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Lasers and Opto-Electronics

When lasers made their dramatic appearance in 1960 it was immediately apparent that a new dimension in electronics had been born, since they offered, for example, the possibility of extending the frequency range of electronic techniques by five orders of magnitude. It was obvious that lasers would have a considerable impact in many other fields too. However, although the principle of coherent light generation had been demonstrated, it was not to be expected that the full exploitation of this technique would be easy. Basically, science and technology are not used to coping with breakthroughs of this extent and many people expected too much too quickly with the result that a few fingers have been singed and a few purses emptied with not always the easy results expected.

In 1964 a conference on Lasers and their Applications was held in London by the Institution of Electrical Engineers with the Institution of Electronic and Radio Engineers as a co-sponsor. This conference provided a review of the new technology and indicated some potential avenues of development. Now, five years later, another conference has been arranged, this time by the I.E.R.E. with the collaboration of the I.E.E., the Institute of Physics and The Physical Society, the Institute of Electrical and Electronics Engineers (United Kingdom and Republic of Ireland Section) and the University of Southampton on the Joint Organizing Committee. The conference, to be held at the University of Southampton, will last three and a half days, and include over 70 papers. These will deal with established theory and techniques and also with up-to-date reports on research still in progress. The overall scope extends beyond that of coherent laser radiation to emphasize techniques and applications in electronics, thus embracing the field of 'opto-electronics', which is of increasing importance.

Already laser development has made significant advances. At the present time gas lasers can provide highly stable outputs, continuous or pulsed, at frequencies from the near ultra-violet to the submillimetre region at power levels up to about 200 W per metre length, generally with low efficiency. Solid-state lasers are essentially pulsed devices and are capable of providing very high powers of up to many gigawatts in pulses as short as 10^{-12} second. Liquid lasers potentially have properties intermediate between those of gas and solid-state lasers. Lasers are basically fixed-frequency devices but the techniques of non-linear optics can produce harmonic generation, sum and difference frequency mixing and parametric interaction, giving rise to tunable light amplifiers and oscillators. A remarkable new development is a recent report of second-harmonic generation from 1·06 μm to 0·53 μm with 100% efficiency.

Lasers have found application in a great many different fields, but their use in radio and electronics is of particular significance to this Institution. The increasing effort 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.

While the so-called 'revolutionary' advantages of lasers are gradually being exploited, there is still need for far more research into laser physics and materials, because the detailed operation of stimulated emission processes is still not fully understood. An important feature of lasers, which has not, in the past, been stressed, is that they involve bulk interaction devices. Looking to the future, it is almost certain that semiconductor junction devices, which although highly developed have fundamental limitations in frequency and power-handling ability, will be superseded by systems in which amplification occurs throughout the bulk of the active material. The laser has provided a very important step in this direction, and this is perhaps one of the most significant contributions to the electronics of the future which will be presented next month in Southampton.

W. A. GAMBLING

Joint Conference on 'Lasers and Opto-Electronics'

At the University of Southampton — 25th to 28th March, 1969

PROVISIONAL PROGRAMME

Tuesday, 25th March

10.00 Welcome and introductory remaiks

10.15–12.30 **OPTICALLY-PUMPED LASERS**

'Recent Advances in Glass Lasers'
C. G. YOUNG, *American Optical Corporation, U.S.A.*

'An Optically-swept Slab Laser'
J. G. EDWARDS, *National Physical Laboratory, Teddington.*

'Laser Damage Mechanisms in Glasses'
J. G. EDWARDS, *National Physical Laboratory, Teddington*

'Some Measurements on the Ho^{3+} in $Y_3Al_5O_{12}$ Laser System'
A. C. EVERITT, E. D. FLETCHER and D. H. PAXMAN, *Mullard Research Laboratories*

'Losses and Energy Transfer in Solid-State Laser Materials'
V. V. GRIGORYANTS, *Institute of Radio Engineering and Electronics, Moscow*

'Transition Cross-sections for Neodymium in Calcium Tungstate'
D. C. HANNA and D. H. ARNOLD, *University of Southampton*

'Experiments with Quasi-continuous Dye-Lasers'
D. ROESS, *Siemens, Munich*

'The Unusual Behaviour of the Liquid Laser $Nd^{3+}/SeOCl_2$ in Non-Resonant Cavities'
D. J. HUNT and T. M. SHEPHERD, *Atomic Weapons Research Establishment, Aldermaston*

2.00–3.30; 4.00–4.50 **PULSE GENERATION AND RESONATORS**

'The Generation and Measurement of Picosecond and Sub-Picosecond Pulses in Neodymium Lasers'
D. J. BRADLEY, S. J. CAUGHEY, G. H. C. NEW and B. SUTHERLAND, *The Queen's University of Belfast*

'Mode Locking by External Loss Modulation of the Saturable Absorber'
J. KERDILES, C.E.A., *Centre D'Etudes de Limeil, France*

'The Production of Two Controlled Q-Switched Pulses from an Electro-optically Q-Switched Laser'
A. R. NEWBERY, *Royal Aircraft Establishment, Farnborough*

'Reflection Properties of an Active Three-Mirror-Resonator'
TH. TSCHUDI, M. SIEGRIST and R. DANDLIKER, *University of Bern, Switzerland*

'The Hurwitz Criteria Optimalization of Reflexion Coefficients of Mirrors in a High-Selective Resonator'
J. KUPKA, *Polish Academy of Sciences, Wroclaw*

'Investigation of a Laser Beam'
J. DE METZ and A. TERNAUD, *Centre D'Etudes de Limeil, France*

4.50–5.30 **GAS LASERS**

'Phase Fluctuations in the Output of a 0·633 μm He-Ne Laser Under Free-running and Self-locked Conditions'
D. G. C. JONES, *University of Sussex*

'Power Broadening of Gas Laser Transitions' and 'Coherence Phenomena in the Coupling of Laser Transitions'
T. HANSCH and P. TOSCHEK, *Institut fur Angewandte Physik der Universitat Heidelberg*

6.00–7.00 Conference Reception

Wednesday, 26th March

9.00–10.40; 11.10–11.40

'Rate Processes in the CO_2 Laser', 'Gain Saturation in CO_2-Laser Amplifiers' and 'High Peak Power from a CO_2 Laser System'
R. C. CRAFER, A. F. GIBSON and M. F. KIMMITT, *University of Essex*

'Electron Temperature Measurements in CO_2 Laser Plasmas'
D. C. TYTE, *Services Electronics Research Laboratory, Baldock*, and R. W. SAGE, *Services Valve Test Laboratory, Haslemere*

GAS LASERS (*continued*)

'A Rate Equation Approach to Saturation in CO_2 Laser Amplifiers'
D. C. Tyte and M. S. Wills, *Services Electronics Research Laboratory, Baldock*

'Self-locking Operation of the CO_2 Laser in Higher Transverse Modes at 10·6 μm'
H. Ito and H. Inaba, *Tohoku University, Sendai, Japan*

'Design and Application of a Continuous HCN Gas Laser'
E. J. S. Becklake, M. A. Smith, C. D. Payne and B. E. Prewer, *E.M.I. Electronics, Wells*

'A Sealed-Off Beryllia Tube Argon Ion Laser'
H. Foster and P. C. Conder, *Services Electronics Research Laboratory, Baldock*

11.40–12.30

SEMICONDUCTOR LASERS

'Injection and Recombination of Carriers in GaAs Laser Diodes Below Threshold'
G. H. B. Thompson, *Standard Telecommunication Laboratories Ltd., Harlow*

'Temperature Effects in GaAs Lasers'
H. R. Wittman, *U.S. Army Missile Command, Alabama, U.S.A.*

'Cooled and Room-temperature GaAs Lasers'
A. R. Goodwin, *Standard Telecommunication Laboratories Ltd., Harlow*

2.00–3.40

NON-LINEAR OPTICS

'Harmonic Generation and Parametric Devices'
E. S. Voronin, *Moscow State University*

'Optical Harmonic Generation in Liquid Crystals'
L. S. Goldberg and J. M. Schnur, *Naval Research Laboratory, Washington, U.S.A.*

'Transient and Steady State Properties of the Optical Parametric Oscillator'
L. B. Kreuzer, *Bell Telephone Laboratories, New Jersey, U.S.A.*

'Optical Parametric Devices using Tellurium'
R. C. Smith and C. R. Stanley, *University of Southampton*

'Image Up-conversion from 10·6 μm to the Visible'
J. Warner, *Royal Radar Establishment, Malvern*

4.10–5.00

OPTICAL MODULATORS AND DEFLECTORS

'The Longitudinal Electro-optic Effect in Lithium Niobate and Other Crystals'
K. F. Hulme and P. H. Davies, *Royal Radar Establishment, Malvern*

'Broadband Light Modulators Design'
M. Bracale and A. Lombardi, *Instituto di Elettrotecnica, Universita di Napoli*

'Magneto-optic Light Modulators'
R. W. Cooper and J. L. Page, *Mullard Research Laboratories, Redhill*

'Temperature Stabilization of the Deflection Pattern of a Digital Light Deflector Containing Single Prisms'
U. J. Schmidt and W. Thust, *Philips Zentrallaboratorium, Hamburg*

'A Television Line Frequency Light Scanner using Piezoelectric Deflector Elements'
J. R. Mansell, *Mullard Research Laboratories*

7 for 7.30 Conference Dinner at the Skyway Hotel, Southampton

Thursday, 27th March

9.00–10.40; 11.10–12.30

OPTICAL DISPLAY AND STORAGE SYSTEMS

'An Optical Data Store'
D. C. J. Reid, P. Waterworth, D. A. Briggs, A. P. Crabb and K. D. Hachfeld, *The Plessey Company Ltd., Towcester and Poole*

'Design of a Large Capacity Holographic Memory'
R. M. Langdon, *The Marconi Company Ltd., Great Baddow*

'High capacity, Random Access Memory'
E. Spitz, *C.S.F., Laboratoire Central, Orsay, France*

'A Laser Photochromic Display System'
G. G. Fuller, *Elliott Space and Weapons Research Laboratory, Camberley*

'Colour Centres in Sodalites for Storage Displays'
P. A. Forrester, S. D. McLaughlan, D. J. Marshall and M. J. Taylor, *Royal Radar Establishment, Malvern*

OPTICAL DISPLAY AND STORAGE SYSTEMS (*continued*)

'An Optical Fixed Store for Use in Computers'
R. J. BOTFIELD and A. N. HILL, *International Computers Ltd., Stevenage*

'Ga(AsP) Visible Lamps and Arrays'
J. R. PETERS and C. E. E. STEWART, *Standard Telecommunication Laboratories Ltd., Harlow*

'Holographic Camera in the Nanosecond Range'
J. C. BUGES and A. TERNEAUD, *C.E.A., Centre D'Etudes de Limeil, France*

'A New Method for Producing Point Holograms'
G. GROH, *Philips Zentrallaboratorium, Hamburg, Germany*

'High Resolution Holography with Large Object Fields'
G. GLATZER, *Philips Zentrallaboratorium, Hamburg, Germany*

2.00–3.20

OPTICAL INFORMATION PROCESSING

'A Holographic Character Generation System'
D. MEYERHOFER, *R.C.A. Laboratories, Princeton, N.J., U.S.A.*

'Optical Data Processing—Pattern Recognition by Coherent Spatial Filtering and Real-Time Correlation'
J. C. BELLAMY and A. WERTS, *C.S.F., Laboratoire Central, Orsay, France*

'Logic Elements on the Basis of Semiconductor Lasers'
N. D. BASOV, V. N. MOROZOV, V. V. NIKITIN and V. D. SAMOYLOV, *P. N. Lebedev Physical Institute, Moscow*

'Dynamic Processes in Semiconductor Lasers with Non-linear Absorber'
N. G. BASOV, V. N. MOROZOV, V. V. NIKITIN and A. S. SEMENOV, *P. N. Lebedev Physical Institute, Moscow*

3.50–5.40

OPTICAL COMMUNICATIONS AND RANGING

'Optical Communication Systems'
F. F. ROBERTS, *Post Office Research Station, Dollis Hill, London*

'Coherent Light Scattering Measurements on Single and Cladded Optical Glass Fibres'
K. C. KAO, T. W. DAVIES and R. WORTHINGTON, *Standard Telecommunication Laboratories Ltd., Harlow*

'Laser Altimeter'
P. A. VAUGHAN, *Elliott Space and Weapons Research Laboratory, Camberley*

'Lasers for Measuring Slant Visual Range'
E. T. HILL, *Plessey Radar Ltd., Cowes*

'Thermal Defocusing and Effects of Wind on the Propagation of CO_2 Laser Radiation in Absorbing Gases'
D. C. SMITH and F. C. GEBHARDT, *United Aircraft Research Laboratories, East Hartford, Conn., U.S.A.*

'Atmospheric Scattering at Wavelengths of 0·69 and 1·06 μm'
M. E. JUDGE, *Elliott Space and Weapons Research Laboratory, Camberley*

Friday, 28th March

9.00–10.30; 11.00–11.40

OPTICAL RANGING AND NEW LASER TECHNIQUES

'Application of Dye Lasers in a Radar used to Probe the Upper Atmosphere by Resonance Scattering'
M. R. BOWMAN, A. J. GIBSON and M. C. W. SANDFORD, *Radio and Space Research Station, Slough*

'New Laser Interferometry Methods of Measuring High-speed Model Missiles'
H. D. VOM STEIN, P. RATEAU, G. SCHULTZE and B. KOCH, *French-German Research Institute, Saint-Louis, France.*

'Influence des Caracteristiques de la Surface Reflectrice D'un Objet sur la Portee D'un Telemetre Laser'
C. VERET, *Office National d'Etudes et de Recherches Aerospatiales, Chatillon-sous-Bagneux, France*

'Laser Navigational Co-ordinate Systems and Aids'
M. R. WALL, *Atomic Weapons Research Establishment, Aldermaston*

'Measurement of Creep and Small Thermal Expansion Coefficients by a Laser Interferometric Method'
O. O. ANDRADE and G. C. THOMAS, *University of Southampton*

'Measurement of Diffusion Constants of Macromolecules by Digital Autocorrelation of Scattered Laser Light'
E. JAKEMAN, C. J. OLIVER and E. R. PIKE, *Royal Radar Establishment, Malvern*

'Measurement of Brillouin Line Widths using a Super-Mode Laser and Frequency Stabilized, Digital Fabry-Perot Interferometer'
D. A. JACKSON, *University of Kent*, and R. JONES and E. R. PIKE, *Royal Radar Establishment, Malvern*

'The Laser Triggered Spark Gap'
S. H. KHAN, *Engineering Laboratory, Oxford*

'A Laser Triggered Spark Gap with Applications to Opto-Electronics'
D. J. BRADLEY, J. HIGGINS, M. H. KEY and S. MAJUMDAR, *The Queen's University of Belfast*

11.40–12.30 Post-deadline papers

Conference Closes

U.D.C. 621·398: 621·39

A Review of Progress in Underwater Acoustics

By

Professor

D. G. TUCKER, D.Sc., C.Eng., F.I.E.E., F.I.E.R.E.†

Reprinted from the Proceedings of the Institution's Convention on 'Electronics in the 1970s' held in Cambridge on 2nd to 5th July 1968.

Summary: The whole field of civil work in sonar and underwater acoustics is reviewed, 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. Civil applications of underwater acoustics are growing rapidly, and 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.

1. Introduction

The term 'underwater acoustics' is to a very large extent synonymous with 'sonar'; it is very doubtful if there is any significant work in underwater acoustics which is not either associated directly with sonar or closely related to it. In this review, therefore, the main theme will be sonar, its applications, and the research associated with its design and use. Two applications of underwater acoustics which are not sonar but are closely related are underwater communication and oceanographic acoustic instrumentation, and these will also be briefly discussed.

In reviewing this subject, the main difficulty is that the greater part of the research, development and application in underwater acoustics is done for naval purposes and is largely classified 'secret'. We can, therefore, discuss only the smaller, but nevertheless very significant part which is done for civil application or which is excluded from the security classification.

Perhaps the most significant matter to report at this stage is the very rapid growth of interest in civil applications of sonar. Echo-sounders have been used in navigation and fish-detection for two to three decades, but in this simple form have been rather 'taken for granted'. In the last few years, however, interest in marine science and technology generally has grown enormously in nearly all countries, and the need for underwater observation and communication has grown in proportion. At the same time, research has shown the way to vastly more refined and powerful sonar systems, and industry has been taking an increasing interest in the matter, although in this country it is still loth to speculate in the civil field. We seem to be poised for a great expansion in the use of sonar.

The review is divided into two parts, one dealing with civil applications of sonar and underwater acoustics, and the second dealing with more basic research.

Part 1. CIVIL APPLICATIONS

2. The Importance and Problems of Very Short Range Sonar

Traditionally, sonar engineers strain to achieve long ranges of detection, and, in the naval field particularly, ranges of many miles are achieved. It is perfectly clear why long ranges are important and why they are difficult to achieve, and we need not dwell on them at the moment. What is worth emphasizing, however, is that there is an ever-growing interest in very short ranges, and these have their own problems too.

† Department of Electronic and Electrical Engineering, University of Birmingham, Birmingham 15.

2.1 *Near-field Resolution*

Sonar offers a means of observation in muddy and turbid water where optical methods are ineffective, and is therefore of potential importance in police searches, in civil engineering (e.g. river and harbour works) and in studying fish behaviour. In the latter case, moreover, it is fairly certain that the high-frequency acoustic waves which are used (say above 100 kHz) are quite undetectable by fish and therefore cannot modify their behaviour. On the other hand, it is well-known that light modifies fish behaviour to a very great extent[1, 2, 3] and is therefore unsuitable as a basis of observation except when natural illumination is adequate. In these applications it is essential to use a sonar giving very high angular (or lateral) resolution as well as high resolution in range. Range resolution can usually be readily increased by using a very short pulse. Lateral resolution, however, raises difficulties.

To obtain a beam of very narrow angular width requires a transducer which has a length of many wavelengths. For example, a 0·5° beam (measured at its 3-dB points) requires a transducer length of approximately 120 wavelengths. To keep the transducer small, as high a frequency as possible is chosen. But the attenuation in the water increases rapidly with frequency. For example, at 500 kHz it is in the region of 0·1 dB/m, while at 1 MHz it is in the region of 0·3 dB/m. Since there must be a specified figure for the maximum range, it is clear that the frequency cannot be increased indefinitely. Thus there is a limit to the reduction in size of the transducer. If, for example, the maximum range is to be about 50 m, then 500 kHz would be a reasonable choice of frequency, the wavelength would be 0·3 cm, and the transducer length (l) for a 0·5° beam would be 36 cm.

The concept of the 0·5° beam, is, of course, a far-field one, based on the idea that the point at which it is measured is well-removed from the transducer which is assumed to be a point from which the beam boundaries are drawn. But at close ranges, it is clear that the concept of angular beamwidth is invalid; if

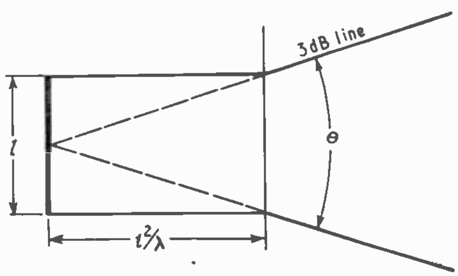


Fig. 1. Geometry of beam near transducer.

the transducer is straight (i.e. unfocused) the beam can hardly be narrower than the transducer length, and the geometry becomes as shown in Fig. 1. The far-field beamwidth is developed only after a range of about l^2/λ, i.e. about 40 m from the transducer. This is almost the maximum range specified, so that we have 'near-field' operation almost throughout. The lateral resolution is therefore clearly not measurable in angle, but in distance, and is thus about 36 cm. The interesting thing is that if the transducer length were halved, the lateral resolution would be *improved* at ranges out to 20 m, although the far-field angular beam-width would be doubled.

The situation described above is clearly a very unsatisfactory one. Fortunately it can be overcome by using multiplicative signal processing on reception.[4] In this system the receiving transducer is divided into two, and the signals from the two half-transducers are multiplied together. Thus a target can give a signal output from the receiver only if it lies in the beams of both half-transducers; moreover the far-field beam-

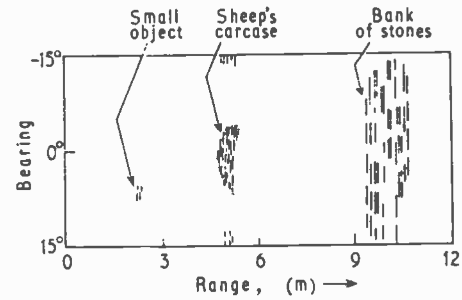


Fig. 2. Display of sector-scanning sonar showing accurate delineation of sheep's carcase in near field, using multiplicative signal processing.

width is one-half of that corresponding to the length of the transducer used normally, and is formed from a distance of $3l^2/16\lambda$ outwards. Thus the near-field beam is extremely narrow, and lateral resolution of one-sixth of the transducer length (or less) is feasible.[5] In the example taken above, the overall length of the transducer would be 18 cm, and the lateral resolution might be quoted as 3 cm or 0·5° whichever is the poorer. This is clearly an enormous advantage for the multiplicative system. An example of a short-range sonar picture obtained on an electronic scanning system using multiplicative processing is shown in Fig. 2.

2.2 *Fish Counting in Rivers*

Another short-range application of sonar which is of very great importance and is attracting a great deal of attention is the counting of fish migrating up rivers. This is of particular importance in respect of salmon, where both sport and commercial fishing represent many millions of pounds annually even in our small country. For the effective control of fisheries, for assessing the effects of waterworks schemes, of hydro-electric schemes, and of pollution, it is necessary to be able to count fish accurately. Sonar offers the best means of doing this in normal river channels.

An American system[6] uses a number of transducers placed on the river bed with upward-looking wide beams. A British system being developed in the author's Department uses transducers at the banks of the river with narrow beams looking across the river.

Two sets of beams, one placed a metre or two higher up the river than the other, and an adequate digital electronic processing system, permit direction of migration to be distinguished; also size can be determined by echo-strength in relation to range, and so on.[7,8] Difficulties due to false counts can arise in rivers with a good deal of weed (e.g. water buttercup), especially when this is cut in the late summer and drifts down the river in clumps. Special electronic discrimination against large and/or diffuse targets may therefore be necessary in some rivers. Aeration of the water may also be a problem in fast turbulent rivers. From this and many other points of view a low frequency, say 50 kHz, would perhaps be preferable to the 500–1000 kHz which has so far been used; but the necessity of obtaining a narrow beam from small transducers has been a dominant factor here, as in so many sonar problems.

3. Long-range Sonar

Since we have to confine attention to civil applications, the best example of a long-range sonar is the system called GLORIA† being developed at the National Institute of Oceanography for acoustic mapping of geological features on the sea-bed at ranges up to 15 km from the ship.

The principles of the use of sideways-looking shipborne sonar for geological mapping are by now well-

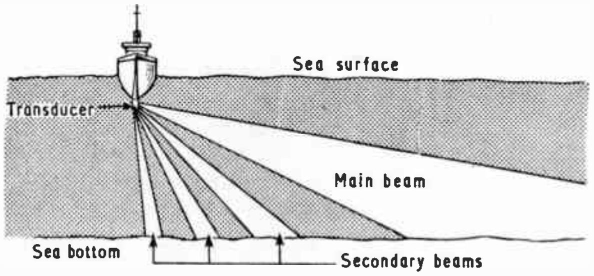


Fig. 3. Sideways-looking sonar for study of surface geology of sea-floor.

known. The technique originated in the Royal Navy in the late 1940's and was further explored under the author's direction in the R.N. Scientific Service in the early 1950's. Since then it has been widely and successfully exploited by the N.I.O. and has led to the acquisition of much new data on the sea-bed. The sonar-beam is narrow in the horizontal plane but wide vertically, so that as the ship proceeds along its route, a map of the acoustic scattering from the sea-bed is built up on recording paper. The sidelobes of the vertical beam pattern can also be exploited, and with experience the records can be interpreted in geological terms, giving in effect a map of the hardness and roughness of the sea-bed rocks or sediments, with clear indications of geological faults, sand waves, ridges, etc. The general arrangement is shown in Fig. 3.

† Geological Long Range Inclined Asdic

Existing equipments (e.g. Ref. 9) have a range of about 1000 m. The effectiveness of the system would clearly be greatly enhanced by a much longer range and although GLORIA is a very expensive instrument because of its high power, large transducer arrays mounted on a towed body with electronic stabilization, etc., yet it will probably represent a very considerable economy in reducing the amount of ship's time required to survey a given area.

GLORIA's parameters may be summarized thus:

3 dB beamwidth: 2·5° horizontal, 10° vertical.

Maximum acoustic power about 60 kW.

Pulse duration: 5, 12 or 30 ms, constant frequency 6·6 kHz, or 1, 2 or 4 s, swept frequency, 6·4±0·1 kHz, (signal-processing gain 24 dB on 4 s pulse).

Range resolution interval minimum 8 m.

It could be argued that a much better angular resolution will be required for further progress to be made in acoustic mapping, and a possible way of obtaining this is the use of aperture synthesis techniques which have been successfully applied in radar mapping.[10] It is believed that no work has been done on this yet for acoustic applications.

4. Fish Detection

Fish detection and location by sonar as an aid to efficient catching in the sea fishing industry is perhaps the most important civil application of sonar. The literature of the subject is vast, but there are some very useful collections of papers (e.g. Ref. 11), and the Institution's Conference on 'Electronics in Oceanography' held at Southampton in September 1966 included several papers on this subject.[12] The author's Buckland Lectures for 1966[3] also covered much of the ground. It therefore seems that we can now treat this topic rather more briefly than its importance would otherwise deserve.

4.1 *The Use of Echo-sounders*

The main bulk of acoustic fish detection in the fishing fleets is still done with echo-sounders. These usually operate somewhere in the frequency range from 10 to 50 kHz, with very wide beamwidths (typically 30°) in order to use only small transducers. For detecting fish in mid-water they are reasonably effective. For detecting fish within a few metres of the sea-bottom (as is required in the demersal trawling industry, which is perhaps the largest part of the British fishing industry) they are not very effective. This is because their wide beamwidth permits powerful echoes from the sea-bottom to swamp most of the fish echoes. Figure 4(a) illustrates the problem; h is the height of the trawlable region, and clearly fish below the arc ABC cannot be detected. If, due to the horizontal movement of the boat or its own motion, a fish passes through the beam, it can in principle be detected as in (b) in the figure, provided the pulse length is less than h.

In practice the bottom profile on the recorder chart is irregular due to the up-and-down movement of the ship, and the fish echoes cannot be easily detected without the aid of the 'white line' device[13] or 'sea-bed lock'.[14]

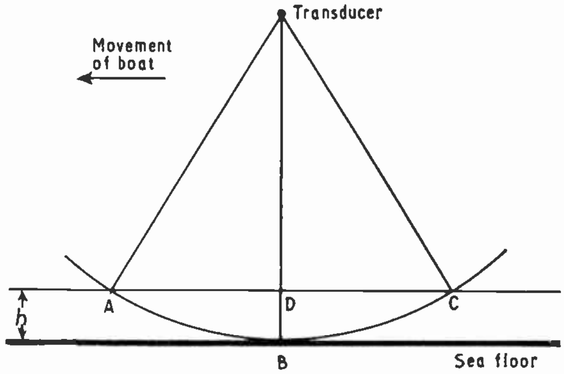
The detection of fish near the sea-bottom is evidently made much surer by the use of much narrower beams. If the beamwidth is brought down to just a few degrees, stabilization of the beam against ship's roll is necessary, and the transducer will have to be very large unless the frequency is raised or unless the new techniques of non-linear acoustics (see Part 2 below) are used. If the frequency is raised, the maximum depth at which fish can be detected is reduced owing to the increased attenuation. Experiments are being conducted by the White Fish Authority with promising results.[15] To avoid the expense of stabilizing large transducers and yet to maintain an acceptable maximum depth, the W.F.A., with the collaboration of the author's Department, is now having a lower-frequency, narrow-beam echo-sounder built with a transmitter using non-linear acoustic wave interaction to give a narrow beam from a small transducer.

The sort of detection performance to be expected from the best commercial echo-sounders is represented by a maximum range of 500 m on a single cod (say 1 m in length).

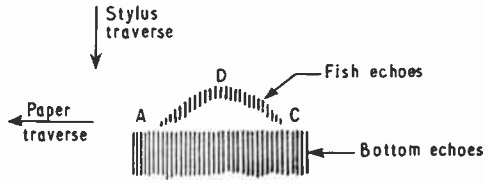
4.2 *The Use of Forward-search Sonar*

For the detection of mid-water shoals of fish, such as herring, a sonar with a nearly horizontal beam which can be directed in any desired direction is very useful and effective, and may have a range of up to 1000 m on a shoal of fish. Such sonars usually operate with frequencies around 50–70 kHz and beamwidths around 10°. Mechanical scanning is used, and with the wide beamwidth is reasonably satisfactory.

For detecting fish near the sea-bottom by forward search, the problem is much more difficult, indeed it may be said to be intractable. The problem is studied in some detail elsewhere.[3, 16] For now we need only say that to pick out a fish echo clear of swamping bottom echo, a very narrow beam is required, and for useful results to be obtained with such a narrow beam, a within-pulse electronic sector-scanning system is essential.[3, 17–21] The scan might be horizontal with a beam which is narrow horizontally and relatively wide vertically, or vertical with a beam which is narrow vertically and relatively wide horizontally. In the first case it is hoped that a fish would occupy so much of the cross-section of the beam that bottom echoes could not dominate; in the latter, it is hoped that the fish and bottom echoes can be separated in range. The latter seems the more promising, and the Fisheries Laboratories, Lowestoft, with the collaboration of the Admiralty, will shortly be conducting trials of the system, with a vertical beamwidth of



(a) Geometry of echo-sounder beam.



