tree diode decode matrix,[3] the number of diodes required is reduced to 96. A better but more expensive method would be to use logic gates to decode the counter outputs. To ensure that more than one gating pulse is able to trigger the same number to the output lines, the matrix outputs should include series diodes so that, when connected to the input of a number gate, they perform the function of an OR gate. The exact circuit configuration will, however, depend on the type of logic used.

4.4. *Method of Sampling*

The numbers are sampled by a bistable switch which, upon reception of a sampling pulse, causes the bistable outputs to take up the state of the inputs at the sampling instant, as shown in Fig. 5. In order to avoid a hazard or race condition when sampling, care must be taken to prevent sampling pulses occurring when the number lines are in transition. If the random digital numbers are required to change at irregular time intervals then the sampling must be performed at these time intervals. The sampling pulses are generated by allowing a noise source to initiate a trigger circuit every time its amplitude passes above a certain

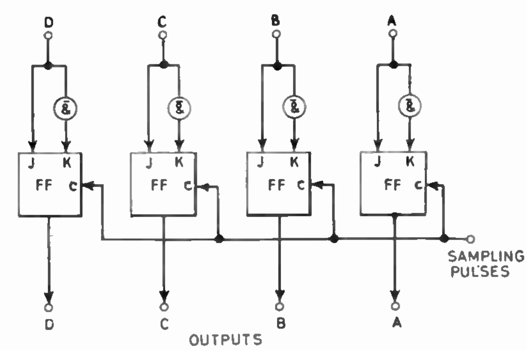


Fig. 5. Bistable sampling switch.

threshold. The sampling time characteristics are dependent on the noise characteristics and it is suggested that a Gaussian noise source is used. In order to set the correct average time between samples, either the amplitude of the noise source or the threshold level of the trigger circuit can be varied, thus controlling the rate at which the threshold level of the trigger circuit is crossed.

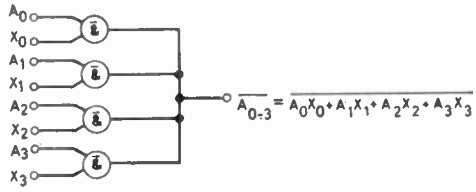


Fig. 6. Collector logic.

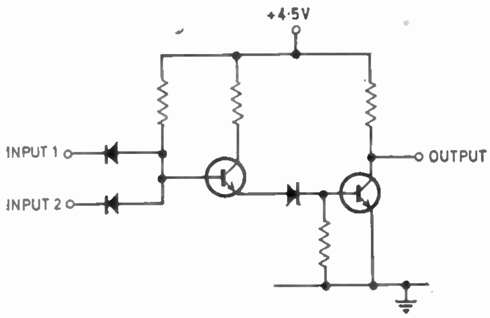


Fig. 7. NAND DTμL integrated logic gate.

5. Number Selection Logic

In the number generator being described 16 digital numbers are available for selection and these are wired to the inputs of 16 two-input NAND gates, the outputs of which are formed into an OR function. This is arranged by using collector logic on the outputs of the NAND gates taken four at a time as shown in Fig. 6. The effect of joining the outputs together is to join together all the collectors of the output transistors of the gates. A circuit diagram of a NAND DTμL (diode-transistor micrologic) integrated logic gate is shown in Fig. 7. If the inputs to these gates are designated A_0, A_1, A_2, A_3, number digits, and X_0, X_1, X_2, X_3, gating pulses, then the output of four NAND gates with collector logic is

$$\overline{A_0X_0 + A_1X_1 + A_2X_2 + A_3X_3} \qquad \ldots\ldots(1)$$

The outputs from other groups of gates are similar.

If eqn. (1) is abbreviated to $\overline{A_{0-3}}$ indicating that number digits A_0 to A_3 are being operated on, then the output A of a four-input NAND gate which has as inputs $\overline{A_{0-3}}$, $\overline{A_{4-7}}$, $\overline{A_{8-11}}$, $\overline{A_{12-15}}$, is given by

$$A = A_{0-3} + A_{4-7} + A_{8-11} + A_{11-15}$$

$$A = A_0X_0 + A_1X_1 + \ldots A_{15}X_{15} \qquad \ldots\ldots(2)$$

This is repeated for the B, C and D digits of the binary numbers. The complete circuit diagram of the number selection logic is shown in Fig. 8.