(b) Echo-trace of fish at height h above sea-bottom as boat moves.

Fig. 4.

0·33° and the rather limited maximum range of about 200 m. It may well be that a beamwidth even narrower than this will be needed (probably a sharp edge on one side of the beam would be easier to obtain and would suffice), and certainly a range of 400–500 m is needed to make the system really useful. Work towards the achievement of these latter figures is proceeding in the author's Department. It cannot, however, be claimed that there is any real confidence in an operationally-acceptable sonar emerging. The operational problems of using such narrow beamwidths seem great.

The within-pulse electronic sector-scanning (w.p.e.s.s.) sonar is not the only kind of sonar which can give a rapid scan over a sector. The continuous-transmission frequency-modulated (c.t.f.m.) sonar system[22, 23] can give a similar performance, although usually at rather greater expense since a large bank of expensive filters is required. The c.t.f.m. sonar has been used for midwater fish detection in the U.S.A.[24, 25] and by using a narrow beam on reception (0·6°–6°) has enabled a detailed picture of the fish distribution to be obtained in the same way as with w.p.e.s.s. (e.g. Ref. 21), and in a way which is quite impossible with mechanically-scanning pulse sonar.

4.3 *The Use of Doppler Detection*

It has long been postulated[26] that since a fish would normally be moving relative to the sea-bottom, a possible way of detecting fish very near the bottom would be to use the Doppler shift of frequency in a c.w. transmission. Since fish speeds might be very low ($<0{\cdot}5$ m/s) compared with ship's speed (say 4 m/s) there might be difficulty in separating the fish echo frequency from the spectrum of the bottom echoes. Research in the author's Department on another aspect of Doppler detection some years ago was rather discouraging, and the application to bottom fish detection was not pursued. But as no other really promising solution to the problem has been found, the matter is now being taken up again.

For midwater fish detection in good conditions, i.e. with little volume reverberation, Doppler detection has been found very effective[24]—but under these conditions *detection* is easy by any method. However, it has been shown that there is a spread of spectrum associated with the fish Doppler echoes which is due to the body motions of the fish, and that this spread is related to the size of the fish. This is therefore a possible means of measuring fish size.

4.4 *The Counting and Sizing of Fish at Sea*

Another aspect of fisheries sonar is that of determining the abundance of fish. For this purpose it is necessary to count individual fish echoes, and to estimate the size of individual fish. The former process can be done automatically by electronic circuits if the acoustic system has sufficient resolution to ensure that all echoes recorded are from single fish, and experimental counting sonars have been made and used.[27] If the echoes are from shoals, then the size of shoal can be estimated from the time integral of echo signal power.[28] It is possible to separate echoes from single fish and from shoals by suitable electronic circuitry, and count them separately.[29] Cushing[30] has applied knowledge of the acoustic system and the statistics of observation to make relatively accurate estimates of fish abundance.

5. Underwater Point-to-Point Communications

Acoustic waves provide the best (and usually the only) means of communication underwater when direct wire links are not permissible. In civil applications, acoustic underwater communication is less than two decades old, however. Two main kinds of system can be distinguished: telemetry and speech. A thorough review of this subject has been recently published,[31] following a meeting of many interested parties, and consequently our discussion here will be very brief.

In underwater acoustic telemetry, data rates are still very low, with a maximum of only a few bits per second, and in most cases there is not a great deal of difficulty in obtaining accurate results over ranges up to 1000 m or so in spite of the very poor medium of transmission. For it must be said that the sea offers many obstacles to acoustic communication, with multipath propagation problems dominant. With its sharp boundaries at the surface and sea-bottom, with turbulence and thermal inhomogeneities, with gaseous and solid particles (including fish), it is a difficult medium.

Although earlier telemetry systems, e.g. for transmitting information regarding depth of a trawl, used analogue signals such as variable frequency,[32] the recent trend is to use pulse signals of binary form,[33] sometimes with time-division-multiplex so that a number of information channels may be provided.[34]

Even at the low data rates used, trouble is experienced from inter-symbol interference due to multipath propagation, and efforts are now being made to overcome this by an adaptive transducer system in which a highly-directional beam is maintained in the direct path using interpolated test pulses.[35]

It seems unlikely that data rates above 400 bits/s and ranges beyond 8 km will be required (or obtainable) in the foreseeable future for fisheries and oceanographic applications.[31]

When it comes to speech and message communication, the situation is rather different in some respects, as larger bandwidths have to be handled, and in the case of communication with and between divers there is the additional difficulty of obtaining intelligible speech in the diver's environment. Nevertheless, underwater communication with submarines is practicable and indeed in general use;[31] and diver's communication equipment,[36] using either direct audio or frequency modulation of a carrier at around 70–100 kHz, has been made commercially.

6. Acoustics in Oceanographic Instrumentation

A great deal of work described in this paper comes into the broad category of acoustics in oceanography, and consequently there is significance in the title of this section. A recent review paper by McCartney[37] gives some coverage of this topic and numerous references to other publications.

Depth measurement is one of the important instrumental problems of oceanography, and the attainment of an accuracy of 0·1% in depths of 6000 m requires a precision echo-sounder instrument as well as knowledge of acoustic velocity throughout the depth. Acoustic velocity can be measured directly, but is more usually obtained indirectly from temperature measurements. Fairly low frequencies have to be used for these great depths—e.g. 12 kHz, yet the beamwidth must not be too great if certainty is required as to the point from which the echo is returned. Thus large transducers are necessary, but the cost and weight can be minimized by using crossed fan-beams.

Acoustic pingers, transponders and hydrophones are widely used in oceanographic instrumentation for monitoring the position and track of surface and submerged buoys of various kinds, and as the basis of a precise but localized navigational system for surveying and data-collection operations.

Acoustic command systems are also used in which instruments (such as a collecting bucket of a biological sampling net) may be operated remotely by means of an acoustic signal.

7. Digital Signal Processing in Sonar and the Influence of Microelectronics

The use of digital computers to process the information obtained from seismic arrays (as also from aerial arrays used in radio-astronomy) is well-established;[38, 39] it is, of course, based on the recording of the signal outputs from the array elements and subsequent (off-line) processing in a general-purpose computer. On-line digital processing was used in a large, long-range wideband receiving sonar array by Anderson and Rudnick.[40, 41] Digital read-out from an echo-sounder has also been developed[42] so that water depths can be fed directly into a data collection system. All these systems have some very interesting features. They are, however, largely unrelated to what the author considers one of the pressing problems of sonar; namely, how to develop sonar systems of a basically digital type to take advantage of the microelectronic digital circuit units already readily available.

In spite of the attractive performance of within-pulse electronic sector-scanning sonar equipments already demonstrated, the fishing industry has been completely resistant to the idea that they should be made commercially for fitting in fishing vessels. The reason given is the price; the existing equipments using analogue (or linear) circuitry of the discrete-element type are a great deal more expensive than the simple fixed-beam or mechanically-scanning sonars at present in use. The higher data rate of the electronic-scanning sonar, which give a complete picture of a sector on every pulse transmission, does not apparently compensate for the much higher cost.

In this situation, it has for some years been the author's contention that the solution lies in designing a within-pulse electronic-scanning sonar which does not cost more than the simple existing sets, and that the only way of achieving this in the near future lies in the use of digital microelectronic circuits. As a result of this policy, a new kind of sonar has been developed in the author's Department, originating with Nairn[43, 44] and continued by Griffiths and Creasey[45] with theoretical investigation by Hudson.[46] In this

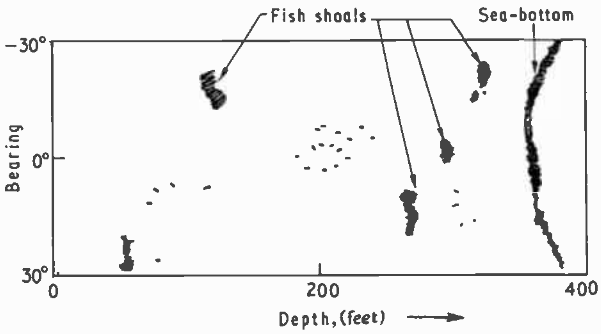


Fig. 5. Display of digital sonar, showing small fish shoals. Sketch is from photograph of echoes from 5 pulses superposed.

sonar the receiving transducer array is, as in other scanning systems, divided into sections. The signal received from each section is amplified and chopped to a predetermined level, so that only the zero-crossing positions remain to give information about the signal. Short pulses are generated at each zero-crossing, and by means of counting circuits driven by a fast clock-pulse, the time differences between the pulses on each pair of adjacent channels are measured and recorded on the digital counting circuits. If there is no background noise, but only coherent signal, then these differences will be the same for all pairs of adjacent channels, and the magnitude of the difference will define the direction from which the signal has come. If, however, there is no coherent signal, but only background noise, then the differences will be random. The average difference is therefore measured by the digital circuitry in order to determine the bearing of a signal; and the departure of each individual difference from this mean is also measured and the average of these over the whole array is determined, so that if it is too large, the sample may be rejected as not representing a true signal. This whole process is repeated every two or three cycles of the received wave, and if consistently-low average departures are obtained, the wave is accepted as signal. A bright-up pulse is then given to the B-scan display at the bearing position determined by the average difference, and, of course, at the correct point in the range scale.

It can be seen that the digital circuitry provided in this sonar is a special-purpose on-line digital computer. It is really quite a simple one, and in the latest experimental model, using commercial microelectronic circuits, the total cost of the electronic equipment (including transmitter and power unit) has been under £200. This model has seven sections to the transducer. Many experimental trials in tanks, reservoirs, rivers, and at sea have been made, and its operational performance is beginning to be understood theoretically.[47,48] It is definitely inferior to the more-expensive analogue electronic-scanning sonars, but recent experiments suggest that this inferiority is not very significant.

Its performance is quite impressive, and a typical display obtained with it is shown in Fig. 5. It is hoped that this sonar will prove a boon to the fishing industry.

In the long term, it will probably be necessary to develop an electronic-scanning sonar using analogue microelectronics. An adaptation of the digital sonar described above, using the same basic concept but analogue circuitry, has been tried but without any impressive success.[49] The correct long-term answer may well be to reproduce the analogue scanning sonars, already so well proved, in microelectronic form, with probably some parts of the process done digitally, and research to this end is proceeding.

Part 2. RESEARCH

8. Non-linear Acoustics

The application of non-linear wave interactions to practical problems is probably one of the most significant developments in underwater acoustic research. That acoustic propagation is inherently non-linear has been known for a very long time, but the exploitation of this knowledge is comparatively recent. That practical applications were possible became clear from work in America early in the present decade,[50, 51] but the main work of exploitation has been that of Berktay and others at the University of Birmingham.[52–60] Further theoretical work has been done in Norway.[61, 62] Some attempts at a simple account of the basic concepts and applications have also been published.[3, 63]

The propagation of acoustic waves is non-linear because the velocity of propagation is a function of the instantaneous amplitude of the wave. This arises because of two separate effects:

(i) the particle velocity adds to the nominal propagation velocity; i.e. when the particle velocity has its maximum positive value, the net propagation velocity is a maximum, and when the particle velocity has its maximum negative value, the net propagation velocity is a minimum.

(ii) the propagation velocity (apart from the above effect) is given by $\sqrt{(\kappa/\rho)}$ where κ is the bulk modulus and ρ is the density of the medium; both κ and ρ vary with the acoustic pressure and so give a propagation velocity dependent on the instantaneous wave amplitude.

Thus the (positive) crests of the acoustic wave travel faster than the (negative) troughs, and the wave at some point distant from the source becomes distorted. If it started as a simple harmonic wave, harmonics are introduced; if it was a complex wave, then intermodulation products are produced. This can be exploited in transmitting and receiving applications, which need to be dealt with separately.

8.1 *Narrow Transmitting Beams from Small Transducers*

Consider a square or circular transducer transmitting two closely-spaced frequencies f_1 and f_2 and having a side or diameter which is many wavelengths at these frequencies. Then a narrow beam is formed which in its near-field is approximately collimated. As the wave of two frequencies propagates along this beam, non-linear interaction takes place and a wave of frequency f_1-f_2 is generated. The phase velocity of this is evidently correct for propagation in the same direction as the 'primary' waves at f_1 and f_2, but it will not be able to propagate in other directions. One can indeed regard the interaction volume as forming an end-fire array of length very great compared with the dimensions of the transducers. Thus a highly-directional transmission is obtained at the low frequency f_1-f_2 using transducers so small that at this frequency they would be practically omni-directional. Moreover, there are no side-lobes. The practical value of this is obvious.

A great deal of theory and experiment have been done, and it is clear that for a given power and (small) transducer size it is possible to obtain as great an intensity at a given distant point by this means as by normal transmission, yet with the most important advantage of high directionality.

The principle is currently being applied to a narrow-beam echo sounder for use in detecting fish near the sea-bottom (sponsored by the White Fish Authority) and to a narrow-beam scanning sonar for forward search.

The use of non-linear acoustic transmission in scanning sonar needs a little explanation. Normally, a within-pulse electronic sector-scanning sonar insonifies (on transmission) the whole sector, while a narrow receiving beam is scanned across it once in each interval of time equal to the pulse duration. To exploit the non-linear effect in a scanning system, however, means doing the scanning on the transmitted beam while using a wide receiving beam. The difference-frequency beam (f_1-f_2) discussed above is easily deflected away from its normal direction by the use of a sectionalized transducer and suitable phasing networks at the high frequencies (f_1 and f_2). It is then necessary to code the signals transmitted in the different directions so that, on detection, the direction of a target can be recognized and the echo displayed at the correct bearing. The simplest way to do this is to use different frequencies for different directions; other coding systems are being investigated. This coding, of course, demands a transmitting bandwidth greater than is necessary for the specified range resolution in a normal system; fortunately, as we shall see, bandwidth is no problem in this non-linear system.

8.2 *Wideband Transmission*

Most efficient underwater electro-acoustic transducers have a fairly high Q-factor, of the order of 5 to 10, and consequently can transmit only a small fractional bandwidth. Often this bandwidth is all that is needed, but for specially high range resolution evidently a wider bandwidth is often necessary. We have already mentioned the wider bandwidth required for transmit-beam scanning. Another system in which an extremely wide bandwidth is required is the wideband sonar[64] to be used for displaying the frequency response (as well as range and bearing) of a target over a bandwidth of ratio 10 : 1. The solution to all these problems is the use of non-linear interactions as described in the previous section.

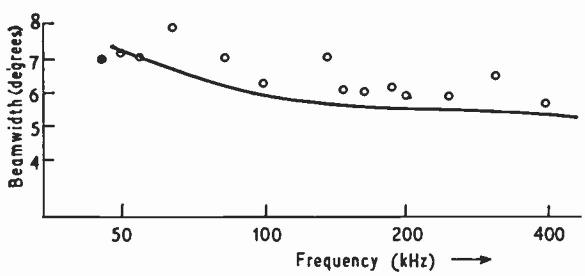


Fig. 6. Constant-beamwidth response of non-linear acoustic transmitter. Circles are experimental observations, line shows theoretical prediction.

In the non-linear transmitting arrangement, the high or primary frequencies f_1 and f_2 have to be confined within the relatively narrow fractional bandwidth (say 0·1 to 0·2) of the transducers. The difference frequency f_1-f_2 will however cover the same *actual* bandwidth in Hz as the primary frequencies, but being at a much lower centre frequency will have a very much wider *fractional* bandwidth. It has been found quite practicable to have f_1-f_2 varying over a 10 : 1 frequency ratio with almost constant level, using ordinary transducers for the primary frequencies. An experimental result is shown in Fig. 6.

A really wideband sonar has thus now become practicable, and initial trials of a breadboard system having been successful,[64] a complete serviceable equipment is now nearing completion.

8.3 *Parametric Reception*

Non-linear wave interactions can be exploited in receiving systems, but not quite so usefully as in transmitting applications. The system resembles in many ways the well-known parametric amplifiers used in electronic systems; and in principle both the up-convertor and negative-resistance types of travelling-wave parametric amplifier can be reproduced in the acoustic system.[65, 66]

The basic principle is that a small transducer launches a high-level pump wave at a fairly high frequency f_p. This transducer is, however, large enough in terms of wavelength to produce an approximately collimated beam. If a comparatively low-frequency signal wave f_s arrives from a direction identical with that of the pump wave, it interacts with the latter continuously along the beam and generates sum and difference frequencies f_p-f_s and f_p+f_s in suitable phase for their propagation in the same direction. In the up-convertor type of parametric receiver, a receiving transducer is placed at a suitable point along the beam and selects either f_p-f_s or f_p+f_s as the desired signal. This can not only have a higher intensity than the original signal at f_s due to the parametric amplification, but also is very sensitive to the direction of the original signal, giving maximum output only when its direction coincides with that of the pump beam, and giving smaller outputs at other directions according to a directional pattern which is of normal appearance (i.e. with side-lobes) but has a beamwidth far narrower than would be appropriate to the transducer sizes if used directly at f_s. A typical experimental result[60] is this:

Given: f_p 2·85 MHz; f_p+f_s selected as output signal; f_s 33 kHz; λ_s 4·5 cm; transducer dimensions 3 cm × 3 cm; distance between pump and output transducers 1·75 m.

Found: 3 dB beamwidth: 14°. Parametric gain: 8 dB.

The pump power was not known exactly, but was of the order of 40 W.

The up-convertor type of parametric amplifier, as described above, is readily made and works well. The negative-resistance kind is more difficult. It works on the principle that in the pump beam the difference-frequency wave at $f_p - f_s$ interacts again with the pump wave to produce an additional signal wave which augments the original one and produces parametric gain at the signal frequency; in the electronic circuit using varactor diodes this is explained most easily in terms of the formation of effective negative-resistance elements—hence the name used. Unfortunately, the sum-frequency wave at $f_p + f_s$ interacts with the pump wave to produce an opposite effect. In the lumped electronic circuit it is easily arranged that the cut-off of the system lies below $f_p + f_s$ and thus this deleterious effect is removed. In the acoustic system this is possible only by the insertion of acoustic filters between the pump and the receiving transducer, and this is difficult and unattractive. Thus no future is foreseen for the negative-resistance acoustic parametric receiver.

These parametric receivers provide a relatively wide bandwidth of response as in the non-linear transmitters described above.

9. Propagation

It is quite impossible to give anything approaching a useful review of the vast subject of underwater acoustic propagation within the confines of one section of a general paper, and the task will therefore not be attempted here. It is proposed to refer only to one or two topics which have specially appealed to the author. However, an excellent detailed and substantial treatment of many of the fundamental problems of underwater acoustic propagation has been recently published by Tolstoy and Clay[67] and there have been numerous relevant papers in the *Journal of the Acoustical Society of America* and elsewhere.[68–75] An interesting treatment of the effect of fish (and especially their swim-bladders) on propagation has been given by Weston.[76]

A review of some recent work on scattering phenomena has been given by Welsby[77] based largely on work done at the University of Birmingham, and this has led to a further programme of experimental work. One of the interesting ideas which has come from this is the use of the receiver of an electronic-scanning high-resolution sonar equipment to study simultaneously both direction-scatter and time-scatter of signals transmitted through water. Providing its resolution is adequate, this is an ideal method since the usual B-scan display with its rectangular co-ordinates of range (i.e. time) and bearing (i.e. direction) is just what is needed for simple presentation of the observations. Some detailed results are given by Welsby and Gazey,[78] and one particular display photograph is reproduced in Fig. 7. Here the vertical axis is direction and the horizontal axis is time. The signal was a short pulse transmitted from a distance along the normal axis of the receiver. The first pulse displayed, on the central axis of the display, is of course that arriving by the direct path. Subsequent pulses displayed have travelled by longer paths involving reflections at the surface and/or bottom and perhaps also involving refraction and scattering due to turbulent patches in the water; therefore they not

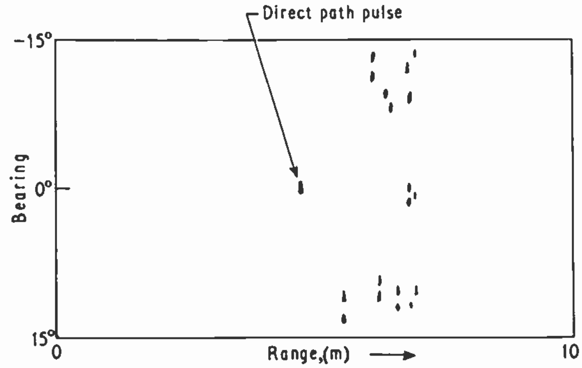


Fig. 7. Time and angle scattering of pulse signal shown on display of scanning sonar.

only appear at a later time, but also arrive from different directions. The time dispersion depicted here is about 2 ms and the angular scattering about $\pm 12°$. The acoustic signal was at 500 kHz and the observations were made in a deep turbulent pool in a river.

In the above illustration it was forward scattering that was being studied. Backward scattering, usually referred to as reverberation (rather confusingly, since those concerned with the type of acoustics generally called seismics have to deal also with real reverberation of the same type as room reverberation in architectural acoustics) is even more important in many practical situations. It sets the limit to detection in many sonar systems whether of short-range or long-range type. The scattering is produced by a rough sea-bottom, by waves, by fish and other biological inhomogeneities, by ice, etc. Numerous papers have been published on it.[79–84]

It is worth noting, in passing, that some of the propagation effects which are difficult to examine with accuracy under water are more easily investigated in sound-in-air transmission because of the larger relative magnitude of the effects in air. An example which has been recently reported[85] is the effect of turbulence due, for example, to the effect of a gusty wind among buildings or to heaters and fans indoors. Here a 6 mile/hour wind caused a worsening of angular resolution from 1° (in still air) to 3° when a 100 ms smoothing was used to reduce amplitude fluctuations. It was shown, however, that if a within-pulse electronic sector-scanning receiver was used on a pulse sonar in air, so that angular resolution on individual pulses could be observed, this was found to be almost as good as in still air; the effect of the turbulence was to deflect the beam randomly and lead merely to incorrect direction on single pulses, but to greatly reduced angular resolution if several pulse returns were superposed. A mathematical analysis of the degradation of a beamshape due to random fluctuations in the refractive index of the medium is given by Bourret.[86]

10. Sound Sources and Receivers

Under this heading we must include the vast and important subject of electro-acoustic transducers which is so fundamental to the successful realization of any sonar system design and also to almost all underwater acoustic research. Unfortunately this subject is most inadequately represented in the literature of underwater acoustics, and the design of individual transducers for particular systems remains, in general, a rather empirical process in which hints and tips are passed on by word of mouth from one worker to another. The basic theory is covered in a few textbooks[87–89] and there are a few review papers[90–91] and occasional descriptions of particular transducer designs.[92–93] A useful review was given recently at a meeting of the British Acoustical Society and it is hoped this will be published.[94]

Most transducers use magnetostrictive, piezo-electric or electro-strictive materials, with the last-mentioned type predominating in modern equipments. These transducers are naturally resonant, and for efficient conversion of electrical to acoustic energy and vice versa have to be used at resonance. This means that such transducers have narrow frequency-bandwidths, commonly around one-tenth of the resonant frequency, and rarely exceeding one-fifth. This is adequate for many (and probably most) normal types of sonar system. Where a wider bandwidth is required, non-linear wave interactions can be used as described in an earlier section of the paper. A method much used hitherto is the operation of the transducers over a wide frequency range well below or well above resonance. This is very inefficient, and is rarely tolerable on transmission (except for special research purposes) because it is necessary to get the signal at the receiver above the ambient acoustic noise level. On reception the inefficiency is often tolerable at frequencies below, say, 100 kHz because the acoustic noise is so far above the electronic noise level in the receiver that a worsening of the latter due to an inefficient transducer has little effect on detection.

As a result of the situation discussed above, there has been some emphasis recently on developing other types of low-frequency sound sources capable of giving a wideband signal of high power. McCartney[95] has reviewed the situation for the frequency range 10 – 3000 Hz. Apart from electromagnetic sources such as the 'boomer',[96] a pneumatic source in which several cubic inches of air at pressures of the order of 2000 lb/in^2 (1·4 MN/m^2) are discharged into the water appears very promising.[97] Work currently in progress at the University of Birmingham is directed towards obtaining a coherent source of low frequency by the repetitive discharge of air into an elastic-walled sphere. The subject of low-frequency sources was given a full discussion at a recent meeting of the British Acoustical Society, of which a record has been published.[98]

There is a growing interest in the frequency range below 1000 Hz. Seismic investigations of the sea-floor, signal transmission in the SOFAR† channel,[67] and detection of fish by swim-bladder resonance,[95] all require this low-frequency range.

11. Arrays

Array theory is very largely common to both acoustic and electromagnetic wave applications, since it is concerned mainly with the relationship between directional effects and discrimination on the one hand, and the dimensions, shape and sensitivity function of the array on the other. A very great deal has been published on this subject, and in 1964 the I.E.R.E. and the University of Birmingham jointly devoted a large symposium to it.[99] Naturally the most prominent work over many years has been that devoted to optimizing the design of arrays, but whereas formerly the emphasis was on synthesizing array patterns of maximum main-beam/side-lobe ratio with uniform spacing of the elements of the array, more recently the emphasis seems to have shifted to obtaining a specified main-beam/side-lobe ratio with the minimum number of elements. Thus much study has been made of arrays with unequally-spaced elements.[100–102]

Superdirectivity has attracted a great deal of attention over many years, but its practical achievement has proved difficult. Since it involves the driving of adjacent elements in opposite phase in spite of spacings of less than half a wavelength there has always been a basic difficulty in practical design because of the considerable acoustic coupling between elements, and the bandwidth of the response of the array has been very narrow indeed because of the large reactive field which has to be tuned out. A method of overcoming these difficulties is to use decoupling between the elements and the point at which their terminals are commoned,[103] so that on reception the individual elements

† SOund Fixing And Ranging

operate as though the array were not superdirective, and on transmission the elements are constrained to the velocities or currents required by the design irrespective of their couplings. Figure 8 shows the arrangement for reception, where buffer amplifiers may be used for decoupling. On reception at any rate, this system necessitates a distinction between 'supergain' and 'superdirectivity', terms which have hitherto been usually regarded as synonymous. The buffer-amplifier system gives superdirectivity but not supergain.

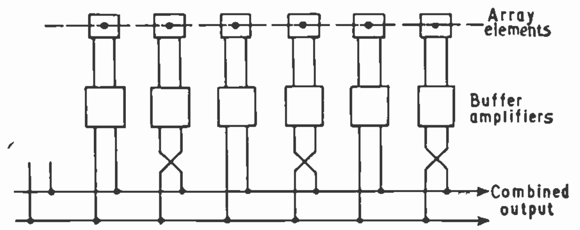


Fig. 8. Superdirective receiving array with buffer amplifiers.

Another topic which is basic in arrays but which has had only sporadic attention is the mutual coupling between elements and its effect on array performance. This has been mentioned above in connection with superdirectivity, but it has potentially serious effects on closely-packed arrays of small elements where it can cause negative radiation resistance in some elements,[104] and in phase-responsive digital systems where the correlation introduced by mutual coupling must worsen the detection performance of the system.[105] The calculation of inter-element coupling is difficult,[106] but it is hoped soon to develop a simple method of measurement.

Arrays with electronic scanning of the beam have been used in sonar for some time, and the within-pulse electronic-scanning sonars have been frequently referred to in this paper. A newer development is the circular array with electronic scanning.[107,108] This enables a beam to be swept around 360°. Techniques for the automatic (or adaptive) control of the position of zeros in the directional pattern have been developed[109] for radio use and may have application in sonar; similarly sonar applications can be envisaged for arrays which automatically return a signal along the direction from which it arrived.[110,111]

Most of the beam-scanning systems so far used merely scan the beam in one plane. The scanning of a pencil beam on a raster, or two-dimensional, basis is also possible, however, and a reasonably simple system has been worked out.[112]

12. Sonar Targets

The acoustic return from the various kinds of object which can be of interest as sonar targets is obviously of paramount importance in the design of a sonar system and in the assessment of its performance. Yet our knowledge of the mechanisms, magnitudes and fine structure of the echoes from even the simplest of targets—i.e. the sphere—is very incomplete. Some thorough work by Freedman[113–115] laid the foundation for a detailed investigation of the echoes from rigid bodies of certain simple geometrical shapes. He showed that echoes should be obtained, not from the body as a whole, but from all discontinuities in the projected area of the target along the range axis, and from all discontinuities in all derivatives of this area/range curve, although these latter give rise to much smaller magnitudes of echo than the former. If the pulse length is very much less than the dimensions of the target, there is a possibility of some of these echoes being resolved, and Freedman achieved a considerable measure of success in his experiments. By using high-frequency beams of very narrow angular width, by using very short pulses, and by providing a very large dynamic range in the equipment, further resolution of the echoes from simple shapes such as spheres, cylinders and rectangular blocks is now being achieved by the author's colleagues. Complications can arise when the targets are solid, since then waves enter the target, suffer mode conversion to and from shear waves, cause repeated internal reflections, and so produce finally an extremely complex echo structure. One particular case, that of a solid metal rectangular block, has been investigated in detail, both theoretically and experimentally, by Dunsiger.[116]

It is, of course, in principle only a direct analytical process to determine what echoes are received from a specified target. However, the real sonar problem is not this at all, but the vastly more difficult one of recognizing any given echo pattern on the high-resolution sonar display as coming from a particular object. It is just possible that if the angular and range resolution were great enough, and if the dynamic range of the system were infinitely large so that lower-order echoes from the derivatives of the projected area/range curve of the target could be distinguished, then an unambiguous classification of the target could be made. In practice it can never be possible to make an unambiguous classification, and if only a single aspect of the target is available, the ambiguity will be large. For example, the first-order echoes from a sphere and a cylinder viewed end-on are indistinguishable, and so are those of a cylinder and a rectangular block viewed sideways. But if it is possible for the sonar to be moved around the target (as it would be if the sonar were ship-mounted and ample time were available) then many ambiguities would be removed. This matter is under investigation at Birmingham.

We have been referring above to a limited class of target, the rigid solid of geometrical shape. Another kind of target of great importance is fish. The investigation of the echo responses from fish has been pursued by a number of people for many years,[117–120] but satisfactory and adequate knowledge of the subject is still far from being attained. At high frequencies where the fish is many wavelengths long, the echoes come from the bony structure and swim-bladder and to a smaller extent from the flesh, and a complicated diffraction effect leads to rapid fluctuation of target strength with frequency. It seems the mean target strength is not strongly dependent on frequency. At low frequencies where the fish is only a small fraction of a wavelength, the echo is only large when the sonar frequency is very near the resonance of the swim-bladder.[95, 121] The gross features of the variation of target strength with frequency are thus predictable, and it has for long been a hope that a very wideband sonar which could display the frequency response of the target strength would be a valuable tool for recognizing and distinguishing fish. As stated in Section 7.2 of this paper, such a sonar is now feasible and results are awaited with interest.

13. Miscellaneous Topics

A few other topics will now be briefly mentioned, because of their potential importance or interest.

13.1 *Ship's Noise*

It has become clear that the performance of echo-sounders used in fish detection falls far short of that calculated from the basic sonar equation on the assumption that sea-noise limits detection. For example, a calculated maximum range for the detection of a single cod of length 1 m with a particular echo-sounder is 1100 m with the noise specified as that due to sea-state 5; but this range is considerably more than twice that claimed as a maximum by the manufacturers, and so is very greatly in excess of the normal maximum range. The explanation of the discrepancy is almost certainly the noise of the ship itself (its machinery and propellor) reflected into the echo-sounder from the sea-bed. There is a great paucity of available measured data on the noise of trawlers (and trawls) and it is clear that this is a matter requiring investigation. Preliminary measurements[122] indicate that ship's noise in an echo-sounder can be 30 dB above normal sea-state noise.

13.2 *Acoustic Imaging*

Over the years there have been a number of efforts to develop an acoustic imaging device or 'acoustic camera' using a lens or focusing mirror.[123–126] It has no doubt seemed a potential advantage of such a system that it produces a display more directly comparable with an ordinary visual image than that of the usual sonar display plotted as bearing versus range. But the author feels that this is fallacious; we are nowadays thoroughly used to radar displays, charts and maps in various forms, and the advantage of an acoustic image cannot be large. Against the imaging system is the strong technical/economic fact that it is extremely costly and difficult, and operationally unattractive, to obtain bearing resolution in two axes to give the same number of resolution cells as the single-bearing/range resolution of the echo-location system. Range resolution by short-pulse techniques can be obtained so cheaply and easily compared with making a huge acoustic lens that it is difficult to see any future potential in imaging systems. Moreover, even if for some reason it is important to obtain a detailed transverse cross-section of the beam, it would still be much cheaper and simpler to use an echo-ranging system with electronically-scanned pencil beam obtained with a Mills cross array[112]—and this would give range information *in addition to* two-axis bearing resolution.

A development which may necessitate a revision of the above conclusions is that of acoustic holography. Since this does away with the need for a lens, the fundamental objection to imaging systems is removed; the difficult but not fundamental problem of the image converter remains of course. A good deal of attention has been paid to acoustic holography recently[127–129] and practical developments are awaited with interest.

13.3 *Detection by Field Distortion*

It is possible, in principle, to detect the presence of objects in an acoustic field by measuring distortion of the field in terms of pressure *and* particle-velocity or displacement changes—or, indeed, in terms of the disturbances in the acoustic wave impedance at various points. This would mean having both pressure-responsive and velocity- or displacement-responsive transducer probes at various points in the field. Very low frequencies would be used so that the wavelength is large compared with the dimensions of the observed volume. It is quite likely that this is the system used by some marine animals for exploring their environment. Welsby[130] has examined this matter in outline, and detailed research is now proposed.

13.4 *Animal Acoustics*

A final topic justifying mention is the general field of animal acoustics and animal sonar. Marine animals make noises, respond to noises, and occasionally use sonar (e.g. the porpoise[131]). Conferences have been held to discuss these topics and a great deal of information appears in their Proceedings.[132, 133]

14. Conclusions

No attempt will be made to draw technical and scientific conclusions from this paper except to suggest that certain important trends are discernible. One is the likely importance of digital circuitry and micro-electronics in the sonar of the future, and another is the new scope offered by the exploitation of non-linear acoustic wave interactions. As the subject of the review is underwater acoustics, no review has been included of the vast amount of work done in recent years under the heading of signal processing, but this work is already having an influence on sonar system design, especially in the naval field.

There has been a great expansion of interest in underwater acoustics and sonar systems in the last decade. The author produced a review paper[134] on 'Underwater Echo-Ranging' for the Institution in 1956—twelve years ago—and it is most interesting to see not only the great developments that have taken place since then, but more particularly the way in which the subject has come out into the open with the growth of civil applications.

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STANDARD FREQUENCY TRANSMISSIONS—January 1969

(Communication from the National Physical Laboratory)

Inclusion of Relative Phase Values in Results of the Standard Frequency Transmissions

From the 1st January 1969 the results for the standard frequency transmissions will include the relative phase differences between the UTC and AT scales derived from the N.P.L. operational caesium standard and the received phase of GBR 16 kHz and MSF 60 kHz, respectively. The phase values will include all phase variations which can be properly attributed to the transmission although in the case of phase discontinuities these will normally be removed in order to obtain the tabulated figure for the mean daily frequency deviation.

In agreement with other publications the convention adopted is to give the difference (N.P.L.–Station) so that an increase in the quoted difference indicates that the phase of the received signal is retarded relative to the N.P.L. standard and consequently the received frequency is low. The difference (N.P.L.—GBR 16) is chosen conventionally to be +500 on 31st December 1968; the corresponding difference for (N.P.L.— MSF 60) of + 468·6 is determined by a previously defined epoch at 31st December 1967.

January 1969	Deviation from nominal frequency in parts in 10^{10} (24-hour mean centred on 0300 UT)			Relative phase readings in microseconds N.P.L.—Station (Readings at 1500 UT)	
	GBR 16 kHz	MSF 60 kHz	Droitwich 200 kHz	*GBR 16 kHz	†MSF 60 kHz
1	− 299·9	− 0·1	+ 0·1	504·0	468·7
2	− 300·1	− 0·1	+ 0·1	529·5	469·4
3	− 300·1	− 0·1	+ 0·1	530·0	470·3
4	− 300·0	− 0·1	+ 0·1	530·0	471·2
5	− 300·1	—	0	531·0	—
6	− 300·0	—	0	531·0	—
7	− 300·0	—	0	531·0	—
8	− 300·2	− 0·1	− 0·1	533·0	471·7
9	− 299·9	0	0	532·0	471·8
10	− 299·9	+ 0·4	0	531·0	468·1
11	− 299·9	0	0	530·0	463·2
12	− 300·1	+ 0·1	0	530·5	462·4
13	− 300·0	− 0·1	− 0·1	530·5	463·0
14	− 300·0	0	− 0·1	530·5	463·0
15	—	0	− 0·1	—	463·0
16	—	0	− 0·1	—	463·2
17	− 300·2	− 0·1	0	—	464·4
18	—	0	− 0·1	—	463·4
19	—	0	0	—	463·1
20	—	0	0	—	462·7
21	− 300·0	0	0	530·5	462·7
22	− 300·1	+ 0·1	+ 0·1	531·0	461·9
23	− 299·8	+ 0·1	+ 0·1	529·5	460·9
24	− 299·9	+ 0·1	+ 0·1	529·0	460·3
25	− 299·9	0	+ 0·1	528·5	460·0
26	− 300·0	+ 0·1	0	528·9	459·3
27	− 299·9	0	+ 0·1	528·3	459·4
28	− 300·0	+ 0·1	0	527·0	458·6
29	− 299·9	0	+ 0·1	526·5	458·3
30	− 299·9	+ 0·1	+ 0·1	524·5	457·6
31	− 300·0	0	+ 0·1	523·5	457·5

All measurements in terms of H.P. Caesium Standard No. 334, which agrees with the N.P.L. Caesium Standard to 1 part in 10^{11}.
* Relative to UTC Scale; (UTC_{NPL} — Station) = +500 at 1500 UT 31st December 1968.
† Relative to AT Scale; (AT_{NPL} — Station) = +468·6 at 1500 UT 31st December 1968.

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Progress in On-line Control by Computer

By

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Reprinted from the Proceedings of the I.E.R.E. Convention on 'Electronics in 1970s', held in Cambridge on 2nd–5th July 1968.

Summary: The paper presents an account of the development of industrial process control by computer. Some of the control functions carried out by computer are described, with examples from a number of industrial processes.

The special features of digital equipment for on-line industrial use are examined. In particular, reliability and fail-safe aspects are emphasized. Some of the methods used and problems encountered in system design and programming activities are illustrated.

1. Introduction

The purpose of this paper is to review the use of digital computers in the control and operation of industrial processes. In the normal use of computers, data prepared manually are fed into the machine which then performs a defined task of processing the information, finally presenting the required output data in a form to be read by a human operator.

This process has two particular features which distinguish it from the type of application considered here. Firstly, it is a batch process. The data are first collected together for input to the machine, the process is run at a convenient time and the output distributed. Secondly, both input and output data require human operation as an essential part of the process.

In contrast, the applications considered here require operations to be carried out at times dictated by the state of the external process, not by the load on the computer, so that the machine must be ready at all times to undertake any one of a number of tasks. Also, data are collected by the machine directly from measuring instruments on the plant and output command signals operate directly on valves and switches. No human intervention is involved.

These features require techniques and equipment which are different from those used in data processing, although, of course, there are many common requirements.

2. Growth of Industrial Systems

This way of using computers requires that they should be available instantly when required, and this implies a high standard of reliability and high operating speed. The first requirement became practical with the use of transistors. The earliest applications were to military systems, and some of the earliest work was done in this country. A good example is the use of computers for preparing and launching guided missiles. This is a complex task involving assessment of the threat, selecting, directing and preparing the missile. It has to be carried out precisely and very quickly. Computers are essential to achieve the necessary speed with minimum manpower. Many systems are now operating throughout the world with a very high standard of availability.

Industrial applications first took place in the early 1960s. In many cases these were not truly control applications, but were mainly used for data acquisition and logging.

One of the first installations in which the computer carried out the whole task of plant control was started in 1962 in this country by I.C.I. By today's standards this was a comparatively simple application, but it has achieved a significant improvement in plant performance, and a high standard of availability.

The rate of expansion since these early days is shown in Fig. 1 which represents installations throughout the world. The picture for the U.K. is also shown in Fig. 1. The rate of growth for this country has been about the same as the overall growth, and we have about 10% of the total number of installations.

Table 1

Percentage distribution by industry

	May 1962	March 1967
Power	44	20
Chemical and oil	27	25
Metal	14·5	17
Others	14·5	38

A broad analysis by industry is shown in Table 1 which compares the distribution of installations over a five-year period during which the total increased by a factor of ten. As might be expected, this shows an extension of applications into new industrial fields.

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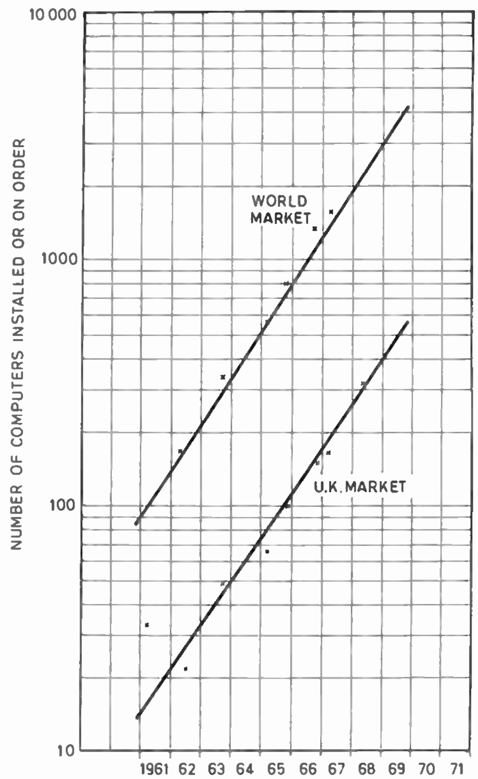

Fig. 1. Growth of U.K. and world market for process control computers.

A growth rate of about 60% per year has been sustained to date. It is difficult to predict the future prospects but 40% per year growth may well be possible for several years. If this turns out to be true, it will very soon tax the available engineering effort, and this seems likely to set the limit to the rate of expansion. This is an activity which has a particularly high content of engineering service. Each installation must be designed to meet the needs of a particular plant, and no two are alike. Each requires an investigation of plant behaviour which may be quite extensive. Engineering staff with the necessary experience are neither readily available, nor rapidly trained.

3. Applications

There is a very wide range of functions to which control computers have been applied, and many of these extend beyond the usually accepted range of closed-loop control. This section describes a few of the ways in which computer-control methods can produce economic benefits.

3.1. *Direct Digital Control* (*D.D.C.*)

A simple d.d.c. system consists largely of separate control loops to control variables such as temperatures, pressures and flows. The computer program examines the value of each controlled variable in turn, compares it with its desired value and calculates the necessary change in demanded valve position. This change is then sent to the output equipment to update the valve position.

In this simple form, the computer is used to replace a number of conventional analogue controllers carrying out the same basic functions. However, it offers considerable advantages.

To appreciate the control improvements that are possible by using the flexibility of computer control systems, let us look briefly at conventional analogue control system as shown in Fig. 2. The difference, if any, between a desired process value D and a measured process value M is called the error E. The controller output which includes adjustable proportions of error derivative and integral moves the control valve in such a way as to bring the error E to zero. Thus a desired value of flow, level, pressure, temperature (or whatever is being controlled) is set by manual adjustment of a lever or knob. The desired value D is compared with a measured value signal M, and the control action attempts to keep M and D equal. To change the desired value subsequently, the operator must intervene.

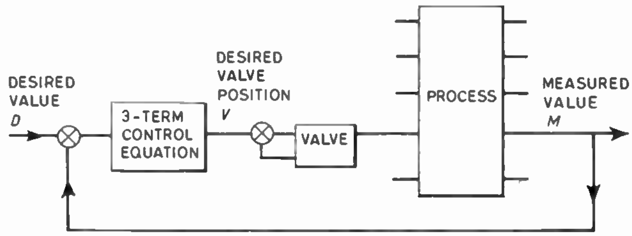

Fig. 2. Simple 3-term control loop.

With computer control, a digital equivalent of the controller can be implemented, with the following variations and advantages:

The desired value D_c used by the computer can be:

(a) a constant value stored in program,

(b) a value which can be set from an operator's control console,

(c) a value calculated by computer program, e.g.
  (i) output of another calculated control action, providing for cascade control, and ratio control.
  (ii) from a mass balance, heat balance or economic calculation,

(d) used to alter the value of the controller settings as throughput varies.

The measured value M_c used by the computer can be either the plant measurement M 'directly', i.e. after it has been converted into an equivalent digital number by the input equipment, or after this digital number has been further modified by programmed calculation as follows:

(a) linearized, e.g. for resistance thermometer signals,
(b) square-rooted, e.g. to convert differential pressure to flow,
(c) compensated for cold junction temperature variations for thermocouples,
(d) integrated, e.g. from rate-meter inputs,
(e) smoothed or averaged by numerical techniques when signals are noisy,
(f) checked against the last measured value for acceptability or to give alarms on rates of change rather than absolute values,
(g) combined with measured values of other physical or chemical properties to define functions which cannot be measured directly.

The error signal can be operated on by digital equivalents of the controller or by many alternative control equations, in order to control process valve positions.

3.2. *Non-interacting Control*

Conventional analogue control schemes consist mainly of separate feedback control loops, although occasionally pairs of loops are connected together for ratio or cascade control. Analogue control schemes typically implement lower levels of control of physical properties such as flows, pressures and temperatures. As the level of performance index rises to include measures such as yield, production rate or quality or profit per hour, it may become necessary progressively to increase the number of variables controlled. When intentionally altering a product flowrate, one often unintentionally alters other properties such as temperature or composition as well. As the number of variables increases, the problem of interaction increases, and to eliminate this often requires more interconnection between control loops than is conventionally used. The wider range of control actions made possible by computer control makes it easier to tackle this problem.

An ideal to aim for would be to have one control which could alter a product-flow without affecting its composition, another to alter the composition independently of the flow and so on. One could then alter these independent controls in turn until the profit was maximized. Such a system would be non-interacting except for its desirable interaction with profitability. Since processes do not normally behave in this way, the next best thing is to find out which combination of control adjustments carried out either simultaneously or with correct time relationships would give the same result as a non-interacting system. In principle, non-interacting control systems can be designed providing we know enough about the process interactions and providing we can measure the right properties. In practice, inadequate knowledge and inadequate measurements mean that we can only approach the ideal. Nevertheless, the principles of non-interacting control systems suggest that control loops should be more interconnected than they are in conventional control schemes.

A good example where a non-interacting system can improve control is on a crude oil distillation column in a petroleum refinery, where up to six product streams may be taken from one column. With present control systems, control actions taken on one side-stream affect properties of adjustment side-streams. At times of process upsets, or during planned changes of product split or of crude oil to be refined, operator actions occur as a series of successive approximations. By computer they can be carried out together, and the new conditions can be quickly established.

3.3. *Conditional Action*

Much of the power of computer control systems derives from their ability to follow one of several alternative sequences of actions, depending on their evaluation of process conditions at a given time. Examples of this capability are:

(a) If the process is to be controlled at certain steady conditions, use control system A. If the process conditions have to be changed quickly to other values, use control system B during the change, then use control system A for the new steady conditions. This facility can take advantage of the fact that the best control system for steady-state control does not always allow rapid changes to be made.

(b) Stand-by equipment can automatically be brought into service on plant failure. Programmed start-up and shut down of plant is also possible, e.g. catalyst regeneration procedures in chemical or petrochemical plants.

(c) At a lower level with a process control scheme, there may be advantage in keeping an off-take flowrate equal to a desired value only so long as a measured pressure lies between safe limits. If the pressure reaches a high or low limit, the flowrate could then be varied to keep the pressure constant.

3.4. *Tracking*

Many production processes require a sequence of operations to be carried out on products which are similar but not identical, and therefore require different processing. The problem of keeping track

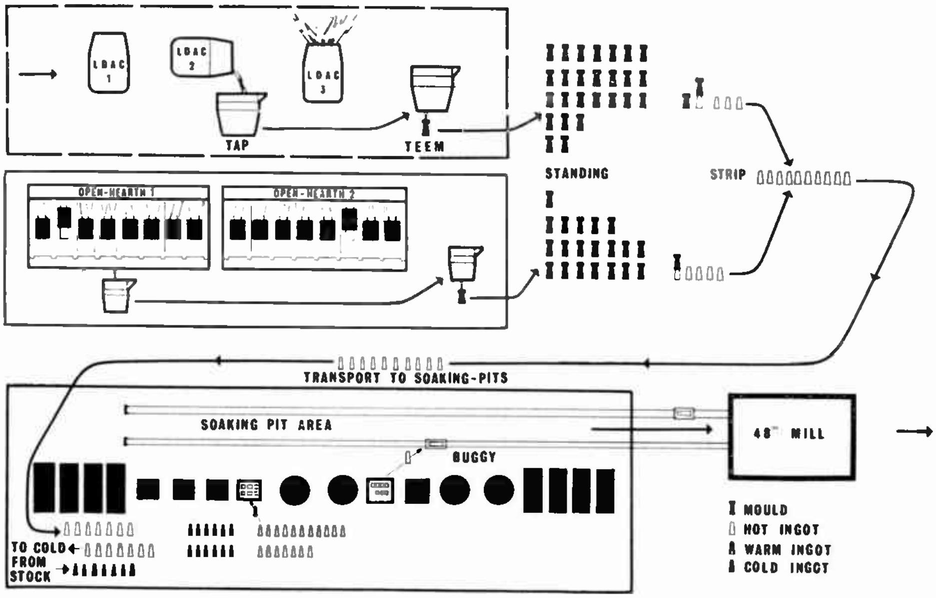

Fig. 3. Soaking pit problem.

of each item on its progress through the plant, and ensuring that it is correctly processed can be complex. The following example of soaking pit operation illustrates the problem (Fig. 3).

Steel ingots are poured into moulds, the moulds stand for various lengths of time until the ingot solidifies, then the moulds are stripped off. The ingots cool in air, again for various lengths of time, and are transported to the soaking pits. Here they may be joined by cold ingots from stock. The problem is to predict the length of time that an individual ingot should be in the pit in order to reach the correct heat content for the subsequent rolling processes. The past history of each ingot is different, i.e. different pouring temperatures, times in moulds and track-times. To arrive at a correct solution, the progress and past history of each ingot must be continuously monitored, and the effect of heat content calculated for various conditions—for example, (a) ingot and mould cooling in air, (b) ingot alone cooling in air, (c) ingot heating in the soaking pit. Proper scheduling of the process results in ingots being presented for rolling in a uniform condition, with consequent improvement in the quality and throughput.

4. Plant Analysis

The examples given in the previous section illustrate two points. The first one is that it usually turns out that the most substantial economic gains result not so much from highly technical considerations of control optimization as from theoretically straightforward 'house-keeping' operations. Once full advantage has been taken of these possibilities it is often doubtful whether the further gains theoretically achievable by optimization would be worth the considerable effort involved in putting it into practice. The great amount of academic attention which this subject receives seems to be out of proportion to the potential value to industry. This is not to say that the real problems are easy to solve or intellectually unrewarding. The second point is that a satisfactory system design can only result from a very detailed study of plant operating conditions and economics. This requires very close collaboration between process engineers, plant operators, control specialists and computer experts.

There is, as yet, no firmly established practice on the question of who should carry out the work of system design. It is difficult to say whether it should be carried out by the user, the manufacturer, or whether this service is best performed by independent consultants. Many of the larger users are well-equipped, and have the necessary staff and experience to do this work for themselves although there are always benefits from close collaboration with the manufacturer, and this is usually done. However, a great many users who could benefit from automation systems are not in this position, and the task must be carried out either by the manufacturer or by a consultant. The author has no doubt that the best result is achieved if the manufacturer is able to provide this service. Programming and system organization is part of the total design problem, and it is an unnecessary and complicating restriction to separate them from equipment provision.

The manufacturer providing this kind of service to industry will design and manufacture equipment specifically for industrial automation, and have available supporting engineering teams with experience in computer techniques, control system design and industrial processes. These are not necessarily different people. Working in the same team, many engineers quickly acquire expertise in fields outside their basic training, and this integration greatly strengthens the overall capability.

5. Programming

The task of programming for industrial control systems is, of course, a major part of the total design process.

Although programming methods have advanced considerably since the early days, there is still a long way to go. The programming languages commonly used in data processing and scientific applications are not well-suited to systems in which timing is all important. Consequently, programming is a task for the expert. This is a considerable handicap to the plant engineer.

The ability to make large or small alterations to the system configuration by means of program changes is one of the most important facilities for plant operations offered by computer systems. The development of more suitable programming languages such as RTL (Real Time Language†) will help considerably, but the services of a programmer will still be necessary and the need for an interpreter between the engineer and the plant greatly diminishes the ease with which changes can be made in practice.

One solution to this problem is offered by the CONSUL system. The object of this system is to enable the plant engineer, by operating a control panel, to make quite major changes to the system configuration. Indeed, it is possible in principle to start with no connection to the plant, and gradually to build up a complete scheme entirely by operation of the panel with no programming at all. This procedure is guided by responses from the computer (Fig. 4).

For example, when inserting a new loop into a control program, the engineer keys NWLP (for new-loop) on the keyboard, followed by the loop identity on special keys. This combined keying sequence initiates the conversational mode of operation. The computer replies by typing requests for information about the main control loop parameters. These requests, which are also given in mnemonic form, remind the engineer to supply, in the correct sequence,

† British Computer Society's Specialist Group, 'A language for real-time system', *The Computer Bulletin*, **11**, pp. 202–12, December 1967.

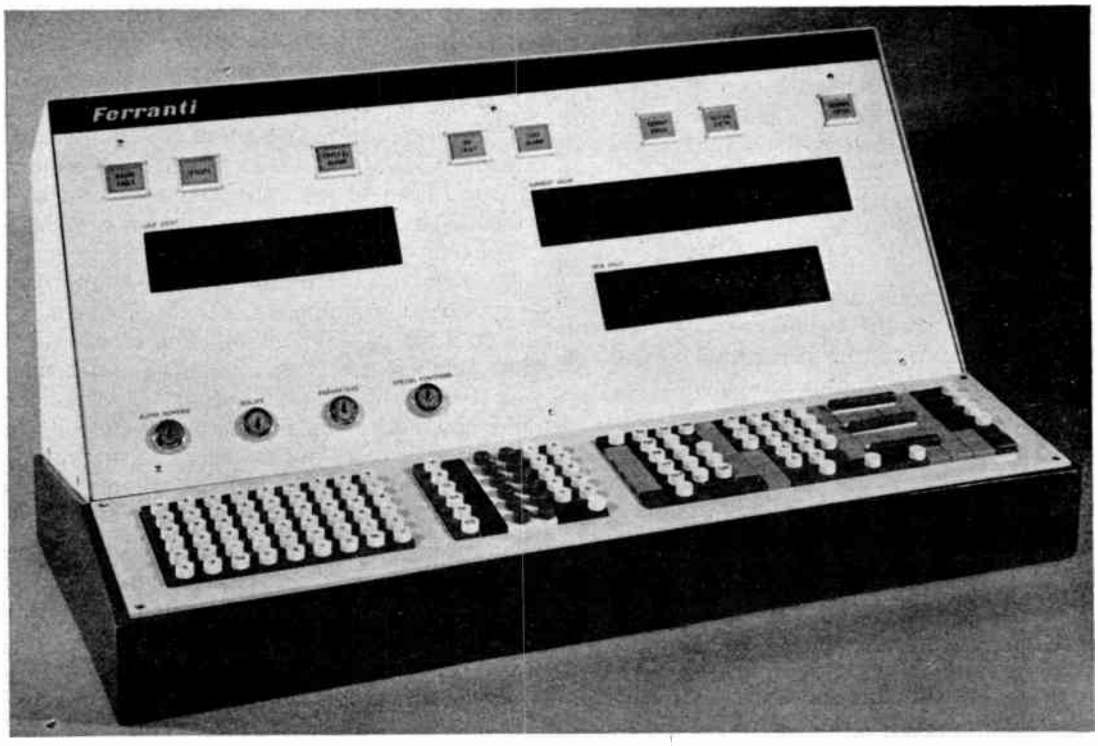

Fig. 4. CONSUL panel.

the necessary information relating to parameter scaling factors and engineering units.

On completion of this first stage, the engineer uses the keyboard to call for the various control functions or algorithms he wishes to use. The computer, in return, requests information on the inputs and outputs as appropriate to the selected algorithm. This information is supplied by the engineer using the most convenient form for reference. For example, the input to one algorithm can use the output of a previous algorithm, a specified decimal constant or the contents of a core store location holding the required data, which is identified by the address of that location.

This procedure is repeated for each remaining algorithm until all algorithms and associated information in the calculation block have been inserted. The engineer signals the end of the insertion sequence by keying the mnemonic FINI.

The control algorithms used are not confined to replacements for control sub-routines in conventional d.d.c. programs. Functions such as control, analytical sequencing and alarm functions are also included.

Non-linear algorithms provide for input linearization and the requirements of non-linear control. Functions such as square-root for linearizing a differential pressure measurement of flow and a signed square (input multiplied by modulus of input) for operating upon the measured value deviation, are included in this category to give a non-linear gain characteristic.

Dynamic algorithms not only provide digital equivalents of conventional analogue control actions—integral only, proportional only, and two- and three-term control—but also provide the digital equivalents of basic systems filters—simple lag, lead/lag and time delay for use in synthesizing more complex control schemes.

Other algorithms perform switching functions such as bumpless transfer cascade switching and general purpose control action switching, and provide routines associated with the generation of suitable control demands for the output equipment.

Facilities are built into the system to enable an engineer to try out experimental systems on-line easily and safely. For example, it is possible to introduce an experimental configuration in parallel with an existing arrangement so that the two may be compared. Operating limits may be set so that the plant returns to the original system if these are exceeded. Systems like this give the plant engineer great freedom to operate the plant in many different ways.

6. Equipment

In equipment design also, the aim has been to provide units which fit into the industrial environment, and are easily handled by engineers. A feature of industrial systems is the large number of input and output signals.

The numbers and types of inputs, outputs and displays vary considerably from process to process, as is shown in Table 2.

Table 2

Equipment requirements in typical processes

Process	Analogue in-puts	Analogue out-puts	Digital in-puts	Digital out-puts	Prin-ters	C.r.t. dis-plays
Chemical plant	300	100	200	20	2	0
Electricity generating station	4000	50	10 000	200	8	20
Steelworks	0	0	500	200	4	20
Electricity distant control centre	0	0	500	200	2	14
Laboratory system	2	2	32	0	1	1

The numbers and types of input and output signals which the input-output equipment must handle vary widely according to the process, and consequently the input-output equipment itself must also vary considerably in size and complexity.