6. Measurement of Random Digital Numbers

A convenient method of measuring the probability of the generated random numbers uses a gate which can be set to give an output only when a certain set number appears at its input terminals. To measure the probability of the numbers appearing at its input, the gate is set to pass each number in turn for a fixed period. From the number of times each number occurs in this period its probability of occurrence can be calculated. As it stands this method of measurement will give incorrect counts, as the counter has no way of telling apart the same number occurring consecutively with itself. To overcome this difficulty, the number is further gated with the gating pulse which produced it in the first place (see Fig. 9), the counter then receives a pulse each time the particular number appears at the input to the gate.

In order to verify the statistical properties of the numbers generated, different histograms of numbers were constructed and the percentage occurrence of each number measured. The method was to take ten measurements of each number in the histogram, using the counting technique and then to produce an average for each number over the ten measurements. The process was repeated many times and a final average of all results produced. The results for a

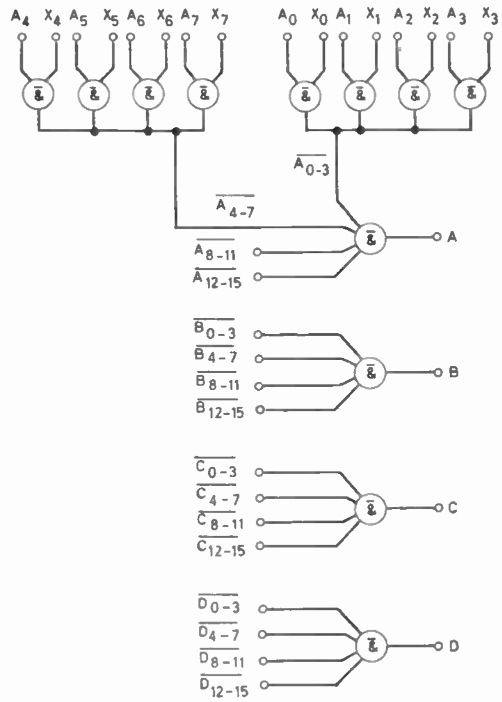


Fig. 8. Number generation logic.

typical test are shown in Table 1 for the histogram shown in Fig. 2. These figures indicate the frequency of occurrence of each number within a one-second period.

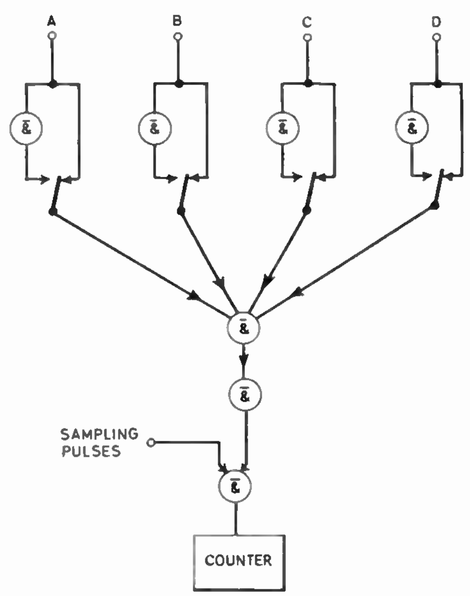


Fig. 9. Measurement of random digital numbers.

Table 1

Frequency of occurrence of each number in a period of one second

Number	1	2	3	4	5
Number of counts in a one second period	60	114	273	162	73
	65	148	259	131	67
	69	149	276	129	65
	68	105	237	104	60
	58	108	267	123	96
	76	110	263	126	77
	51	141	242	111	47
	79	144	227	119	57
	60	110	222	152	61
	73	119	241	135	66
Percentage occurrence	10·3%	19·6%	39·3%	20·3%	10·5%

The results of ten such tests when averaged gives the percentage occurrence of each number as

Number	1	2	3	4	5
Measured occurrence	10·2%	19·8%	39·8%	20·1%	10·1%
Programmed occurrence	10%	20%	40%	20%	10%

This shows that the measured and programmed values are in good agreement. Individual samples are seen to fluctuate quite widely in regard to percentage occurrence, due to fluctuations of the mean level of the noise source used.