On the one hand, therefore, there is the requirement for flexibility in the design of the equipment. On the other hand there is the need for minimum special design, both to reduce design costs and to facilitate maintenance, for a standard design is easier to maintain than a special. These two requirements can be reconciled by the use of a modular approach in the input-output equipment. The equipment is built up from a set of standard building blocks to deal with the numbers and types of inputs and outputs required by the particular application; the specification of the computer itself is determined only by the speed required to carry out the specified program functions in the time allowed, and the amount of core store needed to hold programs and data.

This has been a basic feature of the *Argus* range of equipment, which has been designed specifically for industrial use. The basic scheme is shown in Fig. 5.

The signals to and from the computer are carried on a standard interface of 108 signals. All connections to input, output and peripheral devices are made through this channel. Beyond this, a sub-division is made according to the type of signal and method of scanning,

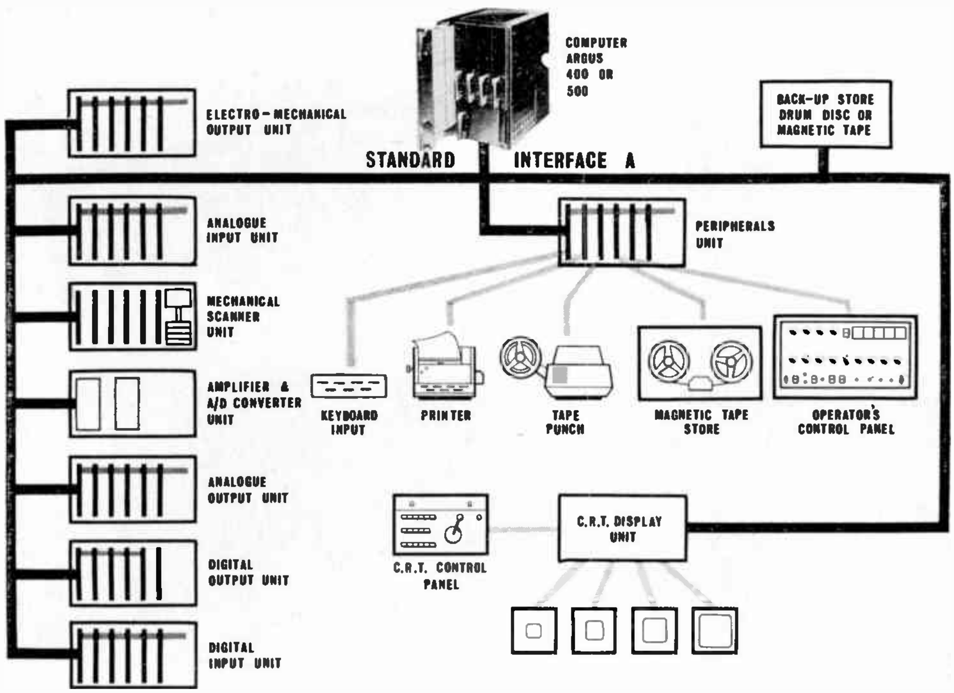

Fig. 5. Standard range of *Argus* micro input output equipment.

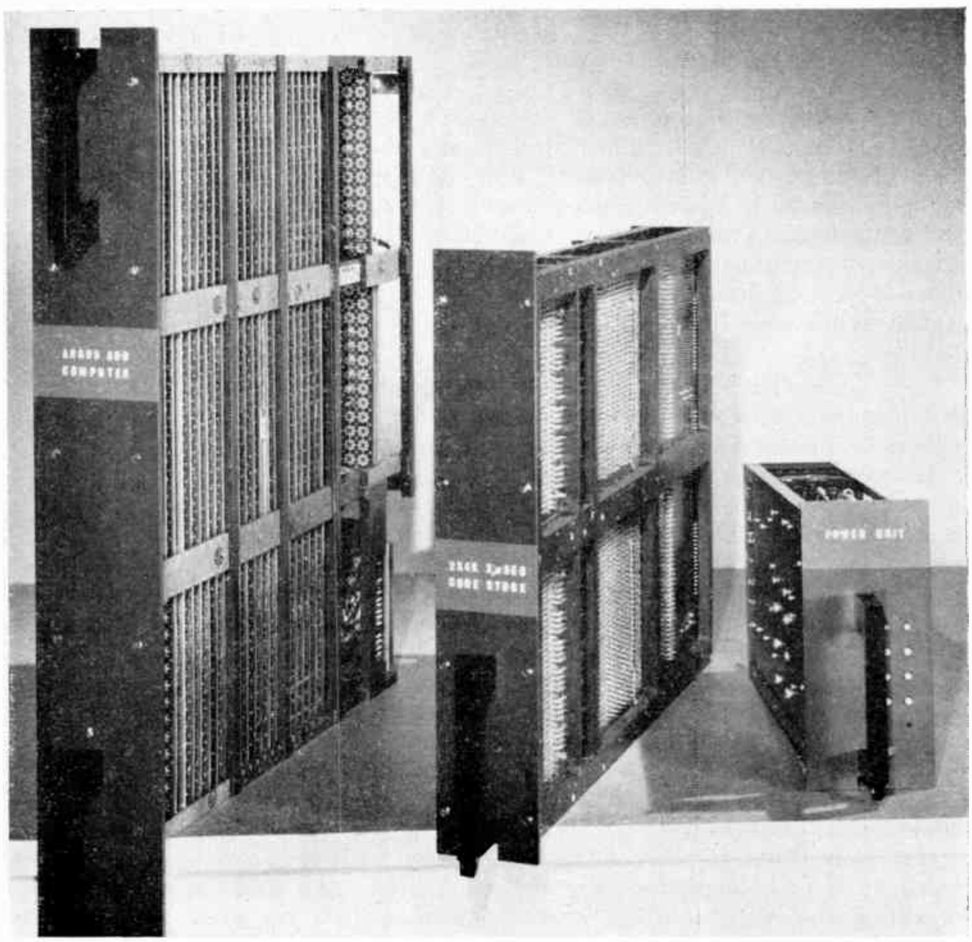

Fig. 6. *Argus* modules.

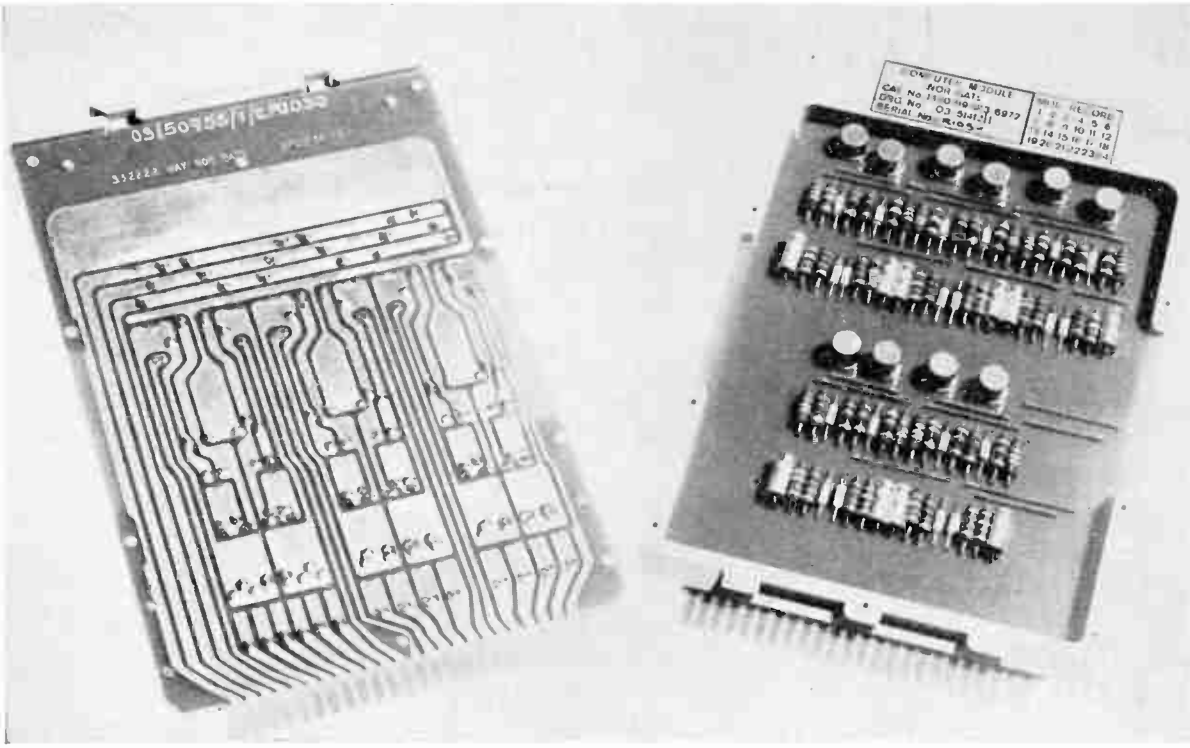

Fig. 7. *Argus 100* card.

and a subsidiary interface produced. On this may be connected a number of selection channels to meet the needs of the particular installation. It is usual to provide a number of spare locations of each type, so that configuration changes can be made simply by plugging in additional selection units of the appropriate type. Connections to the plant are taken straight from connectors on the selection units themselves.

This principle has been extended to the processors and store units. Both are arranged as plug-in interchangeable units. Store capacity may be changed from 4000 to 61 000 words by plugging in extra units, and both processors in the range are compatible with all other modules and interchangeable with each other (Fig. 6).

6.1. *Maintenance*

This equipment, like most digital systems now being designed, makes extensive use of integrated circuits. One consequence of this has been a radical change of concept in the matter of repair and maintenance. The previous generation of equipment usually consisted of a number of plug-in units each performing relatively simple logical functions (Fig. 7). A complete equipment contained large numbers of a few types of unit, a great many plug-in connectors, and a mass of interconnecting wire as shown in Fig. 8. Although chang-

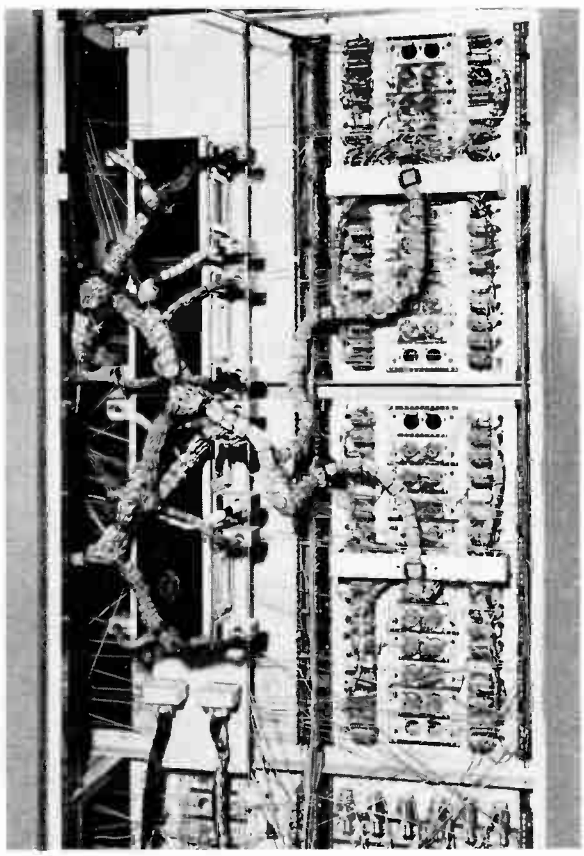

Fig. 8. Plug-in connectors and associated wiring.

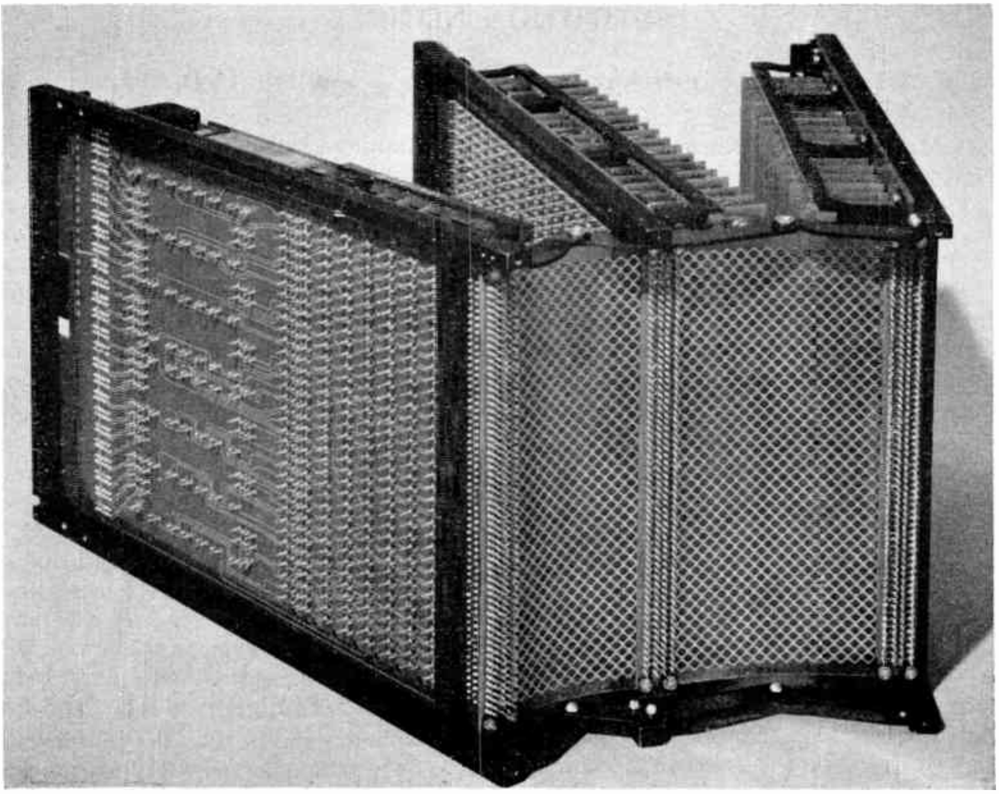

Fig. 9. *Argus 400* processor and store.

ing a faulty unit was easy and inexpensive, locating it was not always straightforward, and frequently faults were associated with the interconnections rather than the packages. In designing the *Argus* range, we considered that the reliability of integrated circuits no longer justified this subdivision into small replaceable units. As a consequence there are no plug-in packages in the processors, and interconnecting wiring is almost completely eliminated in favour of multi-layer printed circuits as shown in Fig. 9. The replaceable unit is now a complete processor or store. The physical size and robustness of the unit makes it quite practical for the maintenance engineer to carry a spare unit, and of course diagnosis to this level is straightforward. The expected mean time between failures of about 5000 hours makes this an economic and quick method of maintenance.

7. Future Trends

The effectiveness of computer control is now firmly established in the continuous process industries. As more and more systems come into operation successfully, confidence is being established to the extent that it becomes accepted that plans for new plant should include an on-line computer. At the same time, the range of applications is extending to other industries.

There is also an inevitable trend towards much wider integration of plant control systems with comprehensive management schemes embracing data processing functions as well as physical control of manufacture. There are clear advantages to be gained from this close coupling of day-to-day plant operation with marketing, supply and economic factors. However, it is not without its dangers. Computers can suffer from the same ailments as other organization systems. Attempts to centralize the control of too many functions can result in a rapid expansion of overhead type operations which soon become unwieldy, inflexible and very expensive. This effect is very familiar in some administrative organizations. The cure is usually to establish autonomous sub-units with the minimum amount of centralized decision making. In the author's opinion, this solution is also effective for large integrated computer systems.

Manuscript received by the Institution on 4th June 1968. (*Paper No. 1240/Comp. 112.*)

INSTITUTION NOTICES

Meeting of I.E.E. and I.E.R.E. Presidents

Following an initiative taken by Admiral of the Fleet the Earl Mountbatten of Burma, who is an Honorary Fellow of the Institution of Electrical Engineers and of the Institution of Electronic and Radio Engineers, and was Charter President of the I.E.R.E., the President, Secretary, and two senior members of each Institution have met to discuss the extent to which the two Institutions can integrate their organizations and activities in the interests of their members and of the profession and industries which both Institutions serve. This first meeting, held at 9 Bedford Square on 8th January 1969, was conducted in a most friendly and co-operative spirit which encourages the hope that arrangements of real significance will ultimately emerge.

There is now increasing collaboration between the Institutions in learned society activities through joint committees and members' meetings and joint sponsorship of conventions and symposia. A first step in working together outside Great Britain was taken a year ago in the arrangement for the I.E.E. to share the offices and administrative services provided by the I.E.R.E. in India. This year, the 'Indian Proceedings' of the I.E.R.E. is to become the 'Joint Indian Proceedings' of both Institutions.

The discussions now taking place are to consider the advantages that could accrue from a closer integration of administrative services, while both Institutions continue to operate under their respective Charters, and to seek an even closer association than at present in the field of radio and electronics.

Paris Components Exhibition

For the fifth successive year the Institution will have a stand at the Salon des Composants Electroniques, to be held at Porte de Versailles, Paris, from Friday, 28th March to Wednesday, 2nd April inclusive. Institution publications in particular *The Radio and Electronic Engineer*, will be exhibited. Committee members of the French Section will be present at the stand during the week.

Canadian Division Activities

A paper on 'Trends in Complex Integrated Circuits' was read at a meeting of the Ottawa Section on 16th October 1968 at the National Research Council Headquarters, by Mr. A. Morton, a member of the scientific staff of Texas Instruments Inc.

On 24th October 1968, members of the Quebec Section visited the Canadian Pacific/Canadian National Data Switching Centre, which is one of the five centres forming the trans-Canada data transmission system.

Mr. G. D. Clifford

In association with past and present Officers of the Institution, the President, Major-General Sir Leonard Atkinson, presented the Institution on 14th January with a portrait of the Director and Secretary, Mr. Graham D. Clifford. Admiral of the Fleet the Earl Mountbatten of Burma and other Past Presidents, and a number of guests including Sir Godfrey Agnew, Clerk of the Privy Council, were present when the portrait by Aubrey C. Davidson-Houston was shown in the new Council Room.

Mr. Clifford was first appointed Honorary Secretary of the Institution in 1937 but for some years previous to that he had been closely associated with the work of the Institution. At the time he was appointed Secretary, the Institution had only a few hundred members; it now has nearly 14,000. In the same period the Institution has grown from a single centre in London to a world-wide organization which has fourteen Local Sections in the United Kingdom, and overseas Divisions or Sections in India, Canada, New Zealand, South Africa, Israel and France.

In presenting this portrait the Officers of the Institution not only thanked Mr. Clifford for past services, but also wished him every success for the future.

Management Techniques Group

Council has approved the formation of a new Specialized Group on Management Techniques. The Inaugural Meeting will be held on Tuesday, 18th March, at 6 p.m. at the London School of Hygiene and Tropical Medicine, when Dr. F. E. Jones, F.R.S. (Fellow), Managing Director, Mullard Ltd., will give an Address.

It is hoped that all whose work brings them in contact with management problems will take the opportunity of attending this and subsequent meetings of the new Group. A second meeting, on 'The Management Concept' will be held on Thursday, 24th April.

Airborne and Spaceborne Computers

The Joint I.E.E.–I.E.R.E. Computer Group Committee are arranging a colloquium on Airborne and Spaceborne Computers to be held at Savoy Place, London, W.C.2, on 21st March 1969. It is hoped that the colloquium programme will include papers on aspects such as systems and hardware design, environmental problems and applications.

Persons wishing to submit contributions are invited to write or telephone the colloquium co-ordinator, Mr. D. M. Taub, IBM (United Kingdom) Laboratories Ltd., Hursley Park, Winchester, Hampshire (Telephone: Winchester 4433).

U.D.C. 681.32

Some Electronic Logic Circuits for Serial-Parallel Arithmetic

By

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Summary: In one form of rapid binary multiplier the number which is to be multiplied is presented serially to a controlled shift register. The multiplier must be presented in parallel to obtain the binary product. It controls the function of the register so that the product is generated serially from it. A variant of this type of multiplier is described in which the control logic is regarded as a recognizable sub-system. A new design for a divider is also suggested which is comparable with that of the multiplier and which uses some control logic which is common to both systems. The method is further extended to include the design of a square-rooter.

1. Introduction

The introduction of microelectronics has led to the development of larger electronic systems having high reliability. This is because it has now become possible to use some fixed volume of space in an electronic machine or instrument to implement a more complex sub-system than had formerly been possible. Medium-scale and large-scale integration have caused designers to think again about the logical functions which may be economically viable for implementation in digital systems. It is now possible to implement several hundred gates on a single chip and designers are seeking sub-systems which are complex enough to warrant the use of medium-scale integration but general enough to be economic propositions through wide acceptance. Some sub-systems are proposed here which may be sufficiently versatile to be considered in this connection.

2. Serial-Parallel Multiplier

Electronic machines such as desk calculators and computers have in the past used serial arithmetic, but this has sometimes been discarded in favour of parallel arithmetic where high speed is required.[1–3] However, the parallel approach uses more circuitry and the additional cost, even when using integrated circuits, may not be justified. The serial-parallel approach used here is a compromise between these two methods. It is faster than a completely serial system whilst the units of which it is composed, being general in nature, are attractive to the integrated circuit systems designer.

Figure 1 shows the basic arrangement of the multiplying system in which the multiplier is available in parallel to a controlled register which consists of bistable elements and a special form of adder between adjacent elements.[4] The system functions in such a way that the multiplier becomes added into the controlled register whenever the least significant digit which remains of the multiplicand is a 1. However, when it is a 0, the number in the register is advanced towards the product shift register without addition taking place at that time. This may be illustrated by Example 1 in which decimal 11 is multiplied by decimal 13.

Example 1: 11 × 13

	T U V W X (0 1 0 1 1)		
1 1 0 1			add 1011 and shift right once
	0 0 1 0 1	1	
1 1 0			shift once without adding
	0 0 0 1 0	1 1	
1 1			add 1011 and shift right once
	0 0 1 1 0	1 1 1	
1			add 1011 and shift right once
	0 1 0 0 0	1 1 1 1	
			shift out and stop
		0 1 0 0 0 1 1 1 1	

Product = 143

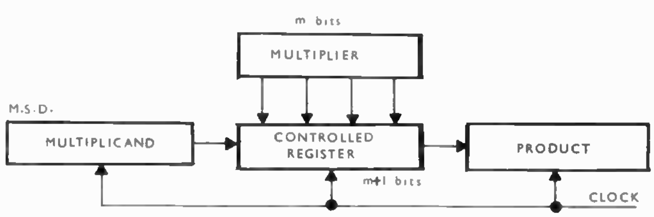

Fig. 1. Functional diagram showing a serial-parallel multiplying system.

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This shows that the controlled register is used to multiply by the shift-and-add method, so that the controlling logic must consist of conditional adders which add when the least significant remaining digit of the multiplicand is a 1 and shift without adding when it is not. This can be implemented as shown in Fig. 2.

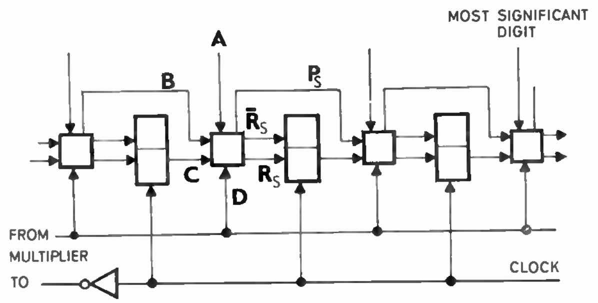


Fig. 2. Schematic diagram showing the logical arrangements for the controlled shift register.

In order to design the logic between the bistable elements a truth-table can be constructed from which the control functions can be deduced. However, in this example a table of this sort is rather trivial.

The control logic is such that R_s is the sum of A, B and C when D = 1, but when D = 0, R_s = C. The carry, P_s, from this sum is independent of D. Karnaugh maps based on this truth-table or in this case on the information just stated, can be used to give the minimized logical functions for P_s and R_s. These maps are shown in Fig. 3.

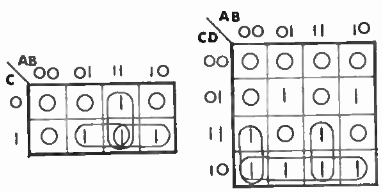


Fig. 3.

$P_s = AB+BC+AC \quad R_s = \bar{C}\bar{D}+\bar{A}\bar{B}C+ABC+\bar{A}B\bar{C}D+A\bar{B}\bar{C}D$

Karnaugh maps showing how a minimal logic function for R_s and P_s can be derived. It should be noted that if NOR-NOR logic is to be used an alternative minimized form will be required.

From these maps the minimal logic between each bistable element can be deduced, and from this, further inspection shows how the logic can be implemented. Figure 4 shows the network fabricated using DTL 2-input gates and inverters. These gates have been used here because of their wired-OR facility.

The clock pulse line to the controlled register is arranged to be out of phase with the multiplicand register so that two operations are carried out alternately: the logic is set up, depending on the input D, and subsequently the bistable elements transfer the data ready for the next addition.

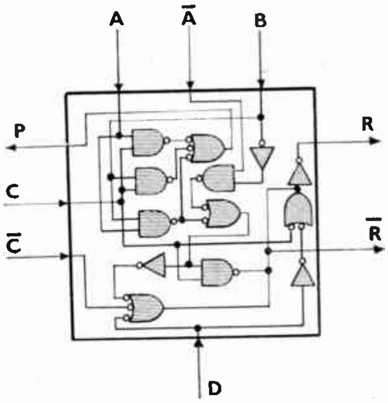


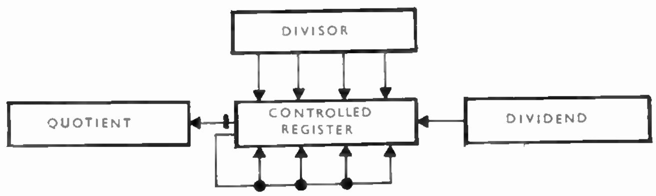
Fig. 4. Logic system required between adjacent bistable elements implemented with DTL gates.

3. Serial-Parallel Divider

Just as the problem of multiplication was solved using a shift-and-add technique, a dividing system will now be considered as shown in Fig. 5, which depends on successive subtraction. However, should the number in the controlled register be about to become negative, then the register advances its contents without subtraction. Example 2, 143 divided by 13, can be used to illustrate this.

The control logic in the controlled register consists of conditional subtractors which subtract the divisor whenever the number in the controlled register is not less than the divisor. The ones complement of the output of the register then gives the quotient.

Fig. 5. Functional diagram showing a serial-parallel divider.



Example 2: 143 ÷ 13

Com-plemen-ted output	TUVWX (0 1 1 0 1)		
		1 0 0 0 1 1 1 1	
			shift left once
	1	0 0 0 1 1 1 1	
			shift left once
	1 0	0 0 1 1 1 1	
			shift left once
	1 0 0	0 1 1 1 1	
			shift left once
	1 0 0 0	1 1 1 1	
			shift left once
	1 0 0 0 1	1 1 1	
			subtract 1101 and shift once
1	0 1 0 0 1	1 1	
			shift left once
1 0	1 0 0 1 1	1	
			subtract 1101 and shift once
1 0 1	0 1 1 0 1		
			subtract 1101, shift and stop
1 0 1 1	0 0 0 0 0		

Quotient = 1011 = decimal 11

Figure 6 shows part of the controlled register, the control logic for each stage of which is such that R_D is the difference $C-A$, with a borrow B, and P_D is the borrow-out to the next more significant stage of the register. However, when the final borrow, P_z in the Figure, is a 1 this shows that A is greater than C. This final borrow is the D input to each stage of the register. When $D = 1$, $R_D = C$. The Karnaugh maps shown in Fig. 7 give the control logic for these subtractors.

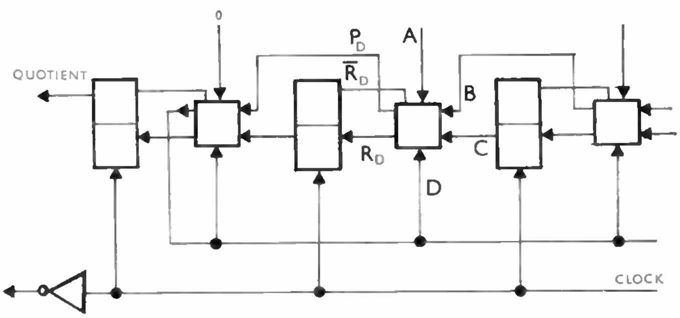


Fig. 6. Schematic diagram showing the logical arrangements for the controlled register acting as a divider.

4. Square-root Generator

The serial-parallel divider just described can be used to generate the square root of a binary number.[5, 6] The divider must be controlled by an auxiliary special-purpose register so that the correct sequence of subtractions can take place.

The register shown in Fig. 8 is suitable for this purpose, and if a 0 is inserted between the two least significant digits of the outputs of the register as shown in the figure, they follow the branching sequence illustrated in Fig. 9. The exact route taken through this sequence depends on the levels at the input Q, in Fig. 8, at successive clock times, the states changing on alternate clock pulses.

The register in Fig. 8 then replaces the divisor register in Fig. 4, whilst the dividend register contains an even number of digits of the number whose square-root is to be obtained. The controlled register in Fig. 5, however, must also be modified so that subtraction can only take place on alternate shifts. This can be done by inserting a bistable element after the borrow-out of the most significant subtractor. This additional flip-flop is then controlled so that its output (which is fed back to control all the subtractors in the register) takes up the level of the borrow-out on alternate clock pulses only. On the remaining clock pulses the subtractors are inhibited so that they advance the number in the register without subtraction, just as they would do were the contents of the controlled register about to become negative. This is illustrated in Example 3, in which the square root of 361 is evaluated. Each line of the example represents two clock-pulse times.

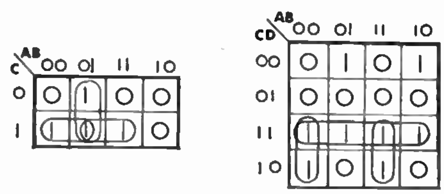


Fig. 7.

$P_D = \bar{A}B + BC + \bar{A}C \quad R_D = CD + \bar{A}\bar{B}C + ABC + \bar{A}B\bar{C}\bar{D} + A\bar{B}\bar{C}\bar{D}$

Karnaugh maps showing the derivation of a minimal logic for R_D and P_D.

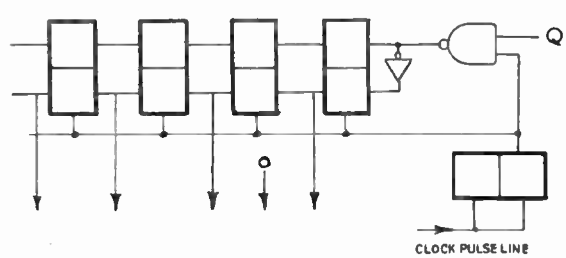


Fig 8. This sub-system replaces the divisor when the dividing system is being used to obtain a square-root.

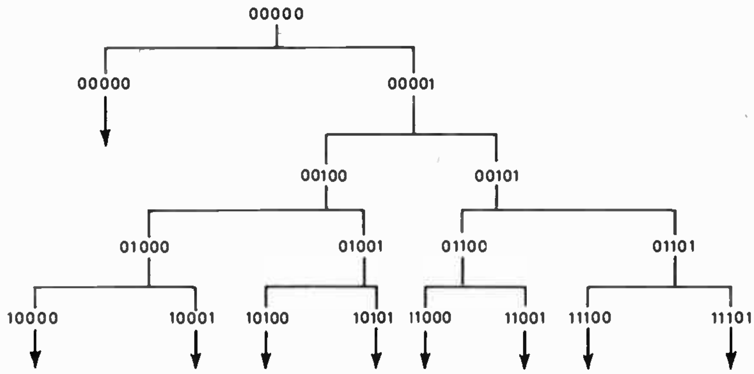


Fig. 9. Tree of binary divisors used in obtaining the square-root of a number.

Example 3: $\sqrt{361}$

Quotient register	Controlled register	Dividend register	Divisor register
		01 01 10 10 01	
	01	01 10 10 01	1
1	01	10 10 01	100
10	01 10	10 01	1000
100	01 10 10	01	10001
1001	10 01 01		100101
10011	00 00 00		

Square-root = 10011 = decimal 19.

The dividend register originally contained the binary code for 361, which was passed to the controlled register two bits at a time, whilst the divisor register followed one of the routes illustrated in Fig. 9, as dictated by the most significant borrow-out of the divider. In this example the final content of the quotient register was the binary equivalent of decimal 19, with zero remainder in the controlled register.

5. Conclusion

Comparing Figs. 3 and 7, R_D differs from R_S only in the negation of D, while P_D differs from P_S only in the negation of A. It would thus be possible to use a single control system for the controlled register composed, for example, of R_S, P_D and an inverter to implement either the serial-parallel multiplier or divider. It may be convenient to provide both the real and the negated outputs of R to operate the bistable elements. Thus a dual-purpose medium scale integrated circuit system might consist of a series of chips each carrying one or more flip-flops and logic networks (as in Figs. 3 and 7), comprising a single stage of the controlled register. Alternatively, since each stage of the controlled register consists of about 20 gates, a 50-stage register could be constructed as a large scale integrated system.

6. References

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U.D.C. 621.396.76

Elimination of Character Smear on Labelled Plan Displays

By

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E. M. SAWTELLE†

AND

L. J. WAPELHORST‡

Summary: Alpha-numeric characters are used to identify radar targets on a p.p.i. display for air traffic control. Normally when using medium- to long-persistence phosphor on the display c.r.t., objectional smear results as alpha-numeric characters are moved. A technique for writing alpha-numeric characters is described which reduces the effect of smear by use of interlace-written monoscope alpha-numerics and a mix of different coloured phosphors to enhance contrast of data.

1. Introduction

An experimental p.p.i. radar display has been built which is capable of writing high quality, non-flicker, colour-contrasted characters on long persistence phosphor without position-update smear. Smear is defined as the trail left by the character blocks (Fig. 1) when the block of information is moved on the cathode-ray tube face from one X-Y position to another while tracking radar targets. The elimination of the smear is achieved by use of a special phosphor mix and special character generation techniques. The special phosphor mix is made by adding a trace of low-persistence phosphor of a suitable colour contrast to a long-persistence phosphor. The special character generator causes the characters to appear as a different colour from the radar data which have the colour of the long-persistence phosphor. Superposition of data blocks on radar video does not result in confusion between the two as normally occurs on a P7 or P33 display. Instead, the high colour contrast between characters and radar video allows quick recognition of both types of data. Further, since there is no character smear, there is no interference with radar target trails. The contrast is such that the radar target is seen through the data block. The data block tends to give one the feeling that it is on a different plane than the radar data. Perhaps this results from the 'see through' effect. The cathode-ray tubes for this type display are essentially the same cost as a single phosphor tube and do not require any additional circuitry.§ The particular colour combination for a P33 with a trace of P11 was a pleasing violet data block while the radar information remained its normal orange. The characters tend to appear as not having hue due to the short duration of the P11 flash. The eye begins to see a hue when the product of light duration and

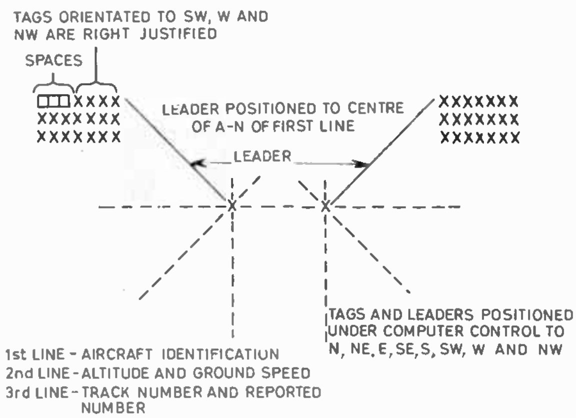

Fig. 1. Character block on c.r.t. radar display.

luminance level exceeds the detection threshold by an increasing margin. The characters are generated by an interlace pattern. Each field of the interlace writes a different part of the character just as do the television interlaced fields. The sequence of interlace and the maximum character size were found to be critical in preventing apparent motion of video dots within a character. Improper sequence caused the eye to see 'crawl' of video dots making up a character. Too large a character causes the eye to see the individual video dot and the effect called 'twinkle' is present. These effects are also a function of the interlace multiple.

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§ The tubes were specially built by Thomas Electronics Inc., Wayne, New Jersey.

2. Smear Reducing Techniques

2.1. *Phosphor Effects*

A major difficulty with characters written on long-persistence phosphors such as a P7 has been the smear or trail when they are moved while tracking with radar targets. This was particularly true when characters were refreshed frequently enough to prevent flicker. Consequently, any character-writing technique developed must effectively deal with the smear problem.

Several techniques are now known by which the smear may be diminished or eliminated. Each of the techniques has received experimental effort in the NAFEC Display Laboratory. The first attempts were made using typotron tubes that were periodically erased and thereby removed the accumulated smear. The technique of selectively erasing local areas was limited in application to the direct view storage tube. A later attempt was briefly demonstrated by using a non-storing vidicon viewing a long-persistency phosphor tube. The viewed display was the 'Eidophor' large screen projection system. In this technique the program control of character refresh was used to inhibit format refresh prior to repositioning of alpha-numeric formats when tracking with radar targets. The technique can be classified as being partially successful in actually eliminating smear as opposed to the first approach.

The effort to provide a smear-free display by means of a phosphor arrangement on the c.r.t. led to experimentation into phosphor combinations. A requirement of the solution desired was that resolution of the display should not be diminished. This condition eliminated the shadow-mask approach and the tiny prism approach. This latter approach called for different phosphors to be deposited on opposing sides —a long-persistence on one and a short-persistence on the other.

2.1.1. Energy-differentiating phosphor combination

Two types of combinations have been considered in this category. The first was a combination in which a shorter persistence phosphor would be activated by virtue of short duration video such as is found in high-speed character writers. The longer persistence phosphor was to activate on raw radar video. This energy-differentiating technique was not successful, but did show the great contrast improvement between alpha-numerics and video that could be produced with colour and its diminution of the effect of smear. The phosphor produced blue characters while radar video was orange coloured. This orange colour which persisted after update was objectionable to air traffic controllers. Pulses of only 200 ns duration were still sufficient to activate both phosphors. The mix was 10% P11 and 90% P38 with 1 s decay time.

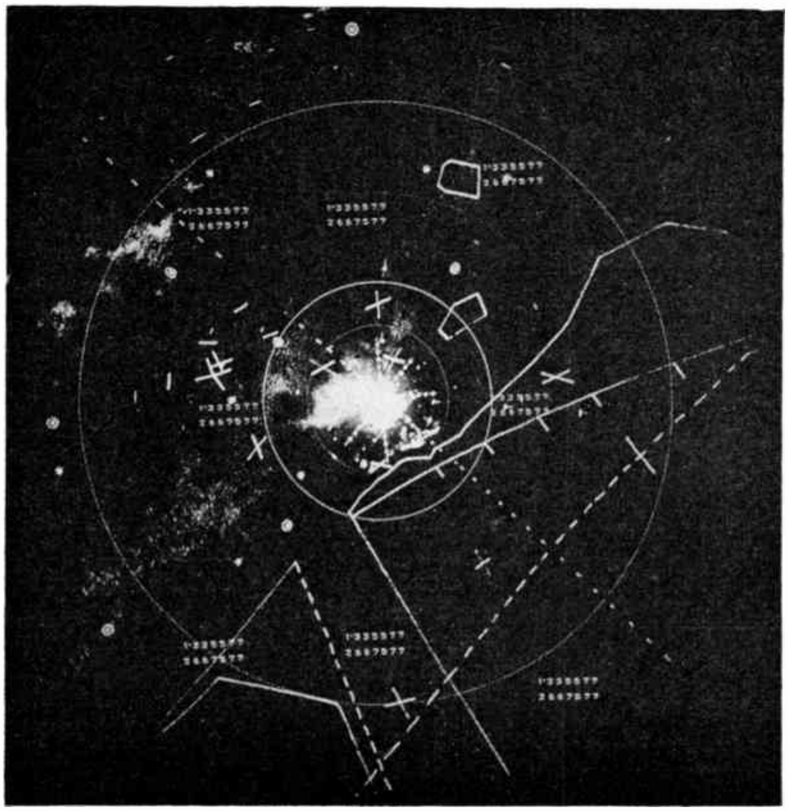

Fig. 2. Static large monoscope-generated alpha-numeric characters with digital sweep-time reduced radar video, map and range rings. 60 radar-miles in approximately 6 radar-mile time.

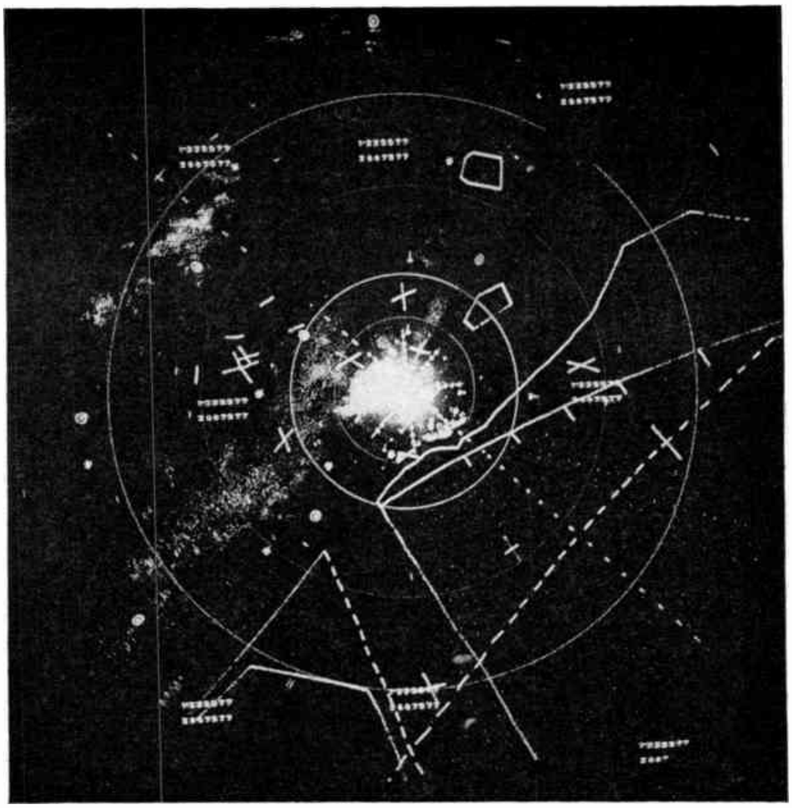

Fig. 3. Moving medium monoscope-generated alpha-numeric character with digital sweep-time reduced radar video, map and range rings. 60 radar-miles in approximately 6 radar-mile time.

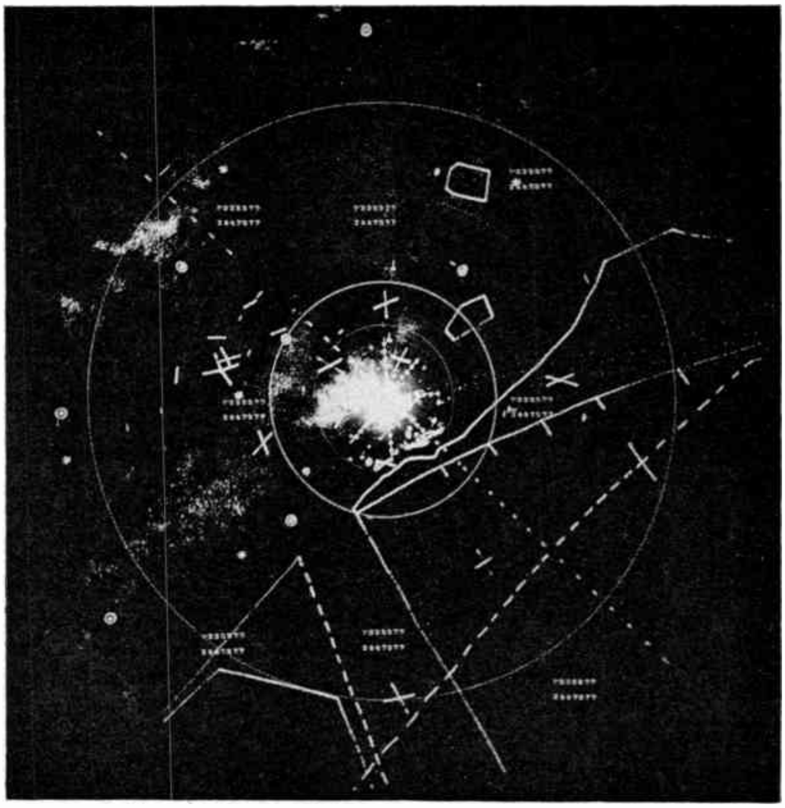

Fig. 4. Static small monoscope generated alpha-numeric character with digital sweep-time reduced radar video, map and range rings. 60 radar-miles in approximately 6 radar-mile time.

A second approach for energy selection calls for the phosphor to be layered and the beam accelerating voltage to be switched to select colour. The colours possible by this technique are widely recognized. However, the time required to switch several kilovolts to achieve colour change is a limiting factor to application in alpha-numeric displays used for air traffic control radar.

2.1.2. Simple phosphor mixes of different persistence and colour

While the objective was not achieved with the first combinations, it did demonstrate the value of reducing the objectionable confusion factor of smear through colour contrast. The brightness difference between trial and refreshed characters was also increased by interrupting refresh prior to update. The limited success of these techniques led to further experiments with simple mixes of phosphor when character-writing time was reduced to 10–20 μs. Earlier character-writing times had been in excess of 100 μs. The effects of the monoscope character generator combined with the phosphor mix greatly diminished smear. This combination made possible use of a long-persistence phosphor such as the P33 having 10 s to 10% brightness as opposed to the P7 with only 0·3 s. The 10% decay time is used for convenience of comparison. The trails used with p.p.i. displays considered were more nearly at the 1% level and below since the radar scan-time is approximately 4 s per scan. Table 1 in the Appendix shows the results of tests using mixes of P33 and P2 phosphors. These tests were made on p.p.i. radar display. Figures 2–4 are photographs of a p.p.i. display in which radar video has been digitally sweep-time compressed. The characters are of diminishing size from Figures 2 to 4. Camera shutter time was approximately 4 seconds while character refresh rate was near 20 per second. The film used was blue-sensitive only. These photographic conditions account for thickness of character lines and lack of radar trail.

While only two colours of phosphor were used, three colours appeared on the display. For the P33–P11 mix, the sweep line of video appears as a blue line rotating with the antenna azimuth. The radar video p.p.i. data is displayed in the P22 orange colour. Finally, the alphanumeric data are lavender. The reason for this shift in colour may be explained, at least in part, by the 'Bunson–Roscoe Law' (see Appendix) which describes the temporal integration of the eye. The hue that the eye perceives is dependent upon the duration of the image and the luminance level. Hue becomes less saturated as the product of duration and luminance approaches the detection level and approaches a no-hue or 'white light' condition. The duration of the P11 flash is less than 50 μs to 10% brightness while the duration of the orange is much longer, while at a much lower level. Even so, the critical chromatic flicker frequency is exceeded and the hue perceived by the eye is a combination of blue and orange, that is, a lavender or violet. The smear effect of the persistent orange is a function of luminance contrast as well as colour contrast. Consequently, as both parameters are maximized, smear is greatly reduced.

2.1.3. Other phosphors

The experimentation was continued to determine the effectiveness of writing the characters with a red phosphor while radar data were displayed in a green phosphor. The red was a short-persistence phosphor while the green was a long-persistence of P2 characteristics.

Characters produced were a non-saturated hue of pink. Again as with the P33–P11 mix, the radar data could be easily read through the characters. Other characteristics of smear and brightness were also equivalent.

2.2. *Character Generator Effects*

The monoscope character technique has resulted in a smearless display for long-persistence phosphors such as the P33 (10 sec to 10% peak luminance). The smear is progressively diminished with increasing interlace ratio. At 8 : 1 interlace of a monoscope character, no smear is observable at the low light ambients used (one-foot candle or less). Instability in the control of the interlace (a temporary laboratory problem only) could cause non-interlace. While operating on 8 : 1 mode, the result was a 4 : 1 level of interlace. Under this condition, dots while fixed (no apparent motion) and not varying in intensity did not diminish character resolution. The aesthetic appearance of the character was not quite as good as the 8 : 1 interlace character. However, when a drop-out of the 4 : 1 interlaced character to the 2 : 1 level occurred, there was a definite deterioration in character quality. Closure of dots defining the character was not achieved. Further, in 2 : 1 interlace as the refresh rate is reduced, flicker occurs at a higher rate than with 4 : 1 or 8 : 1 interlace.

3. Monoscope Character Generator Design

The task was to provide the airport surveillance radar (ASR-4 display console) with a character generator. The console is of vacuum-tube design and contains a deflection yoke of approximately 9 mH inductance. The character generator was to cause a minimum modification, have a character writing speed of 20 μs or less, and be inexpensive. Further, the generator was to be used with a sweep-time reduced display.[2] The radar video was to require a maximum of 150 μs out of an available 833 μs period between radar triggers. Alpha-numeric characters were to be written in radar 'dead time', the period of time from the end of a radar sweep to the start of the next radar sweep. The system was to write a format consisting of two lines of seven characters each per dead time.

Based on the bandwidth constraints of the existing system the use of a stroke character generator was not possible. A sine-wave written character appeared to be feasible if the frequency could be taken high enough to give the necessary resolution. The first experimentation was to attempt to read out a 7-by-5 dot matrix using the linear portion of the sine-wave. Resolution of the displayed characters was too limited by this approach and was abandoned for monoscope-generated characters which when used with interlace showed much high resolution and less time per character.

3.1. *Deflection Systems*

Tests to determine the deflection system bandwidth showed it to be about 50 kHz and the yoke resonant frequency to be at 80 kHz. It was shown, however, that an undistorted sine-wave of 280 kHz could be passed, and resulted in a total deflection of $\frac{3}{8}$ in (9·6 mm).

This was adequate for the system whose characters were to be normally $\frac{1}{10}$ to $\frac{1}{8}$ in (2·5 to 3·2 mm) high.

Based on the above, and since the monoscope may be used with sine-wave scan of characters, the monoscope was selected as a basic building block for the character generator. The monoscope is a secondary-emission characteristic c.r.t. which contains a matrix of 64 characters. When the etched portion of the selected character is scanned, a secondary-emission change is sensed, amplified, and used to intensity-modulate the grid of the display c.r.t. Since display deflection signals are synchronized with the monoscope scanning pattern, the character is thus reproduced on the display c.r.t. The monoscope used, a Raytheon CK 1414, has electrostatic deflection. Phase-shift correction was therefore necessary between electrostatic monoscope and electromagnetic display. This same correction also served to correct the above resonance phase shift in the 250 kHz deflection sine-wave for proper synchronization.

Movement of the monoscope and c.r.t. beams to new position for new characters was first attempted by turning off the sine-wave drive while a ramp was applied for movement. The large current drive to the yoke caused ringing when this was done. Movement to new characters was accomplished in half a cycle so that the sine-wave generator need not be turned off and thereby prevent the yoke ringing. The 9 mH yoke was, by proper phase correction, driven at 320 kHz with the existing deflection amplifier. This upper limit represented the available power supply capability.

3.2. *Monoscope Character Resolution*

The monoscope character resolution is firstly a function of the scanning sine-wave frequency, interlace multiple, video processing, character size, c.r.t. luminance level, and secondly of phosphor persistence and the ambient lighting. An increase in either of the latter two parameters will necessitate an increase in other system parameters. Another factor of concern is uniformity of character luminance.

The resolution of the monoscope characters was found to be satisfactory when scanned with not less than 10 passes, i.e. five cycles through a character. Further, video from the monoscope was standardized as will be described below. This number of cycles can be reduced as the interlace ratio is increased. Interlaced characters were made from two or more fields occurring successively by giving the deflecting sine-wave an appropriate phase shift. With four passes and an 8 : 1 interlace, each character has 32 passes. (This is a satisfactory character, but we wanted to allow the variation of size of character to be wide from small to large and so provided a four-cycle character.) Such a character was of high definition without the expense of additional time.

The most difficult character to produce was a three horizontal bar character. The character was produced without intensity variance between the centre bar and those above and below it.

Consideration to the order of the interlace sequence for 4 : 1 and 8 : 1 ratios was important to the appearance of the character. The order of interlace for 4 : 1 was the sequence of 0°, 180°, 90° and 270°. The 8 : 1 interlace follows a similar pattern of 45° intervals, i.e. 0°, 180°, 45°, 225°, 90°, 270°, 135° and 315°. The purpose of this order of interlace was to prevent a crawl or apparent motion of the character lines. When a capital 'T' was written without proper interlace, the spot would appear to move part way across the cross. The amount of motion was dependent upon the interlace ratio and on the passes per field for a particular phosphor. The amount of crawl was maximum for the 8 : 1 interlace when improperly interlaced. Excellent interlace stability was obtained by using a completely digital interlace generator. An interlaced sine-wave results by gating the appropriate 250 kHz square-wave to the output and then eliminating all but the fundamental by utilization of a 250 kHz resonant circuit. If a leader is required, the leader sweeps are gated into the deflection amplifiers. The sine-wave oscillator can be left running if an interlaced sine-wave derived signal is also fed to the Z-axis of the c.r.t. The leader is then a series of dots though it can

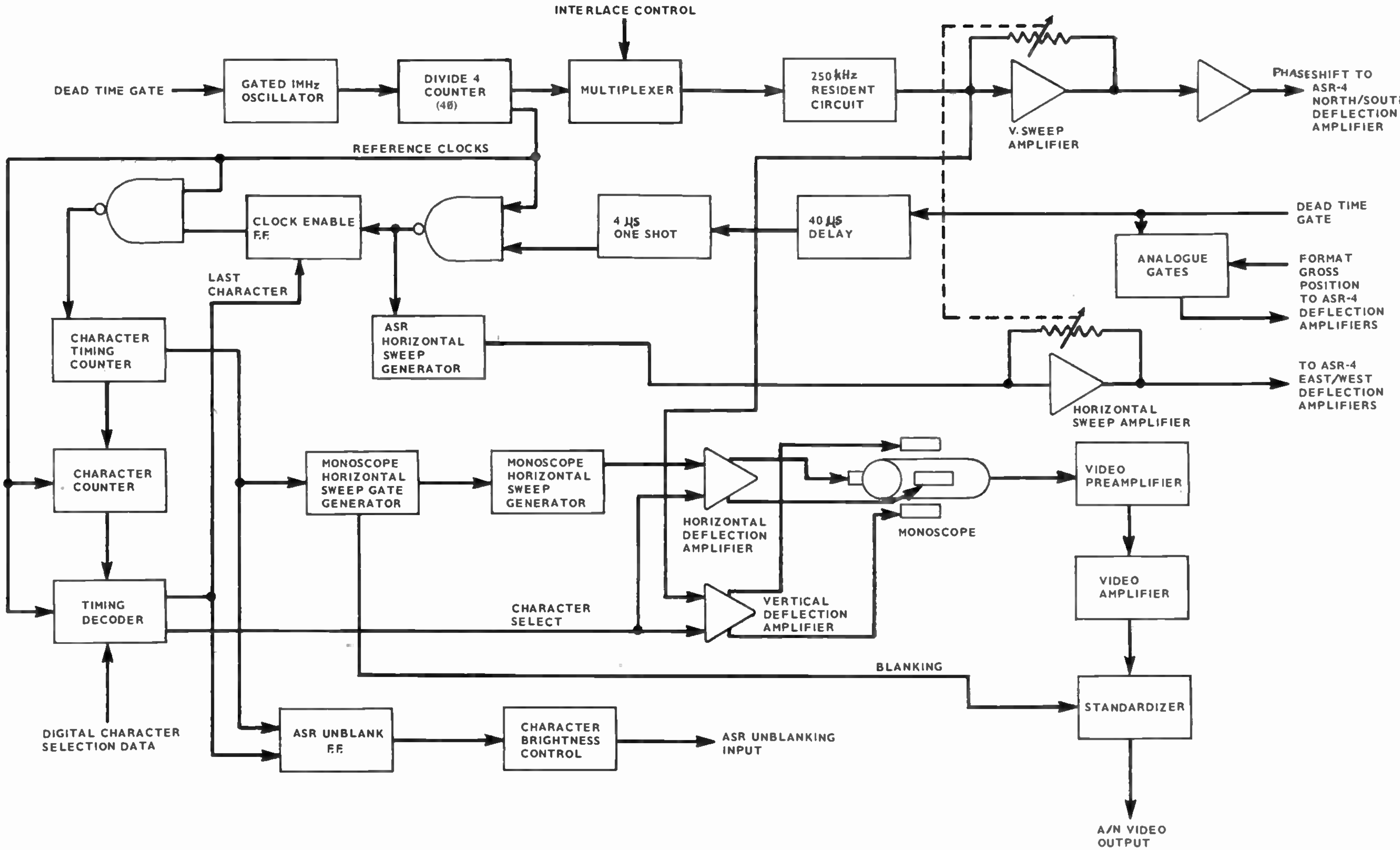

Fig. 5. Monoscope character generator block diagram.

be made to look like a solid line. This dot make-up again reduces the storage problem (see Appendix).

3.3. *Video Amplifier*

The read-out of video from the monoscope was amplified by a standardizing pre-amplifier which chopped all video signals to give standard amplitude and width (see Fig. 6). This standardization prevented hot spots from appearing in the displayed characters. Failure of this circuitry to provide standard video will result in phosphor-type noise similar to that occurring in the event of interlace instability.

3.4. *Size and Spacing*

The size and spacing of the characters are controlled by one potentiometer which varies the sine-wave amplitude to the display. This amplitude control affects only the height of characters. Characters less than $\frac{1}{10}$ in (2·5 mm) are easily read.

Adjustment of spacing and width of characters is varied by change of the X-sweep ramp. This causes spacing to be held proportional to the character width.

When large characters are required, and they are to be viewed at 3 to 4 ft (1·0 to 1·3 m) as well as at 2 ft (0·65 m), then characters should be written with five to six cycles per field. Large characters are $\frac{3}{16}$ to $\frac{1}{4}$ in (4·8 to 6·4 mm) in height.

4. Description of Monoscope Character Generator

The sequence of events in the character generator (Fig. 5) will first be described without interlace. Interlace operation as well as discussion of interlace experimentation results will then be presented.

4.1. *Operation*

At the end of the radar sweep the 'dead time' gate enables the gated 1 MHz oscillator, gates on the format gross position information, and triggers a 40 μs delay circuit. This delay circuit is required so that the display beam can approach the format position from any direction along the outer periphery of the c.r.t. and settle to 0·1%. If insufficient time were allowed before writing characters, the first character written would 'orbit' or be pulled in the direction from which it was approached.