7. Distributions Generated

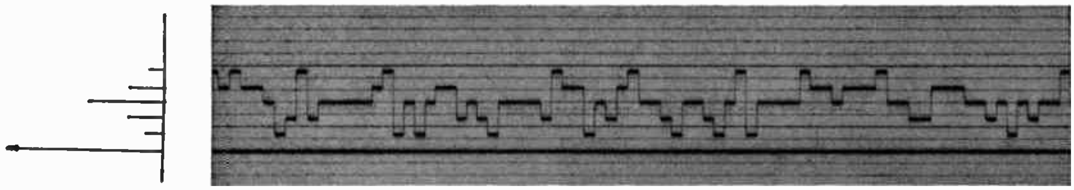
Several well-known statistical distributions of digital numbers have been generated and analogue representations of these numbers, plotted against time, are shown in Fig. 10. The analogue form is obtained by decoding the numbers in a digital to analogue converter. It is at this point that the usefulness of the digital number generator for generating random analogue waveforms should also become apparent.

Figures 10(a) and 10(b) show the distribution of random digital numbers for the histogram designed earlier in the article. The numbers shown in Fig. 10(a) are generated at fixed time intervals by sampling with a fixed frequency train of square-waves, whilst Fig. 10(b) shows the same distribution of numbers generated by sampling at random time intervals. An exponential distribution of numbers generated at fixed and random time intervals is shown in Figs. 10(c) and 10(d). Figures 10(e) and 10(f) show a uniform and a Poisson distribution, while Figs. 10(g) and 10(h) show more unusual types of distribution which it is possible to generate.

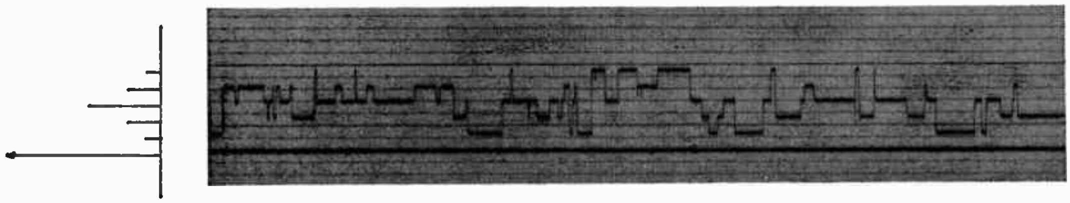
8. Conclusions

The experimental results produced by using the method described for generating random digital numbers illustrate clearly the ease with which different probability histograms can be set up. The distribution of generated numbers can be altered merely by changing pre-programmed patch-boards. A disadvantage of the method is that only discrete values of the probability of numbers in the histogram can be produced. A more versatile generator would allow a continuous variation of probability to occur. A limitation of the present equipment is that it allows only four-bit numbers to be generated. Theoretically, numbers containing any number of bits can be generated as they merely require to be gated to the output highway. The number of digital numbers in the pseudo-random sequence limits the size of the histogram, the total occurrence of all numbers in this sequence being also the number of gating pulses required.

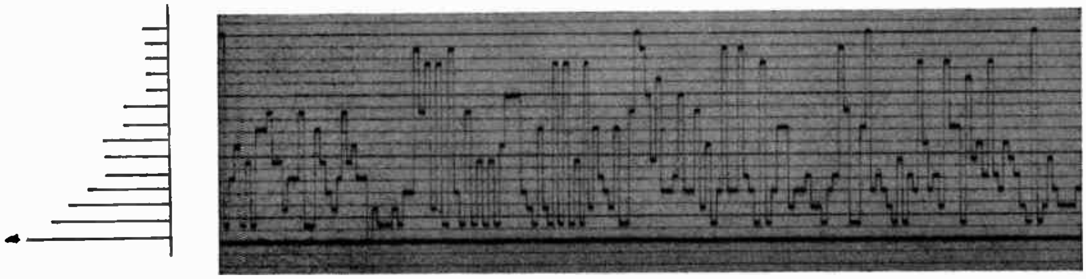
The possibility also exists of using the generator coupled to a digital-to-analogue converter for pro-



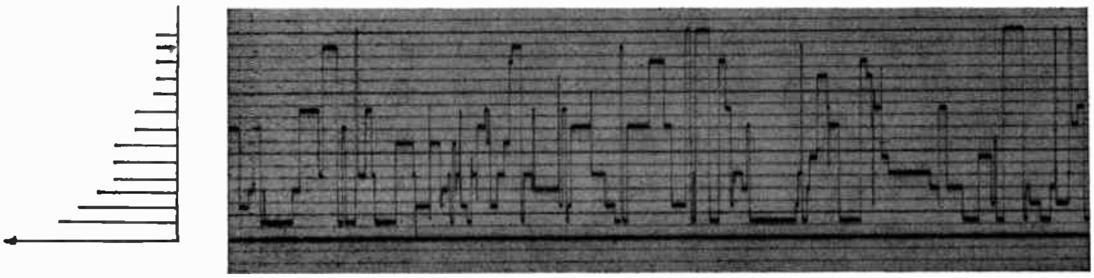
(a) Approximate Gaussian distribution with fixed sampling.



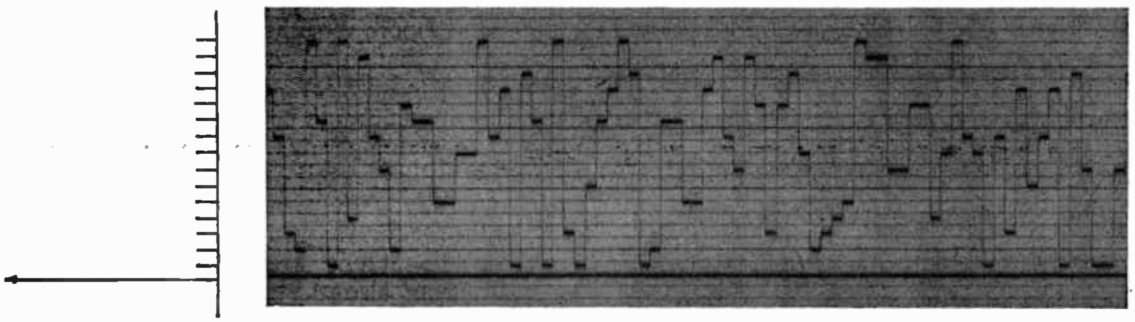
(b) Approximate Gaussian distribution with random sampling.



(c) Exponential distribution with fixed sampling.

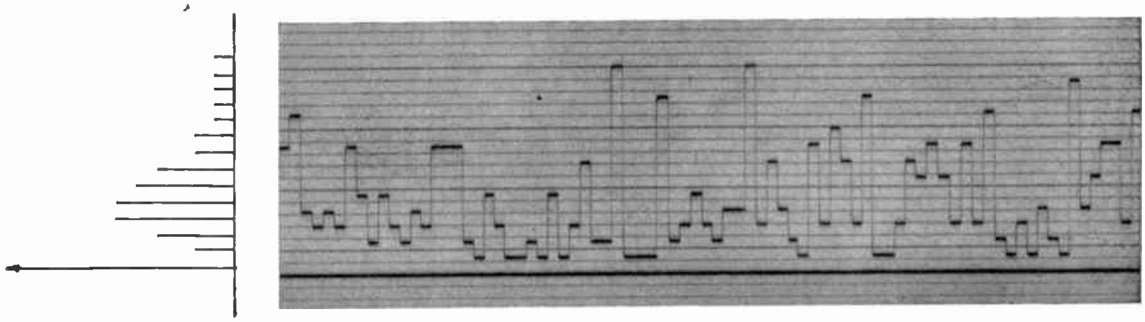


(d) Exponential distribution with random sampling.