The 'divide by 4' counter produced the four phases of 250 kHz square-waves. One phase was used as the reference clock for the associated logic and all are multiplexed to be used as the source of the vertical scanning pattern. A series-resonant circuit was used to filter out all but the fundamental, to produce the four-phase sine-wave for deflection. At the completion of the 40 μs delay, the display beam is settled at the formal gross position point. A 4 μs one-shot multivibrator enables the next clock pulse to trigger the

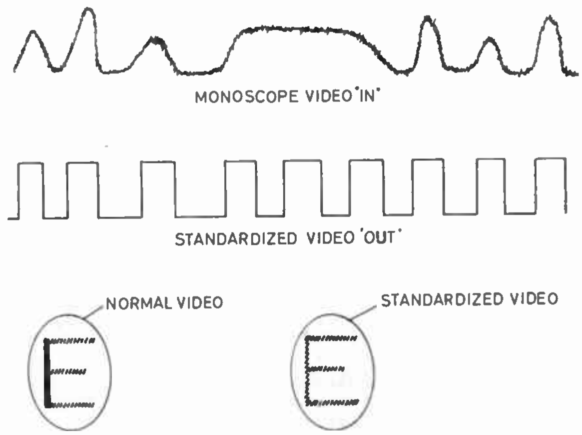

Fig. 6. Brightness correction by character video standardization.

horizontal ASR-4 character sweep and 'clock-enable' flip-flop. The ASR-4 character sweep lasts for the complete line of characters. The digital character selection data are then gated to the monoscope deflection amplifiers. The monoscope horizontal sweep gate generator then enables the horizontal sweep generator for the next four clock pulses or 16 μs.

4.2. *Standardizer*

The ASR-4 contains a 2·5 MHz video amplifier. Because of this bandwidth limitation, when normal character video was used, the brightness at various points in the character varied widely because of the difference in video bandwidth requirements at these points (Fig. 6). These requirements are the most severe when the selected character contains a horizontal line near the centre where the sine-wave velocity is greatest. Further, as the monoscope ages, a change in secondary emission occurs, resulting in differences in brightness of characters if not compensated. This problem was corrected by gating a pulse train whose length is a function of character video duration. This pulse train is therefore 'character synchronized'. The threshold of this gating circuitry determines the amplitude of visible monoscope noise. The bandwidth of the display video amplifier can be reduced significantly as a result of this technique.

The standardizer was developed to eliminate inter-character shading due to bandwidth limitation. The standardizer is a stable multivibrator with the 'on' time pulse-width set to coincide approximately with the narrowest video pulse (about 100 ns) and the period is set so that a vertical line produced four pulses. Since the standardizer is turned on by the video and each of the four phases passes over a slightly different portion of the character, the resulting characters are of an extremely high quality.

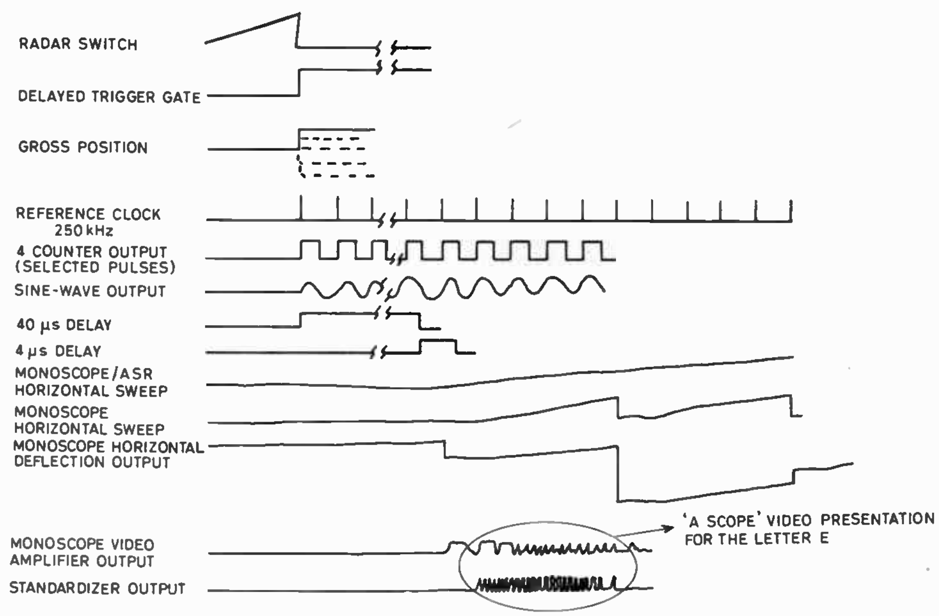

Fig. 7. Character generator timing diagram.

Character size could be varied from $\frac{1}{16}$ to $\frac{3}{8}$ in (1·6 to 4·8 mm) in height and remain quite legible throughout this range. This size variation was accomplished by merely changing the gain of the horizontal and vertical sweep amplifiers (Fig. 5). The character generator timing diagram is shown in Fig. 7.

4.3. *Interlace*

Interlace is accomplished in the following manner. All data to be written on the display are written on a single selected phase. The basic refresh rate on the system varied from 15 to 35 Hz. The next time through the list at this rate the data are switched to a different sine-wave phase. Therefore the *complete character* refresh rate is the ratio of the nominal character refresh rate to the number of interlaced phases.

Phase sequence selection was very important. When the sequence 0°, 90°, 180°, 270° was used the eye could follow this sequence when operating at 15 Hz. However, a sequence of 0°, 180°, 90°, 270° resulted in a very satisfactory character even at 15 Hz (a particular character element refresh rate of less than 4 Hz) on the P33 phosphor.

5. Conclusion

The proposed method for writing alpha-numeric characters using a monoscope is one of several techniques presently available to the Federal Aviation Administration. The other systems, utilizing high-current yokes with main deflection speeds in the order of 6 to 12 μs for the typical 16-in cathode-ray tube, anode voltage of 12 kV, are being studied as are the new printed yokes with their low capacitance and high repeatability of characteristics from yoke to yoke. The advantages of colour-mix to enhance contrast is applicable to all these writing methods. The advantage of the monoscope is the use of the sine-wave which can be employed to obtain relatively high-speed characters by the proposed interlacing method. The use of standardized video and digital control of the interlacing phase shift has removed many of the objections to this technique. The size of the characters can be varied within wide limits, i.e. from less than $\frac{1}{10}$ in (2·5 mm) to approximately $\frac{1}{4}$ in (6·2 mm) in height while maintaining a predetermined aspect-ratio by the turn of a single control available to each display without adding any complications to the system. The use of a pulse train allows the employment of a video amplifier of relatively low pass-band. In this particular case, the bandwidth of the video amplifier was measured at 2·5 MHz, and it was the amplifier that belonged to this terminal radar display. No changes were made to the amplifier while it was operating at a sine-wave frequency of 250 MHz.

6. References

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Table 1

Combinations of phosphor mixes and their characteristics

Phosphor mix	Persistence to 10% (s)	Colour	Radar video					Alpha-numeric video					
			Sweep-line colour	Video-trail colour	Luminance relative to P7	Filter required	Number of radar trails	Colour: Initial	Colour: Steady	Colour: Smear	Spot crawl	Filter required	Refresh rate for c.f.f.
P-33	10	orange	blue	orange	high	no	5 or more	lavender	lavender	none	no	no	12
P-11	10×10⁻⁶	blue											
P-2	5·0	green-yellow		green	low		5 or more						
Europium (25% of P-16)	2×10⁻⁶	red	red					red	yellow	low	yes	no	25
P-16	0·1×10⁻⁶	ultra-violet											
			orange	green		yellow			yellow	green	yes	no	20
P-2	5·0	green											
Europium red		red	gold	green	low		5 or more	yellow	yellow†	no	yes	no	20
Ultra-violet		faint blue											
P-7	0·33	yellow-green	blue	yellow-green	1	yellow	4 or more	white	yellow	no	little	yes	24

†Formats red to fast eye movement.

7. Appendix

7.1. *Bunson–Roscoe Law*

The cathode-ray tube is an excellent device in the manner in which it achieves economy of bandwidth in deflection systems and video amplifiers while also effecting great economy of system cost. Its operation is in accordance with the Bunson–Roscoe law which defines the threshold of eye excitation, K, for light of duration T (when less than 50 to 100 ms). In a light stimulus, luminance, B, and time, T, are interrelated so that

$$BT = K$$

This law describes the phenomenon of temporal integration which the eye exhibits in producing 'after-images'. The duration of the luminance will have a definite effect on the hue received. If the after-image is of the same hue as the image it is termed 'homochromatic'; however, when the stimulus is of short enough duration, the after-image is not a saturated hue, but tends toward no hue-white image. This effect may have important application in display systems to enhance contrast through colour differences. Finally, consideration of the temporal integration to display design may significantly reduce the refresh rate of characters and television raster displays by the appropriate use of interlace.

7.2. *Types of Phosphor Screens*

Several different combinations of phosphor screens were studied in the course of experimentation. The P33–P11 mix is described in the body of the paper. This mix together with the other mixes and their responses relative to the intended use for air traffic control radar display are shown in Table 1. Very relative ratings are listed since the response is completely dependent upon the phosphor mix percentage.

7.3. *Phosphor Capability*

The range of high to low traffic density of terminal facilities require a comparable range of display capacity. The type of phosphor used sets the allowable repetition rates and capability. Listed in Table 1 are quantities of formats that are possible if a mix of P33–P11 phosphors is used in lieu of the P7 phosphor for an ASR-4 display tube. For all cases, the radar trigger interval is taken to be 833 μs; dead-time used is 91 μs and is defined as the time between 60-mile radar video end and the next trigger. The calculations are based on a refresh rate of 15 per second for P33–P11 phosphor mix as opposed to a refresh rate of 20 to 30 per second for the P7 phosphor.

Eighty (80) two-line formats could be written without sweep time compression of radar while on the 30-mile range. Forty (40) formats may be written at 40 miles, again without time compression. Each such format is composed of two lines of seven characters per line, a leader, and target symbol.

A system without time compression is a possibility if format size is limited. The limited formats would consist of a target symbol, leader, and two lines of characters. The first character line would be written with four characters giving beacon code. The second line has only three characters to give altitude. If the system used these clutter-reducing data blocks, 26 could be written in dead-time after real-time display of 55 miles of video. The same number of formats could also be written in the configuration of a single seven-character line plus leader and target symbol.

This character generation would require a refresh memory and control logic. The functions could be performed by a small computer of moderate speed having a memory of 4000 to 8000 words. Under program control the computer would accept keyboard inputs, slew-dot data, address data time coincident for selected beacon code and altitude. In addition, it would control such functions as interlace control of characters, field select, format configuration, and blink.

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Two-port Representation of Multi-node Networks by Matrix Partitioning

By

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Summary: A unified approach to linear circuit analysis is described which apart from its simplicity reduces the time required for circuit analysis on an undergraduate electronics course. Representation of a network by its indefinite admittance matrix is well established and for a *n*-node network would result in a $n \times n$ matrix. Reduction of this matrix to a 2×2 matrix is described and an easy to memorize procedure in terms of the input and output nodes is given.

List of Symbols

$[Y]$	short-circuit admittance matrix
$Y_{11}, Y_{12}, Y_{21}, Y_{22}$	elements of the admittance matrix
$\lvert Y \rvert$	determinant of the admittance matrix
g_m, g'_m	mutual conductance
r_a, r'_a	anode slope resistance
μ, μ'	amplification factor
$[Y_i]$	indefinite admittance matrix
$[Z]$	open-circuit impedance matrix
$\underset{\longleftarrow}{[\;\;]}$	arrow indicates current direction at port 2
$[Y_E]$	equivalent 2×2 admittance matrix
$\frac{Y_{11E}}{K}, \frac{Y_{12E}}{K}, \frac{Y_{21E}}{K}, \frac{Y_{22E}}{K}$	elements of equivalent 2×2 matrix
K	determinant of the original matrix with rows 1 and 2 and columns 1 and 2 removed
$\operatorname{adj}\begin{bmatrix} k & 1 \\ m & n \end{bmatrix}$	adjoint $\begin{bmatrix} k & 1 \\ m & n \end{bmatrix} = \begin{bmatrix} n & -1 \\ -m & k \end{bmatrix}$
$h_{11}, h_{12}, h_{21}, h_{22}$	elements of hybrid (h) matrix
$\lvert h \rvert$	determinant of h matrix
$[a]$	transmission matrix

Matrix Conversions

$$Y_{11} = \frac{1}{h_{11}}$$

$$Y_{12} = \frac{-h_{12}}{h_{11}}$$

$$Y_{21} = \frac{h_{21}}{h_{11}}$$

$$Y_{22} = \frac{|h|}{h_{11}}$$

$$[M]^{-1} = \frac{\operatorname{adj} M}{|M|}$$

$|M|$ = determinant of the non-singular square matrix M

1. Introduction

Established methods exist for the representation of networks by node voltage equations of the form

$$[I] = [Y][E] \qquad \text{......(1)}$$

and mesh current equations of the form

$$[E] = [Z][I] \qquad \text{......(2)}$$

The result in the case of complex networks is a high-order admittance matrix $[Y]$ or a high-order impedance matrix $[Z]$. The method selected is usually the one yielding the lower-order matrix and is a function of the network topology.

Traditional methods of analysis tend to be laborious and have been replaced to a large extent by well-known matrix methods which in their more conventional form apply the rules of matrix interconnection to a network represented by individual element matrices. When the more conventional methods are used with transistor and valve circuits the following steps are necessary:

(i) Determine the interconnections to be employed and hence the required matrix (Z, Y, h, etc.).

(ii) Determine the valve or transistor configurations (common-cathode, common-base, etc.).

(iii) Derive the required 2×2 matrix from known information about the active device (usually

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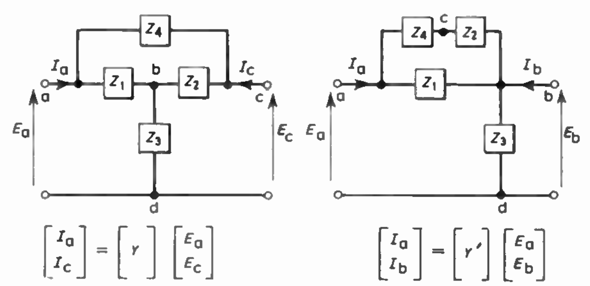
Y in common-cathode for valve and h in common-emitter for transistor).

Occasionally step (iii) is impossible, for instance, in the case of a valve having a singular admittance matrix (eqn. (5)) an impedance matrix does not exist.

The disadvantage of singular matrices, requiring several matrices for the devices and having to determine the actual mode of operation is removed in the method to be described.

The voltage and current at all nodes in a network are not always required and in many cases analysis reduces to terminal characteristics such as voltage transfer function, input impedance and output impedance. This is generally the case with transistor and valve amplifiers that have two ports of interest while the internal complexity may be great. It is therefore important to be able to reduce the high-order matrix for the complete circuit to a 2×2 matrix relevant to the two ports of interest.



(a) (b)

Fig. 1. Two arrangements showing the ports of interest in a simple network.

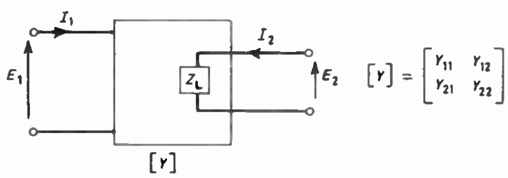


Fig. 2. A simple two-port network.

In the method to be described since there is an arbitrary selection of the input and output ports the performance at any node can be obtained. Figures 1(a) and (b) show two possible arrangements for the ports of interest in a simple network.

If the two-port network of Fig. 2 is represented by the nodal matrix given by

$$\begin{bmatrix} I_1 \\ I_2 \end{bmatrix} = \begin{bmatrix} Y_{11} & Y_{12} \\ Y_{21} & Y_{22} \end{bmatrix} \begin{bmatrix} E_1 \\ E_2 \end{bmatrix} \qquad \ldots\ldots(3)$$

the terminal characteristics can be obtained from the following equations:

$$\text{voltage transfer function} = \frac{-Y_{21}}{Y_{22}} \qquad \ldots\ldots(4a)$$

$$\text{input impedance} = \frac{Y_{22}}{|Y|} \qquad \ldots\ldots(4b)$$

$$\text{output impedance} = \frac{1}{Y_{22}} \qquad \ldots\ldots(4c)$$

The effect of a terminating impedance at port 2 is included in the Y_{22} term.

2. Indefinite Admittance Matrix

Perhaps the most common matrix representation of a network is the nodal matrix in which one of the nodes has been selected as a reference. An n-node network would be represented by a $(n-1)\times(n-1)$ admittance matrix.

If none of the n nodes are defined as a reference a more useful $n\times n$ matrix is formed. This is termed the indefinite admittance matrix and the following are some of its properties.

(i) The sum of every row and every column is zero.

(ii) If any nodes are connected by a zero impedance the rows and columns in the matrix corresponding to these nodes may be added to give a single row and column.

(iii) Any node can be made the reference node by removing from the indefinite admittance matrix the row and column corresponding to the selected reference node.

The nodal admittance matrix for a valve in the common-cathode mode is given by:

$$[\underleftarrow{Y}] = \begin{bmatrix} 0 & 0 \\ g_m & 1/r_a \end{bmatrix} \qquad \ldots\ldots(5)$$

The corresponding indefinite admittance matrix is of the form

$$\begin{bmatrix} I_g \\ I_a \\ I_k \end{bmatrix} = [Y_I] \begin{bmatrix} E_g \\ E_a \\ E_k \end{bmatrix} \qquad \ldots\ldots(6)$$

Applying rule (i), the indefinite admittance matrix for a valve is given by:

$$[\underleftarrow{Y_I}] = \begin{bmatrix} 0 & 0 & 0 \\ g_m & 1/r_a & -g_m-1/r_a \\ -g_m & -1/r_a & g_m+1/r_a \end{bmatrix} \ldots\ldots(7)$$

The nodal matrix in the common-anode mode would be found by using rule (iii).

Removing the row and column corresponding to the anode, i.e. 2

$$\underleftarrow{[Y]} = \begin{bmatrix} 0 & 0 \\ -g_m & g_m + 1/r_a \end{bmatrix} \qquad \text{......(8)}$$

A rule that is easy to memorize and apply obviously exists for the formation of the indefinite admittance matrix. The object of this paper is to produce an equally easy approach for the reduction of matrices to a 2×2 matrix such that the results of eqns. (4) can be used.

3. Reduction of a Multi-node Network to a 2×2 Matrix by Matrix Partitioning

The general admittance matrix for a n-node network is given by

$$\begin{bmatrix} I_1 \\ I_2 \\ I_3 \\ \cdot \\ \cdot \\ I_n \end{bmatrix} = \left[\begin{array}{cc|c} Y_{11} & Y_{12} & Y_{13} \ldots\ldots Y_{1n} \\ Y_{21} & Y_{22} & Y_{23} \ldots\ldots Y_{2n} \\ \hline Y_{31} & Y_{32} & Y_{33} \ldots\ldots Y_{3n} \\ \cdot & \cdot & \cdot \ldots\ldots \cdot \\ \cdot & \cdot & \cdot \ldots\ldots \cdot \\ Y_{n1} & Y_{n2} & Y_{n3} \ldots\ldots Y_{nn} \end{array}\right] \begin{bmatrix} E_1 \\ E_2 \\ E_3 \\ \cdot \\ \cdot \\ E_n \end{bmatrix} \qquad \text{......(9)}$$

The partitioned matrix can be rewritten as

$$\begin{bmatrix} I_1 \\ I_2 \\ I_3 \\ \cdot \\ \cdot \\ I_n \end{bmatrix} = \left[\begin{array}{c|c} A & B \\ \hline C & D \end{array}\right] \begin{bmatrix} E_1 \\ E_2 \\ E_3 \\ \cdot \\ \cdot \\ E_n \end{bmatrix} \qquad \text{......(10)}$$

Consider the case where all nodes other than 1 and 2 are internal then the current generators at nodes 3 to n will be zero. This condition can be represented by the matrix of eqn. (11).

$$\begin{bmatrix} I_3 \\ \cdot \\ \cdot \\ I_n \end{bmatrix} = [\, C \mid D \,] \begin{bmatrix} E_1 \\ E_2 \\ E_3 \\ \cdot \\ \cdot \\ E_n \end{bmatrix} = 0 \qquad \text{......(11)}$$

This reduces to

$$[C]\begin{bmatrix} E_1 \\ E_2 \end{bmatrix} + [D]\begin{bmatrix} E_3 \\ \cdot \\ \cdot \\ E_n \end{bmatrix} = 0 \qquad \text{......(12)}$$

Hence

$$\begin{bmatrix} E_3 \\ \cdot \\ \cdot \\ E_n \end{bmatrix} = -[D]^{-1}[C]\begin{bmatrix} E_1 \\ E_2 \end{bmatrix} \qquad \text{......(13)}$$

The required two-port representation is of the form given by

$$\begin{aligned} I_1 &= f^n(E_1, E_2) \\ I_2 &= f^n(E_1, E_2) \end{aligned} \qquad \text{......(14)}$$

From eqn. (10), one gets

$$\begin{bmatrix} I_1 \\ I_2 \end{bmatrix} = [A]\begin{bmatrix} E_1 \\ E_2 \end{bmatrix} + [B]\begin{bmatrix} E_3 \\ \cdot \\ \cdot \\ E_n \end{bmatrix} \qquad \text{......(15)}$$

Substitution of the result of eqn. (13) into eqn. (15) would give

$$\begin{bmatrix} I_1 \\ I_2 \end{bmatrix} = [A]\begin{bmatrix} E_1 \\ E_2 \end{bmatrix} - [B][D]^{-1}[C]\begin{bmatrix} E_1 \\ E_2 \end{bmatrix} \qquad \text{...(16)}$$

$$\begin{bmatrix} I_1 \\ I_2 \end{bmatrix} = ([A]-[B][D]^{-1}[C])\begin{bmatrix} E_1 \\ E_2 \end{bmatrix} \qquad \text{......(17)}$$

The resultant admittance matrix is given by:

$$\underleftarrow{[Y]} = [A]-[B][D]^{-1}[C] \qquad \text{......(18)}$$

Resubstituting the original values before partitioning, the following equations are obtained

$$\underleftarrow{[Y]} = \begin{bmatrix} Y_{11} & Y_{12} \\ Y_{21} & Y_{22} \end{bmatrix} - \frac{\begin{bmatrix} Y_{13} \ldots Y_{1n} \\ Y_{23} \ldots Y_{2n} \end{bmatrix} \operatorname{adj} \begin{bmatrix} Y_{33} \ldots Y_{3n} \\ \vdots \quad\ \vdots \\ Y_{n3} \ldots Y_{nn} \end{bmatrix} \begin{bmatrix} Y_{31} & Y_{32} \\ \vdots & \vdots \\ Y_{n1} & Y_{n2} \end{bmatrix}}{\begin{vmatrix} Y_{33} \ldots Y_{3n} \\ \vdots \quad\ \vdots \\ Y_{n3} \ldots Y_{nn} \end{vmatrix}} \qquad \text{......(19)}$$

$$\underleftarrow{[Y]} = \frac{1}{\begin{vmatrix} Y_{33} \ldots Y_{3n} \\ \vdots \quad\ \vdots \\ Y_{n3} \ldots Y_{nn} \end{vmatrix}} \left\{ \begin{bmatrix} Y_{11} & Y_{12} \\ Y_{21} & Y_{22} \end{bmatrix} \begin{vmatrix} Y_{33} \ldots Y_{3n} \\ \vdots \quad\ \vdots \\ Y_{n3} \ldots Y_{nn} \end{vmatrix} - \begin{bmatrix} Y_{13} \ldots Y_{1n} \\ Y_{23} \ldots Y_{2n} \end{bmatrix} \operatorname{adj} \begin{bmatrix} Y_{33} \ldots Y_{3n} \\ \vdots \quad\ \vdots \\ Y_{n3} \ldots Y_{nn} \end{bmatrix} \begin{bmatrix} Y_{31} & Y_{32} \\ \vdots & \vdots \\ Y_{n1} & Y_{n2} \end{bmatrix} \right\} \qquad \text{......(20)}$$

Reduction by partitioning has been applied to the impedance matrix $[Z]$ (Ref. 5) and results in an equivalent 2×2 matrix. Admittance reduction does however offer more general usage when applied to electronic networks, since it is not always possible to set up the original impedance matrix. The singular admittance matrix of eqn. (5) does not yield an equivalent impedance matrix and is one such example.

4. Reduction Procedure

Perhaps the most important feature of the method is the simple relationship between the elements of the equivalent 2×2 matrix and the original admittance matrix to be reduced. The equivalent 2×2 matrix can be obtained from eqn. (20), however in its present form a simple procedure is not obvious. This can be overcome by considering a particular case such as $n = 4$.

The admittance matrix for $n = 4$ is given by:

$$\underleftarrow{[Y]} = \begin{bmatrix} Y_{11} & Y_{12} & Y_{13} & Y_{14} \\ Y_{21} & Y_{22} & Y_{23} & Y_{24} \\ Y_{31} & Y_{32} & Y_{33} & Y_{34} \\ Y_{41} & Y_{42} & Y_{43} & Y_{44} \end{bmatrix} \quad \ldots\ldots(21)$$

Equation (20) having $n = 4$ reduces to

$$\underleftarrow{[Y]} = \frac{1}{\begin{vmatrix} Y_{33} & Y_{34} \\ Y_{43} & Y_{44} \end{vmatrix}} \left\{ \begin{bmatrix} Y_{11} & Y_{12} \\ Y_{21} & Y_{22} \end{bmatrix} \begin{vmatrix} Y_{33} & Y_{34} \\ Y_{43} & Y_{44} \end{vmatrix} - \begin{bmatrix} Y_{13} & Y_{14} \\ Y_{23} & Y_{24} \end{bmatrix} \begin{bmatrix} Y_{44} & -Y_{34} \\ -Y_{43} & Y_{33} \end{bmatrix} \begin{bmatrix} Y_{31} & Y_{32} \\ Y_{41} & Y_{42} \end{bmatrix} \right\} \quad \ldots\ldots(22)$$

$$\underleftarrow{[Y]} = \frac{1}{\begin{vmatrix} Y_{33} & Y_{34} \\ Y_{43} & Y_{44} \end{vmatrix}} \begin{bmatrix} \begin{matrix} Y_{11}Y_{33}Y_{44} - Y_{11}Y_{34}Y_{43} - \\ -Y_{13}Y_{44}Y_{31} + Y_{14}Y_{43}Y_{31} + \\ +Y_{13}Y_{34}Y_{41} - Y_{14}Y_{33}Y_{41} \end{matrix} & \begin{matrix} Y_{12}Y_{33}Y_{44} - Y_{12}Y_{34}Y_{43} - \\ -Y_{13}Y_{44}Y_{32} + Y_{14}Y_{43}Y_{32} + \\ +Y_{13}Y_{34}Y_{42} - Y_{14}Y_{33}Y_{42} \end{matrix} \\ \begin{matrix} Y_{21}Y_{33}Y_{44} - Y_{21}Y_{34}Y_{43} - \\ -Y_{31}Y_{23}Y_{44} + Y_{31}Y_{43}Y_{24} + \\ +Y_{23}Y_{34}Y_{41} - Y_{24}Y_{33}Y_{41} \end{matrix} & \begin{matrix} Y_{22}Y_{33}Y_{44} - Y_{22}Y_{34}Y_{43} - \\ -Y_{32}Y_{23}Y_{44} + Y_{43}Y_{24}Y_{32} + \\ +Y_{23}Y_{34}Y_{42} - Y_{24}Y_{33}Y_{42} \end{matrix} \end{bmatrix} \quad \ldots\ldots(23)$$

Equation (23) can be rewritten in the form

$$\underleftarrow{[Y_E]} = \frac{1}{K} \begin{bmatrix} Y_{11E} & Y_{12E} \\ Y_{21E} & Y_{22E} \end{bmatrix} \quad \ldots\ldots(24)$$

The number of terms in each element of the equivalent matrix is large even when n is as low as 4. Thus, as n increases the terms would become cumbersome. A more conventional approach would be equally cumbersome and more laborious.

4.1. *Reduction by Inspection*

Setting up a nodal matrix from an indefinite admittance matrix is a simple process of removing appropriate rows and columns as discussed in Section 2. A similar approach of eliminating rows and columns can be applied to the determinant of the original matrix to obtain the elements of the equivalent 2×2 matrix.

Y_{11E} relates to the input and is formed by removing row 2 and column 2 corresponding to the output node. In a similar way, Y_{22E} which relates to the output has row 1 and column 1 removed.

4.2. *Statement of Multi-node Method*

A multi-node network having two ports of interest can be represented by an equivalent 2×2 matrix relevant to the ports of interest and having the form:

$$\underleftarrow{[Y_E]} = \frac{1}{K} \begin{bmatrix} Y_{11E} & Y_{12E} \\ Y_{21E} & Y_{22E} \end{bmatrix} \quad \ldots\ldots(25)$$

The terms of the equivalent matrix can be found directly by considering the determinant of the original matrix with appropriate rows and columns removed as shown in Table 1.

(Application of the results to eqn. (21) again yields eqn. (23).)

Table 1

	To be removed	
Term	Column	Row
K	1, 2	1, 2
Y_{11E}	2	2
Y_{12E}	1	2
Y_{21E}	2	1
Y_{22E}	1	1

4.3. *Convention*

The validity of the method depends on the input and output nodes being selected as nodes 1 and 2 respectively. The remaining nodes may be selected in an arbitrary manner.

5. Numerical Verification

The simplicity of the method is best seen by example. Consider the cathode coupled amplifier of Fig. 3(a).

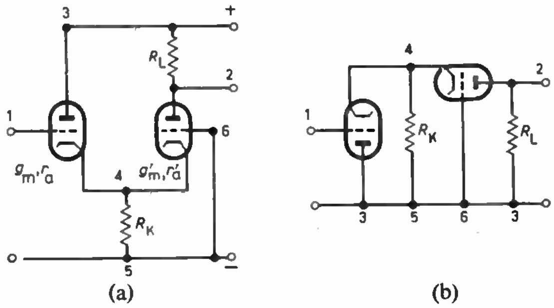


Fig. 3. Circuit for cathode-coupled amplifier and its cascade equivalent.