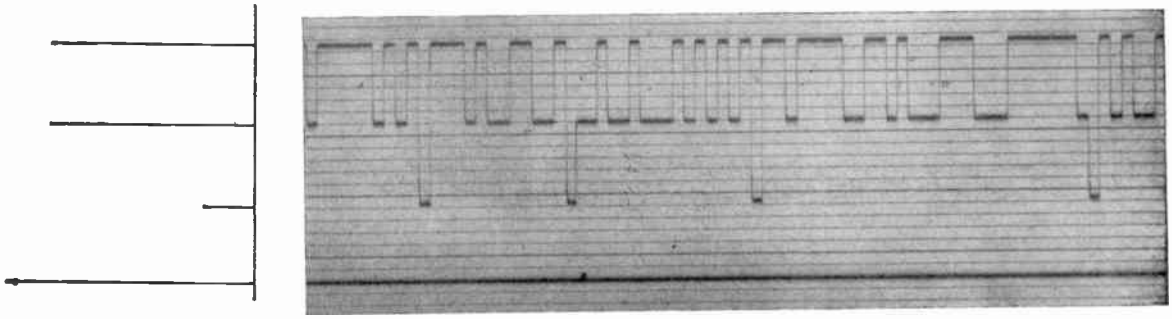


(e) Uniform distribution with fixed sampling.

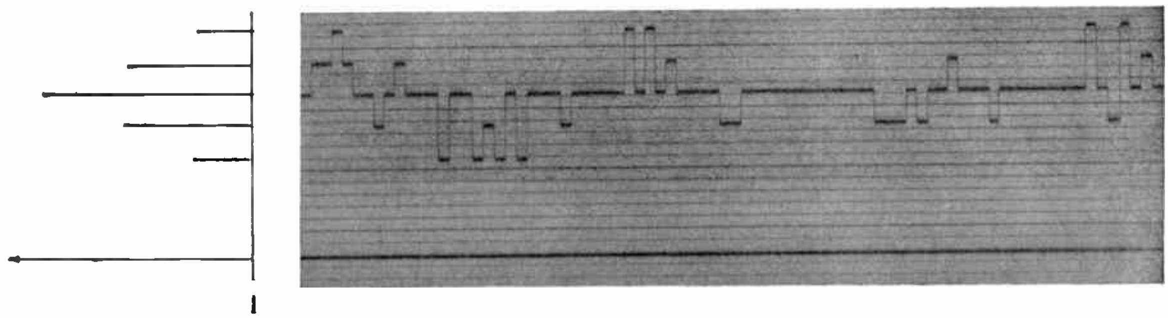
Fig. 10. Analogue representation of statistical distribution of digital numbers.



(f) Poisson distribution with fixed sampling.



(g) see text.



(h) see text.

Fig. 10. (contd.)

ducing random analogue waveforms, in which it is possible to set the probability of the signal having any discrete value.

9. Acknowledgments

The work described in this paper was carried out in the Department of Electronic and Electrical Engineering, at the University of Birmingham. The author wishes to express his appreciation of the facilities provided and wishes to thank his colleagues for their discussions.

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Radio Engineering Overseas . . .

The following abstracts are taken from Commonwealth, European and Asian journals received by the Institution's Library. Abstracts of papers published in American journals are not included because they are available in many other publications. Members who wish to consult any of the papers quoted should apply to the Librarian giving full bibliographical details, i.e. title, author, journal and date, of the paper required. All papers are in the language of the country of origin of the journal unless otherwise stated. Translations cannot be supplied.

M.T.I. IN PULSED-DOPPLER RADAR

Radar returns from fixed targets frequently interfere with those from moving targets. If the transmitter signals are coherent, conclusions for the suppression of completely correlated fixed target returns can be drawn from the spectrum of a limited number of returns at different antennæ pattern shapes. The clutter returns occurring in practice can be classified into several categories. The correlation time differs widely and depends on a large number of events that are difficult to assess. The theory of the statistical signal extraction, given in a German paper, shows that with a coherent transmitter system only a relatively small gain can be achieved over an incoherent system if the clutter is uncorrelated. With correlated clutter the coherent system is far superior. With a detection system designed to be almost optimum for uncorrelated clutter, a high degree of reduction of completely correlated clutter is already possible at, for instance, 10 returns. Clutter suppression depends markedly on the Doppler frequency.

'Extraction of moving targets by pulsed-Doppler radar', S. Nestel, *Nachrichtentechnische Zeitschrift*, **21**, No. 1, pp. 15–23, January 1968.

PLASMA LENSES

One of the methods most frequently employed in the focusing of the electromagnetic energy is the use of lenses. The design and construction of lenses for regions with different frequencies (including that of light) depend on the properties and the index of refraction of the lens material. For the most part, lenses are made of materials with a homogeneous index of refraction; an exception to this rule is represented by Luneberg lenses, for example, which have an inhomogeneous index of refraction. All these lenses have fixed parameters, i.e. neither the index of refraction nor the focal length of the lens can be changed by external action. The necessary change of shape of the constant phase surface of an electromagnetic wave in its passage through a lens with a constant and homogeneous index of refraction is achieved by a suitable choice of lens shape.

Ionized gas is a medium with sufficient refractivity that may be successfully used for the focusing of the microwave energy. Unlike that of the materials hitherto used in the manufacture of microwave lenses, its index of refraction can be easily varied by direct variation of either the plasma parameters or of the magnitude of the external magnetic field in which the plasma is placed. The lenses in which plasma is used for the focusing of the electromagnetic energy, called plasma lenses, are of cylindrical shape formed, e.g., by a short glass cylinder (whose length is of about the same order as its diameter), in which the plasma is enclosed. The electromagnetic energy is focused by means of a radially variable index of refraction which in plasma lenses is the result of a radial density distribution of charged particles in the cylinder. Since the index of refraction can in this case be varied continually by the variation of the external magnetostatic field, plasma lenses may serve as zoom-lenses, i.e. as lenses with variable focal length. Due to the circumstance that plasma parameters may be chosen in a way that makes the losses of the electromagnetic energy in its passage through the lens negligible, plasma lenses stand a real chance of becoming an important element in practical applications.

In a Czechoslovakian paper an analysis is presented of the possibility of focusing the microwave energy by means of a magnetoactive plasma. The basic parameters of a plasma lens are calculated, and the way a magnetoactive plasma may be used in the making of lenses with a variable focal length is pointed out.

'Plasma lenses', J. Musil and F. Žáček, *Czechoslovak Journal of Physics*, **B-18**, No. 1, pp. 66–74, 1968.

DECISION-ADAPTIVE F.S.K. RECEIVERS

Two methods are suggested to improve the detection of orthogonal frequency shift keying signals in the presence of noise and an unknown carrier phase. To get a decision about a transmitted mark or space symbol the past decisions are used and also the received signal during two or more signal intervals. In one scheme a reference is derived from the next preceding signal interval by taking the output from either the mark or the space filter depending on the previous decision and combine this reference signal with both the output from the mark and space filter during the interval under investigation. In the second scheme a separate reference is established for the mark and space channel.

In the analysis a possible phase difference between the mark and space channel is taken into account. The probability of error for the first system is shown to be fairly insensitive to any difference less than about 20°. It is also proved to be conditionally optimal at no phase difference. The second system is suggested as a compromise to avoid the effect of large phase differences and still have a simple detector.

The receivers both incorporate envelope detectors and a simple logic and they are adaptive in the sense that earlier decisions affect the signal processing. It is proved that for both schemes the error probability comes close to that for a coherent detector when the signal-to-noise ratio is about 7 dB or larger.

'Decision-adaptive f.s.k. receivers', L. H. Zetterburg. *Ericsson Technics*, **24**, No. 1, pp. 31–72, 1968 (in English).

SUBJECT INDEX

Papers and major articles are denoted by printing the page numbers in bold type

INDEX OF PERSONS

Names of authors of papers published in the volume are indicated by bold numerals for the page reference

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Czechoslovak Journal of Physics

E.B.U. Review (Switzerland)

Electronics and Communications in Japan (English language edition of *Denki Tsushin Gakkai Zasshi*)

Ericsson Technics (Sweden)

Nachrichtentechnische Zeitschrift (Germany)

Proceedings of the Institution of Radio and Electronics Engineers Australia

Radio Engineering and Electronic Physics (English language edition of *Radiotekhnika i Elektronika*) (U.S.S.R.)

Review of the Electrical Communication Laboratory (*NTT*) (Japan)

Telecommunications and Radio Engineering (English language edition of *Elektrosvyaz* and *Radiotekhnika*) (U.S.S.R.)