The indefinite admittance matrix can be written down by inspection

$$[\underleftarrow{Y_i}] = \begin{array}{c|cccccc|} & 1 & 2 & 3 & 4 & 5 & 6 \\ \hline 1 & 0 & 0 & 0 & 0 & 0 & 0 \\ 2 & 0 & \frac{1}{r'_a}+\frac{1}{R_L} & -\frac{1}{R_L} & -\frac{1}{r'_a}-g'_m & 0 & g'_m \\ 3 & g_m & -\frac{1}{R_L} & \frac{1}{r_a}+\frac{1}{R_L} & -g_m-\frac{1}{r_a} & 0 & 0 \\ 4 & -g_m & -\frac{1}{r'_a} & -\frac{1}{r'_a} & \frac{1}{r'_a}+g'_m+\frac{1}{R_K}+g_m+\frac{1}{r_a} & -\frac{1}{R_K} & -g'_m \\ 5 & 0 & 0 & 0 & -\frac{1}{R_K} & \frac{1}{R_K} & 0 \\ 6 & 0 & 0 & 0 & 0 & 0 & 0 \end{array} \qquad \ldots\ldots(26)$$

With respect to a.c., nodes 3, 5 and 6 are common and if taken as the reference then the nodal matrix may be obtained by removing rows 3, 5 and 6 and columns 3, 5 and 6. The result is a 3×3 nodal matrix given by:

$$[\underleftarrow{Y}] = \begin{bmatrix} 0 & 0 & 0 \\ 0 & \frac{1}{r'_a}+\frac{1}{R_L} & -\frac{1}{r'_a}-g'_m \\ -g_m & -\frac{1}{r'_a} & \frac{1}{r'_a}+g'_m+\frac{1}{R_K}+g_m+\frac{1}{r_a} \end{bmatrix} \qquad \ldots\ldots(27)$$

The equivalent 2×2 matrix is obtained using the results of Table 1.

$$[\underleftarrow{Y_E}] = \frac{1}{\frac{1}{r'_a}+g'_m+\frac{1}{R_K}+g_m+\frac{1}{r_a}} \begin{bmatrix} 0 & 0 \\ g_m\left(-g'_m-\frac{1}{r'_a}\right) & \left(\frac{1}{r'_a}\right)\left(-\frac{1}{r'_a}-g'_m\right)+ \\ & +\left(\frac{1}{r'_a}+\frac{1}{R_L}\right)\left(\frac{1}{r'_a}+g_m+\frac{1}{R_K}+\frac{1}{r_a}\right) \end{bmatrix} \qquad \ldots(28)$$

The voltage gain from eqn. (4a) is $-\dfrac{Y_{21E}}{Y_{22E}}$

$$\text{gain} = \frac{\mu(\mu'+1)R_L R_K}{R_L\{(1+\mu)R_K+r_a\}+R_K\{r_a(\mu'+1)+r'_a(1+\mu)\}+r_a r'_a} \qquad \text{......(29)}$$

This result can be verified from text books.

As a second example, consider the transistor amplifier of Fig. 4(a) in which the transistor parameters are as shown in Table 2.

Table 2

	h_{11}	h_{12}	h_{21}	h_{22}
TR1	2400	$4{\cdot}04 \times 10^{-4}$	37·5	$13{\cdot}85 \times 10^{-6}$
TR2	281·8	$0{\cdot}7613 \times 10^{-4}$	58·44	$136{\cdot}4 \times 10^{-6}$

The corresponding Y parameters are derived from the given matrix conversions.

Table 3

	Y_{11}	Y_{12}	Y_{21}	Y_{22}
TR1	$4{\cdot}166 \times 10^{-4}$	$-1{\cdot}684 \times 10^{-7}$	$1{\cdot}563 \times 10^{-2}$	$7{\cdot}712 \times 10^{-6}$
TR2	$3{\cdot}549 \times 10^{-3}$	$-2{\cdot}702 \times 10^{-7}$	0·2074	$1{\cdot}206 \times 10^{-4}$

The indefinite admittance matrix can now be set up.

$$[\underleftarrow{Y_i}] = \begin{array}{cc} & \begin{array}{ccccc} 1 & 2 & 3 & 4 & 5 \end{array} \\ \begin{array}{c} 1 \\ 2 \\ 3 \\ 4 \\ 5 \end{array} & \left[\begin{array}{ccccc} (Y_{11})_1 & 0 & (Y_{12})_1 & -(Y_{11})_1-(Y_{12})_1 & 0 \\ 0 & (Y_{11})_2+(Y_{12})_2+(Y_{21})_2+(Y_{22})_2+\dfrac{1}{470} & -(Y_{11})_2-(Y_{21})_2 & -\dfrac{1}{470} & -(Y_{12})_2-(Y_{22})_2 \\ (Y_{21})_1 & -(Y_{11})_2-(Y_{12})_2 & (Y_{22})_1+(Y_{11})_2+\dfrac{1}{10^4} & -(Y_{21})_1-(Y_{22})_1 & (Y_{12})_2-\dfrac{1}{10^4} \\ -(Y_{11})_1-(Y_{21})_1 & -\dfrac{1}{470} & -(Y_{12})_1-(Y_{22})_1 & (Y_{11})_1+(Y_{12})_1+(Y_{21})_1+(Y_{22})_1+\dfrac{1}{470} & 0 \\ 0 & -(Y_{21})_2-(Y_{22})_2 & (Y_{21})_2-\dfrac{1}{10^4} & 0 & (Y_{22})_2+\dfrac{1}{10^4} \end{array}\right] \end{array} \qquad \text{...(30)}$$

where $(\ )_1$ and $(\ )_2$ relate to transistors TR1 and TR2 respectively.

Substituting numerical values the following result can be obtained:

$$[Y_1] = \begin{array}{c|ccccc|} & 1 & 2 & 3 & 4 & 5 \\ \hline 1 & 4\cdot166\times10^{-4} & 0 & -1\cdot684\times10^{-7} & -4\cdot164\times10^{-4} & 0 \\ 2 & 0 & 2\cdot128\times10^{-3}+0\cdot21105 & -0\cdot2109 & -2\cdot128\times10^{-3} & -1\cdot206\times10^{-4} \\ 3 & 1\cdot563\times10^{-2} & -3\cdot549\times10^{-3} & 7\cdot712\times10^{-6}+10^{-4}+3\cdot549\times10^{-3} & -1\cdot563\times10^{-2} & -10^{-4}-2\cdot702\times10^{-7} \\ 4 & -1\cdot563\times10^{-2} & -2\cdot128\times10^{-3} & -7\cdot7\times10^{-6} & 2\cdot128\times10^{-3} & 0 \\ 5 & 0 & -0\cdot2075 & 0\cdot2074-10^{-4} & 0 & 10^{-4}+1\cdot206\times10^{-4} \end{array} \quad \ldots(31)$$

A 3×3 matrix results when nodes 4 and 5 which are common with respect to a.c. are taken as a reference.

$$[Y] = \begin{bmatrix} 4\cdot166\times10^{-4} & 0 & -1\cdot684\times10^{-7} \\ 0 & 2\cdot132\times10^{-1} & -2\cdot109\times10^{-1} \\ 1\cdot563\times10^{-2} & -3\cdot549\times10^{-3} & 3\cdot657\times10^{-3} \end{bmatrix} \quad \ldots\ldots(32)$$

The gain is again found from eqn. (4a), and if this is the only requirement of the analysis then only two terms of the equivalent matrix are required.

Applying the rules of Table 1 to the matrix of eqn. (27), we get

$$Y_{21E} = \begin{vmatrix} \multicolumn{2}{c}{\text{Removed row and column}} \\ 0 & -2\cdot109\times10^{-1} \\ 1\cdot563\times10^{-2} & 3\cdot657\times10^{-3} \end{vmatrix}$$

$$= \begin{vmatrix} 0 & -2\cdot109\times10^{-1} \\ 1\cdot563\times10^{-2} & 3\cdot657\times10^{-3} \end{vmatrix}$$

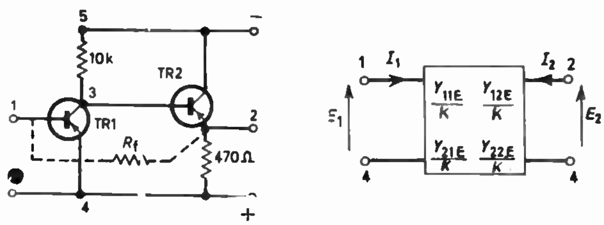
$$= 3\cdot296\times10^{-3}$$

$$Y_{22E} = \begin{vmatrix} \multicolumn{2}{c}{\text{Removed row and column}} \\ 2\cdot132\times10^{-1} & -2\cdot109\times10^{-1} \\ -3\cdot549\times10^{-3} & 3\cdot657\times10^{-3} \end{vmatrix}$$

$$Y_{22E} = 3\cdot12\times10^{-5}$$

$$\text{Gain} = \frac{-Y_{21E}}{Y_{22E}} = -105\cdot6$$

This example can be verified by an alternative solution.[2]



(a) (b)

Fig. 4. A simple amplifier circuit using transistors and its two-port representation.

6. Conclusions

The following advantages are achieved by this method.

(1) The same procedure is used for all networks, active or passive, and eliminates the need to know particular mathematical manipulations which may be necessary with certain networks.

(2) The matrix representation is required for only one mode of operation. In example 1, only the common-cathode mode was required. However, if the more conventional method had been used both the common-anode and the common-grid modes would have been needed.

In example 2 the common-emitter mode was used whereas this, together with the common-collector mode would be needed for the more conventional method.[2]

(3) It is easy to change the reference node and the input and output nodes.

(4) Additional circuit elements may be connected between the nodes without changing the procedure. For example, if feedback is applied to the circuit of Fig. 4 (a) by connecting a resistor R_f between nodes 1 and 2 the following simple additions would be made to matrix (30).

$$\begin{array}{c|cc} & 1 & 2 \\ \hline 1 & \frac{1}{R_f} & -\frac{1}{R_f} \\ 2 & -\frac{1}{R_f} & \frac{1}{R_f} \end{array} \qquad \ldots\ldots(33)$$

More conventional methods require much more work and in many cases need an interchange of parameters from say the (a) matrix to the (Y) matrix as shown in Reference 2.

(5) The method eliminates the problems associated with singular nodal matrices when used in particular two-port interconnections.

Consider the circuit shown in Fig. 5 with its series two-port equivalent. The rules for series interconnection are given by:

$$[Z_T]=[Z_1]+[Z]_2 \qquad \ldots\ldots(34)$$

The admittance matrix for the valve, as given by eqn. (5), is singular thus the inverse matrix (impedance matrix) does not exist. A 2×2 matrix from eqns. (5) and (34) is therefore not possible.

This problem is overcome by the use of the indefinite admittance matrix in conjunction with the method described.

The indefinite admittance matrix for the valve with cathode resistor R_K is

$$[\underleftarrow{Y_i}] = \begin{bmatrix} 0 & 0 & 0 & 0 \\ g_m & \frac{1}{r_a} & -g_m - \frac{1}{r_a} & 0 \\ -g_m & -\frac{1}{r_a} & g_m + \frac{1}{r_a} + \frac{1}{R_K} & -\frac{1}{R_K} \\ 0 & 0 & -\frac{1}{R_K} & \frac{1}{R_K} \end{bmatrix} \qquad \ldots\ldots(35)$$

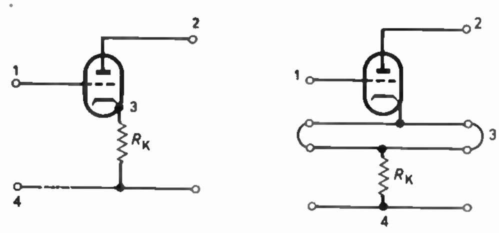


Fig. 5. A valve amplifier circuit and its two-port equivalent.

This can now be reduced to a 2×2 matrix by the method given and results in

$$[\underleftarrow{Y}] = \begin{bmatrix} 0 & 0 \\ \dfrac{g_m}{1+R_K\left(g_m+\dfrac{1}{r_a}\right)} & \dfrac{1/r_a}{1+R_K\left(g_m+\dfrac{1}{r_a}\right)} \end{bmatrix} \qquad \ldots\ldots(36)$$

(6) The procedure, being applicable to a variety of circuits, indicates that only the principle requires teaching time, the applications to particular networks being the subject of tutorials.

(7) All the terminal characteristics can be obtained from the equivalent 2×2 matrix $[Y_E]$ without additional analysis by using the following equations:

$$\text{voltage transfer function} = -\frac{Y_{21E}}{Y_{22E}} \qquad \ldots\ldots(37a)$$

$$\text{input impedance} = \frac{K\,Y_{22E}}{Y_{11E}Y_{22E}-Y_{12E}Y_{21E}} \qquad \ldots\ldots(37b)$$

$$\text{output impedance} = \frac{K}{Y_{22E}} \qquad \ldots\ldots(37c)$$

A separate analysis for each term is often required with the more traditional methods.

7. Acknowledgment

The author wishes to thank his colleague, Mr. L. Hoare, for valuable discussions.

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The Maxson Multi-beam Antenna: Theory and Design for Non-interacting Beams

By
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Summary: A multi-beam antenna known as the Maxson antenna is described. It consists of a linear array of elements fed in a serial fashion by a matrix of waveguide directional couplers. The antenna designed produced two 1° beams and operated over a 10% bandwidth at C-band. The feeder coupling matrix was designed to produce no interaction between beams. Limits on orthogonal and non-orthogonal beam-shapes and beam-positions are derived. Applications of this type of antenna are discussed.

List of Symbols

V	normalized voltage
P	input voltage to feeder 1
p_r	voltage at antenna element r due to P
P_{L1}	voltage at load terminating feeder 1 due to P
Q	input voltage to feeder 2
q_r	voltage at antenna element r due to Q
q_{L1}	voltage at load terminating feeder 1 due to Q
q_{L2}	voltage at load terminating feeder 2 due to Q
θ_r	phase angle between p_r and q_r
θ_{L1}	phase angle between p_{L1} and q_{L1}
q_r^*	complex conjugate of q_r
q_{L1}^*	complex conjugate of q_{L1}
$G_p(Z)$	far-field pattern due to P
$G_q(Z)$	far-field pattern due to Q
Z	phase increment between antenna elements
d	distance between elements
θ	angle between radiated beam and aperture normal
λ	free space wavelength
C	coupling factor of directional coupler
ϕ	phase lag suffered by wave passing over directional coupler
$A^2(n)$	power at nth antenna element
N	total number of antenna elements
$Z_{0,1,2}$	impedance
β	propagation constant
l	matching length of dielectric on phase insert
w	width of phase insert
$\phi_{1(2r)}$	calculated phase length of phase insert $(2r)$
$\phi'_{1(2r)}$	total phase length of phase insert $(2r)$
λ_g	waveguide wavelength
$A(\bar{x})$	aperture distribution
$\bar{x}$	normalized aperture co-ordinate
a	constant
η^2	feeder efficiency
u	continuous phase variation across aperture

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1. Introduction

The advent of high-speed aircraft and missiles has resulted in the need for high-data-rate, high-resolution radars. These demands can be met by electronically scanned antennas or multi-beam antennas. This paper considers a multi-beam antenna first described by Blass[1] and now commonly described as the Maxson antenna. It consists of a linear array of elements fed by a matrix of waveguide directional couplers (Fig. 1). The input power is serially-coupled to the array element as opposed to the alternative parallel feeder proposed by Butler[2] (Fig. 2). The fundamental restrictions on beam-shapes and beam-positions are the same for either feeder but the Maxson feeder is more suitable when a limited number of beams are required from an aperture containing many array elements. For efficient use the Butler feeder should produce as many beams as array elements and must contain an even number of antenna elements. When therefore one requires fewer beams than the number of array elements the Maxson array is a more economic arrangement than the Butler matrix array.

The initial concept of a series-fed r.f. beam-former was first described by Blass[1] in 1960. Further experimental and theoretical work on this topic was undertaken by Blass and his colleagues and a series of reports[3] were issued. This beam-former was also theoretically studied at the Lincoln Laboratories under their phased array radar programme.[4,5] The author conducted the present work during the period October 1964 to September 1966.

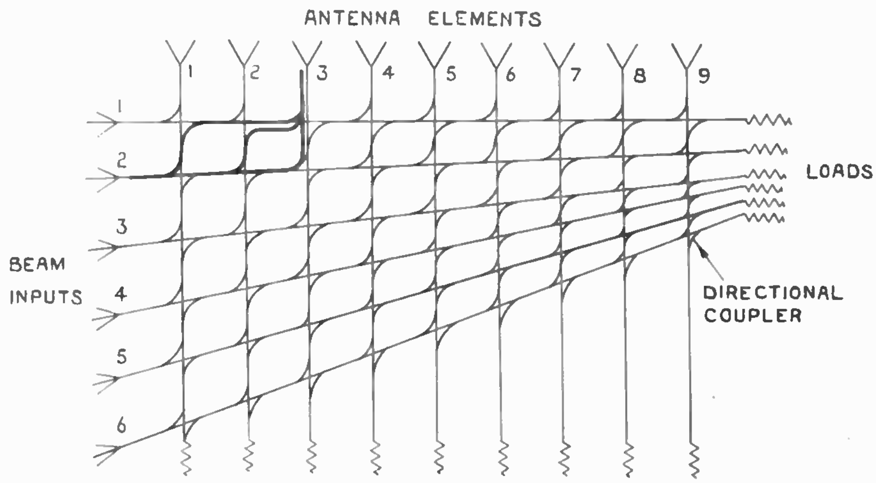

Fig. 1. Maxson matrix.

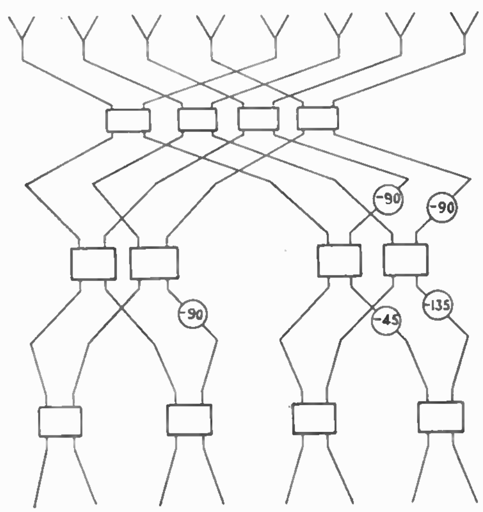

Fig. 2. Butler matrix.

2. Multiple-beam Formation

The formation of multiple-beams is best described with reference to Fig. 1. The basic elements of the structure are the cross-guide directional couplers, which are arranged so that power fed into any horizontal feeder arm is essentially serially coupled to the antenna elements via the cross-arms. The remaining power in the feeder is dissipated in the terminating load of the feeder. No power reaches the loads terminating the cross-arms due to the directivity of the couplers.

In the case of feeder 1 the antenna elements are fed in a purely serial fashion. Feeder 2 and succeeding feeders, although routing the majority of the power in a serial fashion, in general possess several secondary parallel paths between feeder input and antenna element. For example, there are one primary and two secondary routes between input 2 and antenna element

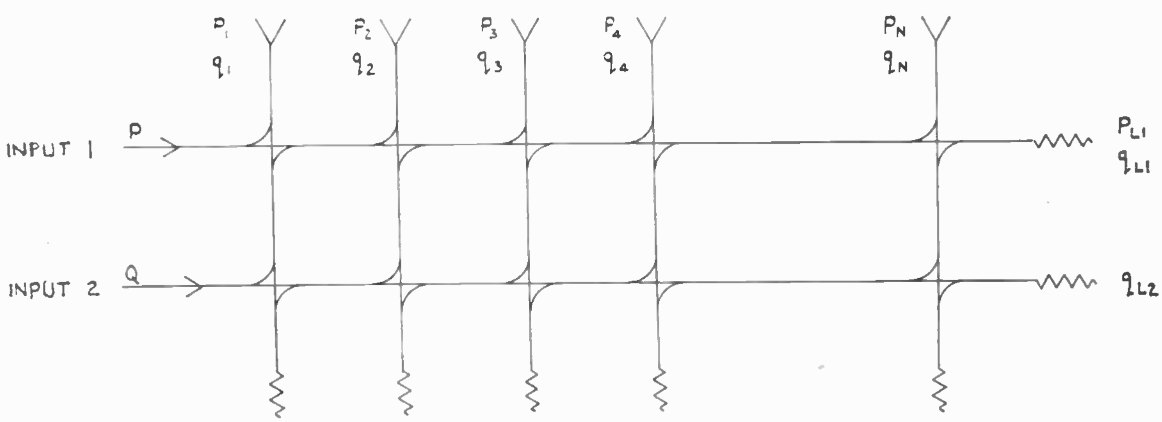

Fig. 3. Voltages produced by feeding inputs 1 and 2.

3. The primary route involves only one coupling from feeder to cross-arm whilst the secondary routes involve three couplings and consequently are weaker.

Irrespective of the routes followed, exciting an input feeder results in the majority of the power reaching the antenna elements and radiating into space. Now if it is arranged that power fed into a feeder arm produces an array aperture distribution possessing a linear phase gradient the radiation into space forms a definite beam. If different feeders produce different phase gradients then to each feeder input can be allotted a beam position in space. This structure is thus capable of simultaneously producing multiple-beams in space. The physical constraints on the beam patterns and beam positions are discussed in the following Sections.

3. Constraints in Multiple-beam Design

Several writers[5-7] have shown that there are fundamental constraints on the production of multi-beams from a common aperture. These constraints apply equally to quasi-optical methods used to produce the beams as to the waveguide beam-former discussed in this report. A non-rigorous derivation, intended solely as a guide to the appreciation of these constraints, is now formulated for the Maxson antenna structure.

Consider any two feeder arms, say 1 and 2, and assume that the feeder structure is completely lossless and matched. Hence we will assume that any power fed into 1 or 2 will always arrive at the antenna elements or feeder loads. Let us adopt the normalized voltage concept[8] to study the flow of energy in the waveguide structure. The normalized voltage V, is chosen so that $\frac{1}{2}|V|^2$ represents the power in the guide and V is complex so that relative phase information is retained.

If input 1 is fed with a voltage P it will result in a set of voltages $p_1, p_2, p_3 \ldots p_N$, arriving at the antenna radiators, also a voltage p_{L1} arrives at the load terminating feeder 1. Similarly a voltage Q fed into feeder 2 results in the set of voltages, $q_1, q_2, q_3 \ldots, q_N$, at the antenna radiators but this time there is the possibility of voltages q_{L2} and q_{L1} in the feeder loads (see Fig. 3).

Now, neglecting the $\frac{1}{2}$ on either side of the relationship, the principle of conservation of energy gives

$$|Q|^2 = \sum_{r=1}^{N} |q_r|^2 + |q_{L1}|^2 + |q_{L2}|^2 \qquad \ldots\ldots(1)$$

and

$$|P|^2 = \sum_{r=1}^{N} |p_r|^2 + |p_{L1}|^2 \qquad \ldots\ldots(2)$$

when the feeders are excited independently.

Exciting the two feeders simultaneously results in a superposition of these voltage sets on the antenna terminals and in the loads.

Now

$$\begin{aligned}|P|^2 + |Q|^2 &= \sum_{r=1}^{N} |(p_r + q_r)|^2 + |(p_{L1} + q_{L1})|^2 + |q_{L2}|^2 \\ &= \sum_{r=1}^{N} |p_r|^2 + \sum_{r=1}^{N} |q_r|^2 + 2\sum_{r=1}^{N} |p_r| \cdot |q_r| \cos\theta_r + \\ &\quad + |p_{L1}|^2 + |q_{L1}|^2 + |q_{L2}|^2 + \\ &\quad + 2|p_{L1}| \cdot |q_{L1}| \cos\theta_{L1} \qquad \ldots\ldots(3)\end{aligned}$$

where θ_r and θ_{L1} are the phase angles between the superposed voltages.

Comparing eqns. (1), (2) and (3), it is seen that for conservation of energy

$$\sum_{r=1}^{N} |p_r| \cdot |q_r| \cos\theta_r = -|p_{L1}| \cdot |q_{L1}| \cos\theta_{L1} \qquad \ldots\ldots(4)$$

This could be written as

$$\text{Re} \sum_{r=1}^{N} p_r . q_r^* = \text{Re}\ p_{L1} . q_{L1}^*$$

The asterisk denotes the complex conjugate.

Note that for the case of no power dissipated in the loads

$$p_{L1} = q_{L1} = q_{L2} = 0$$

Then

$$\sum_{r=1}^{N} p_r . q_r^* = 0$$

This result is significant for it states that a completely lossless multiple beam feeder structure can be produced but the sum of the product of cross-coupled fields over the aperture must be zero. Alternatively by using the Wiener–Khintchine theorem, this result can be related to the far-field beam patterns.

The far-field patterns are linked to the aperture field distributions by the Fourier relationships

$$G_1(Z) = \sum_{r=1}^{N} p_r \exp(\mathrm{j}rZ)$$

$$G_2(Z) = \sum_{s=1}^{N} q_s \exp(\mathrm{j}sZ)$$

where $Z = \dfrac{2\pi d}{\lambda} \sin\theta$; d is element spacing, λ the free space wavelength, and θ the angle the radiated beam makes with the aperture normal.

$$G_1(Z) . G_2^*(Z) = \sum_{r=1}^{N} p_r \exp(\mathrm{j}rZ) . \sum_{s=1}^{N} q_s^* \exp(-\mathrm{j}sZ)$$

$$\begin{aligned}\int_{-\infty}^{+\infty} G_1(Z) . G_2(Z)\,\mathrm{d}Z &= \sum_{r=1}^{N} p_r \sum_{s=1}^{N} q_s^* \int_{-\infty}^{+\infty} \exp\{\mathrm{j}(r-s)Z\}\,\mathrm{d}Z \\ &= \sum_{r=1}^{N} p_r \sum_{s=1}^{N} q_s^* . \delta(r-s) = \sum_{r=1}^{N} p_r . q_r^* \qquad \ldots\ldots(5)\end{aligned}$$

It is in terms of the beam patterns that the completely lossless result is usually quoted. The statement

$$\mathrm{Re} \int_{-\infty}^{+\infty} G_1(Z).G_2^*(Z)\,\mathrm{d}Z = 0$$

implies that beam pattern $G_1(Z)$ is orthogonal to beam pattern $G_2(Z)$.

This is an important constraint on multiple beam design using a common aperture as without its knowledge many designers had tried to achieve impossible systems possessing non-orthogonal beams and high feeder efficiency.

The value of

$$\mathrm{Re} \int_{-\infty}^{+\infty} G_1(Z).G_2(Z)\,\mathrm{d}Z$$

tends to become smaller as the two beams are spaced further apart. The cross-over levels of adjacent beams must not be too low or it leads to holes in the radar coverage afforded by the multi-beams. It is found that for orthogonal spacing of the beams, a high cross-over level is unfortunately associated with high side-lobe patterns. A flat amplitude distribution over the aperture generates a beam pattern of the form $\sin u/u$ (u is defined in Appendix 2). These beams are orthogonal when displaced $u = \pm\pi/2$ and possess a cross-over level of $-3{\cdot}92$ dB. This aspect together with the design of non-orthogonal beams is discussed in Section 8.

4. Maxson Feeder Design Philosophy

The constraints on multiple-beam design were discussed in terms of the amplitude distribution resulting across the aperture when a feeder was excited. Whilst this discussion showed the limits set on the cross-coupled aperture distribution it did not suggest a method of designing for those aperture distributions.

Blass's approach to the design problem was to consider only the primary routes through the waveguide matrix to produce the desired pattern. Each feeder arm possessed coupling coefficients identical to the first and positioning of the beams was performed by altering the angle the feeder line made with the cross-arms. He then analysed the resulting structure and estimated the undesirable effects produced by the secondary routes through the matrix. It is only when appreciable power is wasted in the feeder end loads that this approach is a good approximation to the desired aperture distributions.

It was realized by Bernella and Pratt[4] and later, but independently, by the author that, if all routes between antenna element and feeder input were considered, it should be possible to design exactly for the desired aperture distributions and produce non-interacting beams. Due to the numerous computations involved in the design of an array with a large number of elements a digital computer was used to perform the calculations. A difference in the design procedure of Bernella and Pratt and the author was that they designed the antenna on a reception basis whilst the author designed on a transmit basis.

5. Design Procedure

Since the feeder consists of numerous directional couplers some aspects of these couplers will be discussed. The directional couplers are assumed to be perfectly matched which implies that they also possess perfect directivity.

It is shown in Appendix 1 that a wave travelling past a perfect directional coupling is not only reduced in amplitude depending on the coupling power C, but also suffers a phase lag ϕ, where $\phi = \sin^{-1}\sqrt{C}$. The coupled wave is phase advanced by the amount $\pi/2-\phi$. No experimental work has been performed to verify this phase angle law but it fits the experimental results given in Reference 3 very closely.

The performance of a directional coupler when excited by two input (i.e. uncoupled) arms simultaneously is obtained by taking a linear superposition of the excitations produced when fed independently. It is thus seen that under all conditions of relative amplitude and phase of the two inputs, there is no mismatch at the input arms. The power distribution at the output arms depends on both amplitude and phase at the input arms. Also knowing the amplitudes and phases of the voltages at any two ports of the coupler the amplitude and phase at the other two ports can be predicted.

The design of the coupling coefficients and phase inserts necessary between feeder lines for the six-beam, six-element array shown in Fig. 4 can now be described. The six beams are orthogonally-spaced $\sin u/u$ patterns, each requiring a flat amplitude distribution. Apart from the phase shift introduced by the couplers the design of feeder 1 is little different from a slotted waveguide array design. The power coupling coefficients are calculated from the relationship

$$C_{1r} = \mathrm{A}^2(r) \Bigg/ \left(P_1' + \sum_{n=r}^{N} \mathrm{A}^2(n)\right)$$

where $\mathrm{A}^2(r)$ is the power distribution required for the rth antenna element.

$$P_1' = \frac{P_1}{100-P_1} \cdot \sum_{n=0}^{N} \mathrm{A}^2(n)$$

where P_1 is the percentage of the total input power dissipated in the end load and N is the total number of antenna elements, in this case six. For example, a uniform distribution with 4% of the input power

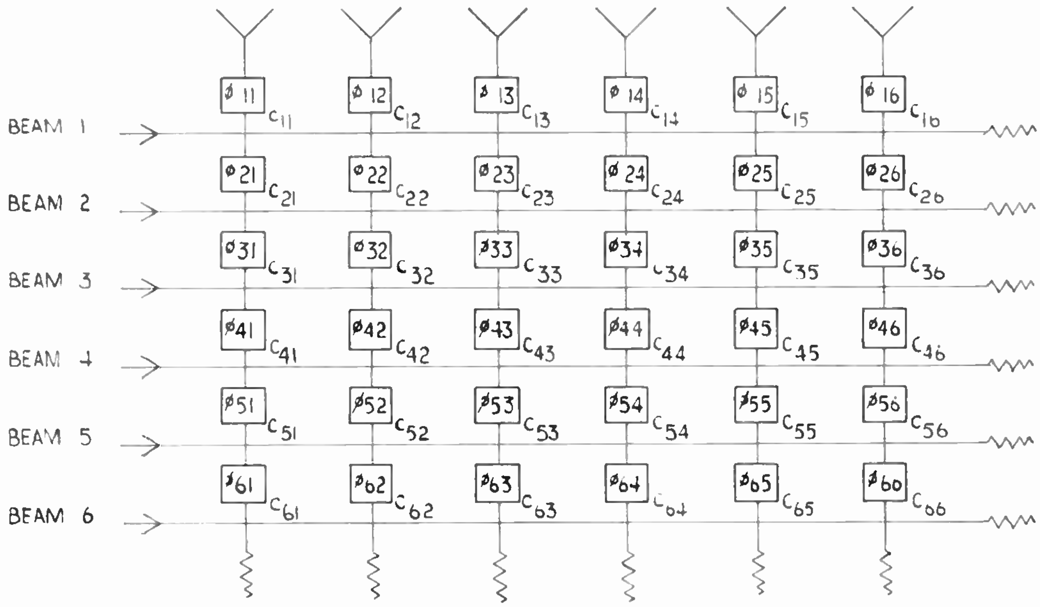

Fig. 4. 6×6 Maxson matrix.

dissipated in the load has

$$C_{13} = 1\Big/\left(\frac{4}{96}\times 6+\sum_{n=3}^{6} A^2(n)\right) = 1\Big/\left(\frac{24}{96}+4\right) = 0{\cdot}235$$

Expressed in decibels $C_{13} = -6{\cdot}28$ dB.

The phase shift past each coupler is calculated by using $\phi = \sin^{-1}\sqrt{C_{ij}}$, e.g. $\phi_{13} = 29° \ 6'$.

The phase front resulting over the aperture when feeder 1 is excited is plotted in Fig. 5 and is given by

$$\sum_{n=1}^{r} \phi_{1n} \text{ at the } r\text{th element.}$$

This phase front is corrected by means of dielectric phase inserts so that the resulting phase front is linear. The values of extra phase inserted are equal to the vertical lines drawn in Fig. 5. This completes the design of feeder 1. Feeder 2 can now be designed to feed through feeder 1 and still produce the correct aperture distribution required.

Again a flat amplitude distribution is required but the linear phase front of beam 2 must have a steeper slope than beam 1. It should exceed the total phase difference across the aperture of beam 1 by 2π radians. This ensures that the beams are orthogonal in space. Feeder 2 and all succeeding feeder arms are designed sequentially from input to load. The arms of any coupler will be labelled N, S, E, W. Consider cross-arm 1, the required amplitude and phase at antenna element is known, there is no secondary path to allow power to flow into feeder 1. Therefore the amplitudes and phase at the inputs, to the coupler C_{11} can be calculated. Also the amplitude and phase of energy at arm W of C_{12} can be found. This will be needed when dealing with cross-arm 2. Returning to cross-arm 1, the input power to coupler C_{21} is of course 100% and the coupled output into arms of C_{11} is known in amplitude and phase. Hence the value of C_{21} and the phase insert ϕ_{21} can be calculated. Also the amplitude and phase of feeder power travelling down feeder 2 past C_{21} can be found.

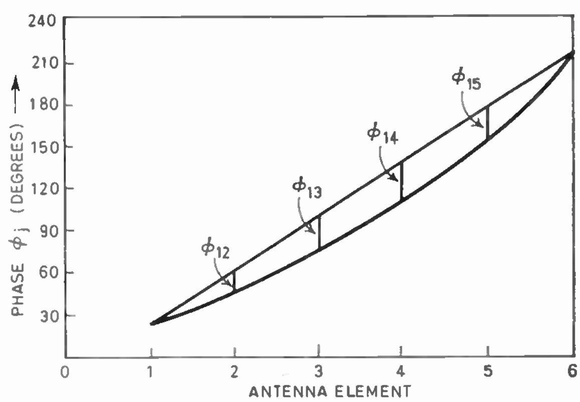

Fig. 5. First beam phase correction.

Now considering cross-arm 2, the amplitude and phase at arm N of coupler C_{12} is known from the required aperture distribution, also the secondary power at arm W has already been calculated. Thus the vector outputs at arm S and E of C_{12} can be found. Now for coupler C_{22} the input to arm W is known and

Table 1

Coupling coefficient matrix for 6×6 feeder producing orthogonal sin u/u beams with 4% dissipation in the loads. Coupling values are given in dB

Antenna element	1	2	3	4	5	6
Feeder arm 1	−7·96	−7·20	−6·28	−5·12	−3·52	−0·97
Feeder arm 2	−7·20	−5·24	−3·66	−2·63	−2·93	−3·52
Feeder arm 3	−6·28	−3·66	−2·72	−4·28	−2·63	−5·12
Feeder arm 4	−5·12	−2·63	−4·28	−2·72	−3·66	−6·28
Feeder arm 5	−3·52	−2·93	−2·63	−3·66	−5·24	−7·20
Feeder arm 6	−0·97	−3·52	−5·12	−6·28	−7·20	−7·96

the required output after phase-shift is known for arm N, thus C_{22} and ϕ_{22} can be calculated. This process is repeated for cross-arms 3 to 6. If the beams are phased for orthogonality it should be found that none of the power entered along feeder 2 reaches the load of feeder 1.

Feeders 3, 4, 5 and 6 are designed respectively in a similar fashion to feeder 2. The cross arms are taken in the order 1 to 6. In general the vector inputs or outputs are known at arms N and W at the top of a cross-arm, this enables the vectors at S and E to be calculated. The process is repeated down the cross-arm until the feeder arm being designed is reached. The problem is then a serial design problem for that feeder and can be solved to find the coupling coefficient and phase insert required.

The coupling coefficients necessary to produce a 6×6 orthogonal sin u/u system are given in Table 1. Each feeder arm dissipates 4% of its power into its own load but none into the other feeder loads. This is because the orthogonal spacing of the beams make the left hand side of equation 4 equal zero.

It should be noted that the coupling coefficients are doubly symmetrical so that $C_{ij} = C_{ji}$

$$\text{and } C_{ij} = C_{N+1-i,N+1-j}$$

The results were obtained using an autocode program on a digital computer. In this program the number of antenna elements and feeder arms can be varied but the size of the computer storage limits the coupler matrix to approximately 600 elements. It was decided to build a two-beam, 72-element antenna as an initial attempt to convert the computer design results into a practical antenna. This choice was based on the desire for a narrow beamwidth antenna and the limited time available for the building program.

6. Practical Design Considerations

The computer program was used to predict the coupling coefficients and extra phase shift required in the feeder matrix to produce two orthogonal, sin u/u beams. To achieve practically realizable coupling values the power dissipated in the load was made 4% of the input power. Graphs of the coupling values and phase inserts required for a 72-element array are plotted in Figs. 6 and 7 respectively. It can be seen that a range of coupling values of from −19 dB to −6 dB, and phase values from 0 to 283° are required.

6.1. *Directional Coupler*

The directional coupler used was the crossed-slot, broad-wall to broad-wall coupler illustrated in Fig. 8.

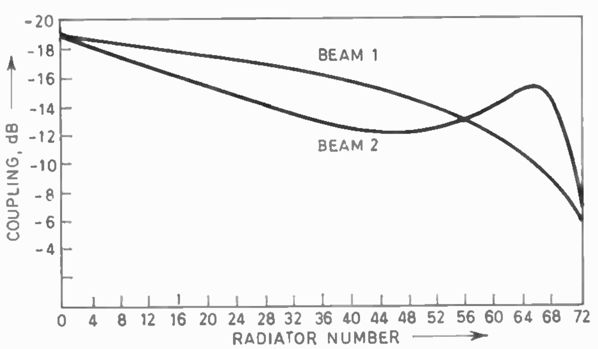

Fig. 6. Coupling coefficients for two beam matrix (flat aperture distribution), orthogonal beams 4% load power.

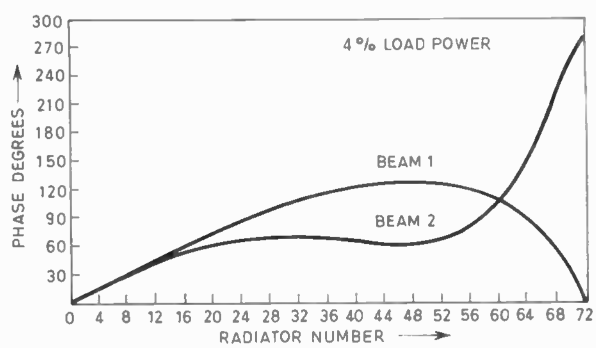

Fig. 7. Compensating phase for two-beam matrix.

The preferred directions of coupling are indicated by the arrows. The coupler can be analysed using the approach of Surdin,[9] but, for large coupling values embodying large slots no accurate theoretical estimate of coupling values to slot size can be made. An experimental programme was undertaken to evaluate coupling value to slot length, whilst the slot width was maintained constant. The feeder arm was in W.G.14, whilst the cross-arms were reduced height W.G.14. The wall currents are increased when guide height is reduced so that coupling becomes stronger. Also to increase the coupling values the slots were flush-filled with fluorosint dielectric which brought them closer to resonance. Even these measures were not quite enough to obtain a −6dB coupler, and for this coupling region large dumb-bell-shaped slots filled with variable thicknesses of fluorosint were used. The experimental results are shown in Figs. 9 and 10. The frequency variation of these couplers over a 10% bandwidth ranged from 1 dB for the weak couplers to 3 dB for the strongest. In general, directivity was better than 20 dB and v.s.w.r. better than 0·95. The worst performance was given by the −6 dB coupler which had a directivity

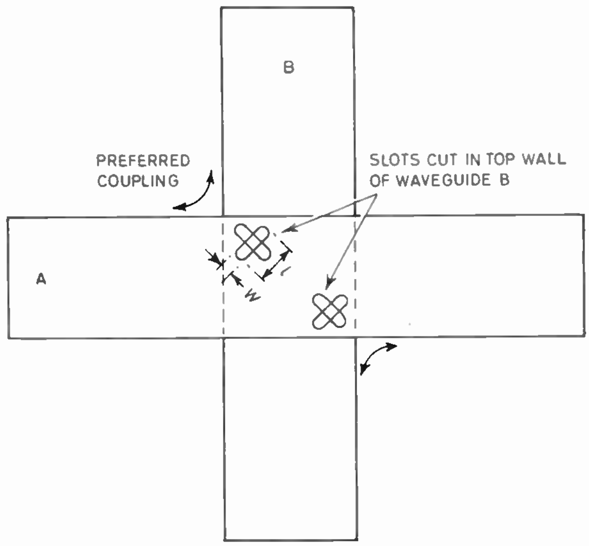

Fig. 8. Cross-slot coupler.

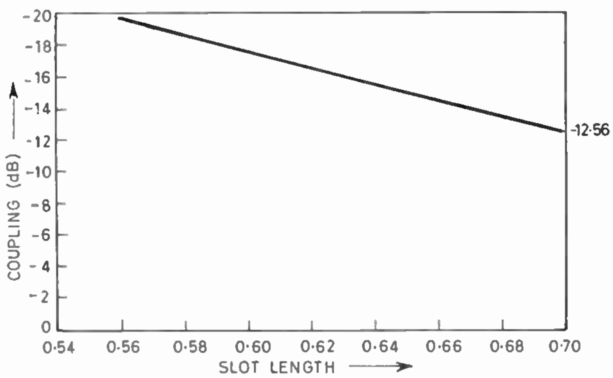

Fig. 9. Graph of experimental coupling versus slot length.

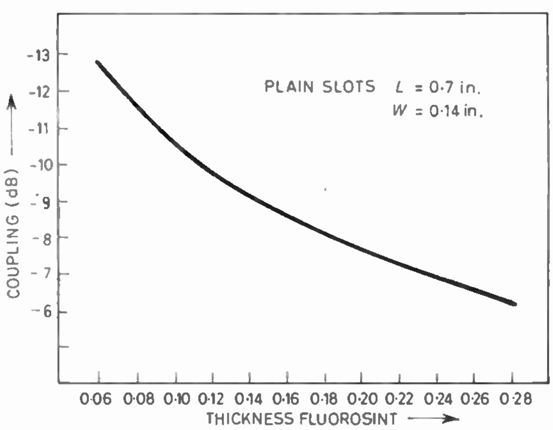

Fig. 10. Variation of coupling versus dielectric thickness.

of −14 dB and a v.s.w.r. of 0·7. From these experimental results the size of slots and thickness of fluorosint filling for the feeder arms could be predicted.

6.2. *Phase Inserts*

Extra phase lag was introduced in the cross-arms by means of polythene inserts, shaped as in Fig. 11. The protruding centre pieces are to effect a quarter-wave match between the completely dielectric-filled waveguide and the empty guide. The characteristic impedances of the dielectric-filled and the empty waveguide are Z_2 and Z_0 respectively.

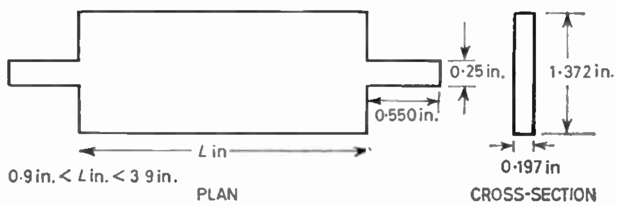

Fig. 11. Dielectric insert for reduced height W.G. 14.

A computer program provided a series of values of impedance, Z_1, and propagation constant, β, for the partially-filled guide as a function of width of dielectric. From this the width, w, which corresponded to Z_1 that satisfied $Z_1 = Z_0 Z_2$ was used. Then the length, l, of dielectric required is chosen to make $\beta l = \pi/4$. Using these values of l and w as starting values, experimental cut-and-try methods were used to achieve a match better than 0·8 over the 10% band. Phase measurements of the matched inserts were then taken on a waveguide phase bridge. The matched inserts had various lengths of completely filled dielectric. From these results a plot of phase shift against length of completely filled dielectric was obtained (Fig. 12). This was, of course, a linear relationship and was used to predict the lengths of solid section of the dielectric inserts. A small length was chosen to represent zero phase and the extra phase required added on to this small length. Thus even the first and last cross-arms above feeder 1 contained a dielectric insert.

6.3. *Construction*

The coupling slots were cut in pairs alternately in the top and bottom walls of the feeder arms. This eased the problems of soldering the cross-arms which have only 0·020 in (0·51 mm) horizontal separation. The cross-arms have a double E-plane bend so that they produce a linear array of open-ended waveguide elements at the aperture. Two strips of brass-angle were soldered at the aperture to provide a flange for mounting a flare.

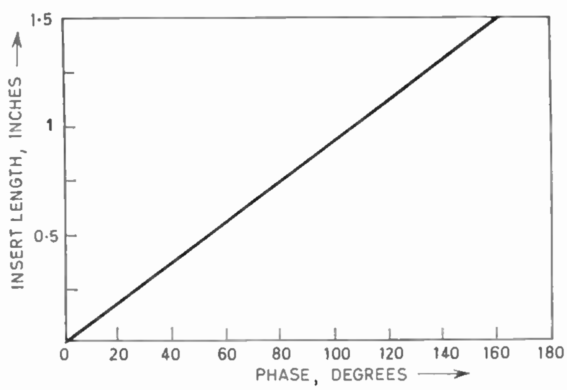

Fig. 12. Plot of phase shift versus dielectric length.

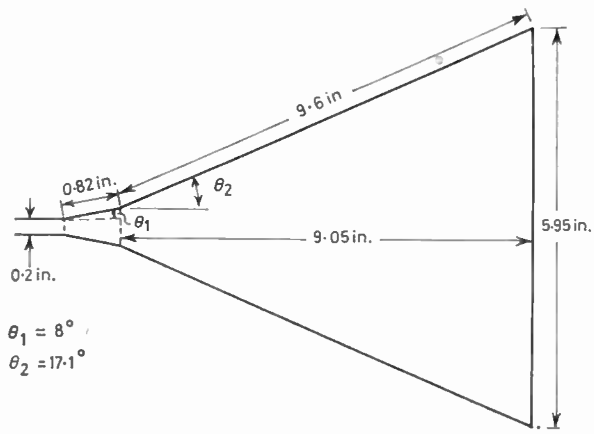

Fig. 13. Flare dimensions.

Another important point is that to the even numbered values of extra phase for feeder 1 predicted by the computer program must be added 180°, i.e.

$$\phi'_{1,2r} = \phi_{1,2r} + \pi \qquad r = 1, 2 \ldots 36$$

This is to introduce a further 180° phase to the $\lambda_g/2$ separation of the cross-arm and hence yield an in-phase distribution over the aperture.

6.4. *The Flare*

The chief purpose of the flare was to provide a reasonable match between the waveguide antenna element and free space. It also provided some vertical directivity. The design is somewhat unusual because the reduced height waveguide cross-arms are only 0·2 in (5·1 mm) and need an abrupt angled flare

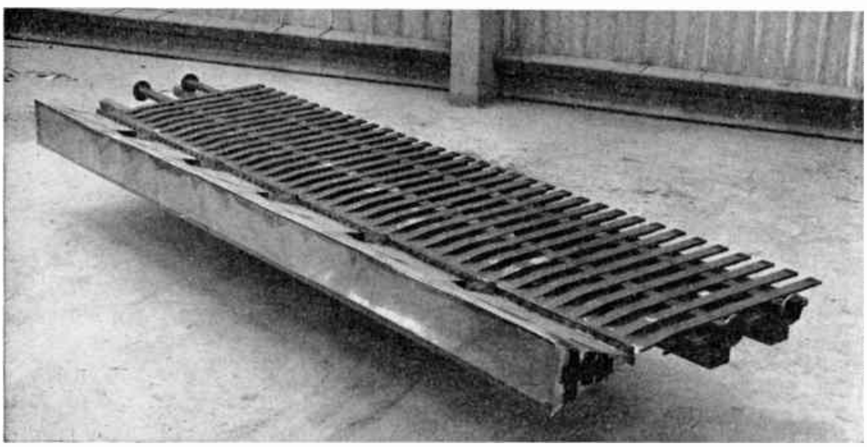

Fig. 14. Experimental model 2 beam Maxson antenna.

to achieve a reasonable vertical aperture (say 6 in or 15 cm) in a short horizontal distance. This abrupt angle would produce a poor match. The flare was thus designed with two flare angles, as shown in Fig. 13. The distance between the two flare sections is $\lambda_g/4$ and the angles are chosen to provide equality of mismatch. The aperture to free-space mismatch is small compared to these two mismatches and is neglected. The prototype flare possessed a v.s.w.r. between 0·7 and 0·9 over the required frequency band.

With the main feeder cross-coupling structure complete, the dielectric pieces were inserted into the cross-arms via the load ends of the cross-arms. When these dielectric inserts were correctly positioned, loads were glued into all cross-arms and the two feeders. The flare was then screwed to the brass angle to complete the antenna. A photograph of the completed antenna is shown in Fig. 14.

6.5. *Bandwidth*

The bandwidth of the Maxson antenna will be limited by such factors as, the design of the directional couplers, the correction of phase shift past the directional couplers, and the use of phase shift inserts to provide the separate beams in place of true line length delays. The complete effect of all such relevant factors has not been analysed, but it would seem that 10% bandwidths are possible.

7. Experimental Results

Due to shortage of time only limited measurements have so far been made on the antenna. The individual patterns produced by feeders 1 and 2 have been measured at centre, low and upper band frequencies. These patterns are shown in Fig. 15. The main beams are well-defined and have the correct beam-width of 1°. The side-lobes are not quite so well defined as might be expected but show a regular periodicity with angle for the side-lobes close to the main beam. The patterns maintain their character over the frequency band but in general the first side-lobes are higher than their design value of −13 dB. The gains of the two beams agreed to within 0·5 dB.

Table 2

Theoretical and measured beam positions as functions of frequency

Frequency MHz	Beam 1 angle Measured	Beam 1 angle Theoretical	Beam 2 angle Measured	Beam 2 angle Theoretical
5300	−6·2°	−6·6°	−5·2°	−5·5°
5600	−0·8°	−1·1°	+0·3°	+0·0°
5900	+3·25°	+3·5°	+4·3°	+4·7°

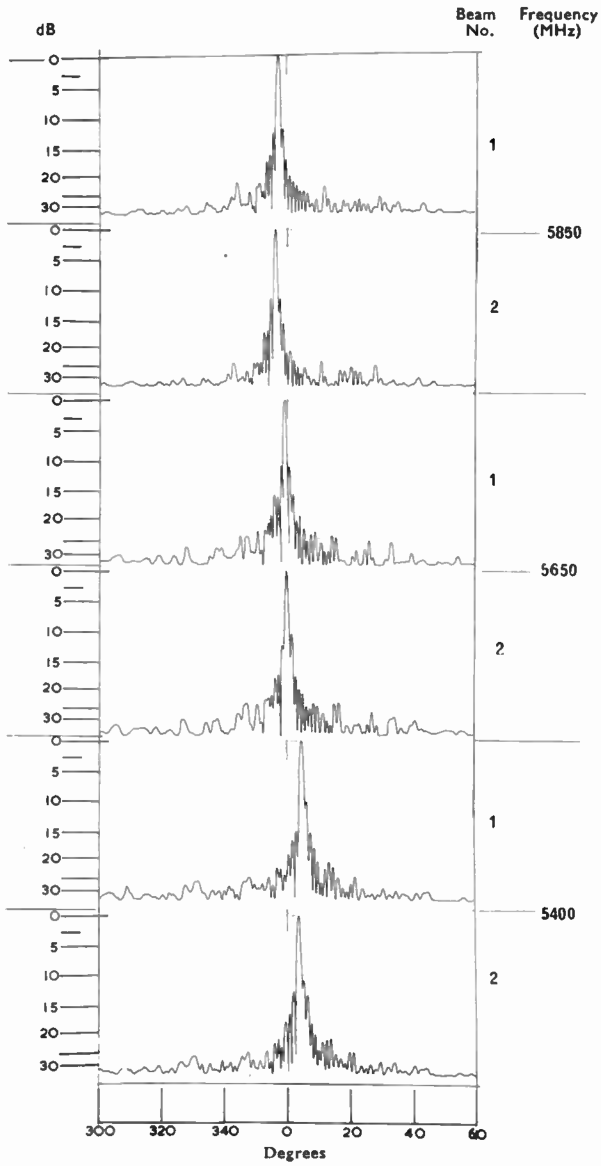

Fig. 15. Two-beam Maxson array, experimental patterns.

Table 2 shows theoretical predictions of beam positions and the measured positions as a function of frequency. Close agreement was obtained indicating that the phase gradients across the aperture were correct and justifying the non-interaction design procedure.

8. Design Compromises in Multiple Beam Design

It was shown in Section 3 that the beams from a loss-less feeder structure must be orthogonal in space. The only beam-shapes discussed so far have been sin u/u beams, which have large −13·2 dB side-lobes. The relationship between crossover levels and side-lobes for orthogonally-spaced beams having tapered aperture distributions is now discussed.

8.1. *Orthogonal Beams*

Using the parameters and definitions given in Appendix 2, a continuous aperture distribution $A(\bar{x})$ produces a far-field pattern $G(u)$ where

$$G(u) = \int_{-1}^{+1} A(\bar{x}) \exp(ju\bar{x})\, d\bar{x} \qquad \ldots\ldots(6)$$

It is also shown in Appendix 2 that:

$$R(u') = \int_{-\infty}^{+\infty} G\left(u+\frac{u'}{2}\right).G^{*}\left(u-\frac{u'}{2}\right) du = $$

$$= \int_{-1}^{+1} A^2(\bar{x}) \exp(ju'\bar{x})\, d\bar{x} \qquad \ldots\ldots(7)$$

When $R(u') = 0$, the beams are orthogonally spaced. It is usually easier to use the indirect relationship involving the aperture power distribution shown above to find orthogonality of two beams. As an example of the use of the above equations, the 'constant plus $\cos^2$' type of aperture distribution has been considered: $A(\bar{x}) = (1-a)+a\cos^2(\pi\bar{x}/2)$ where $0 \leqslant a \leqslant 1$.

For each $A(\bar{x})$ the lowest value of u' which makes $R(u') = 0$ was found. Then using the pattern $G(u)$ corresponding to the $A(\bar{x})$ considered, the crossover level is given by $G(u'/2)/G(0)$. Also the maximum side-lobe level of $G(u)$ was found. The calculations were performed on a digital computer and the results are given in Fig. 16. This is a graph of crossover level against side-lobe level for orthogonal spacing of beams produced by 'constant+$\cos^2$' aperture distributions. It can be seen that as the side-lobe levels are reduced the crossover level falls. Crossover levels of less than say -6 dB could not be tolerated in a multiple beam surveillance system or 'troughs' in the radar coverage would appear. This crossover level of -6 dB corresponds to a side-lobe level of -19 dB. To achieve high crossover levels and low side-lobes it is necessary to use non-orthogonal spacing of the beams.

8.2. *Non-orthogonal Beams*

When the beams are not spaced for orthogonality, the feeder structure cannot be lossless and a reduced feeder efficiency is obtained. Feeder efficiency $|\eta|^2$ of a beam is defined as the ratio of power radiated to the power exciting the feeder input of that beam. The Maxson antenna is well suited to the design of non-orthogonal beams as the amount of loss and where it occurs in the feeder structure can be predicted. Referring to eqn. (4) of Section 3, it is seen that the summation of the cross-coupled aperture fields is equal in magnitude but opposite in sign to the cross-coupled fields in the feeder loads. Expressed in terms

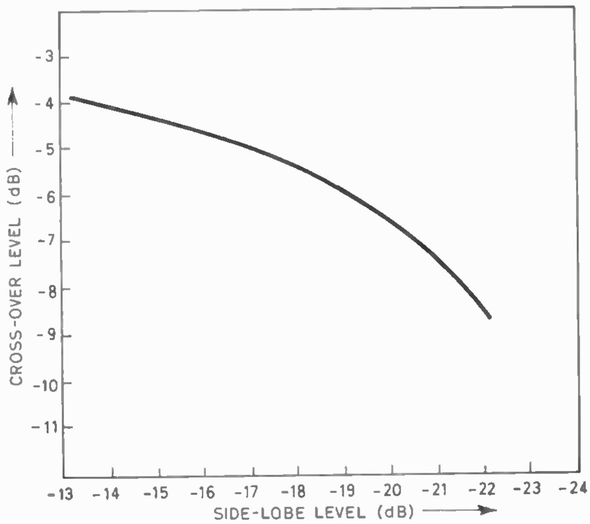

Fig. 16. Constant + $\cos^2$ distribution (orthogonal spacing).

of the continuous distributions used in this Section:

$$\int_{-1}^{+1} A^2(\bar{x}) \cos(u'\bar{x})\, d\bar{x} = -|p_{L1}|\cdot|q_{L1}| \cos\theta_{L1} \qquad \ldots\ldots(8)$$

It is convenient to normalize

$$R(u') = \int_{-1}^{+1} A^2(\bar{x}) \cos(u'\bar{x})\, d\bar{x}$$

so that $R(0) = 1$.

Then for beams having equal efficiency factors, η^2, eqn. (8) becomes

$$\eta^2 R(u') = -|p_{L1}|\cdot|q_{L1}| \cos\theta_{L1}$$

The largest value that the cross-coupled load fields can take is when

$$q_{L2} = 0$$

and

$$\theta_{L1} = \pi$$

so that

$$-|p_{L1}|\cdot|q_{L1}| \cos\theta_{L1} = 1-\eta^2$$

Therefore,

$$\eta^2 R(u') \leqslant 1-\eta^2$$

$$\eta^2 \leqslant 1/\{1+R(u')\} \qquad \ldots\ldots(9)$$

A general result for multiple beam systems similar to eqn. (9) has been derived by Stein[10] using a matrix theory approach.

Equations (6), (7), (8) and (9) can be used to design a non-orthogonal multi-beam system. Again the 'constant+$\cos^2$' distribution was considered. Since high crossover levels are desirable it was decided to maintain the crossover level at $-3{\cdot}92$ dB for all the beam shapes. Stipulating the crossover level gives the value of u', the angular spacing of the beams. Then the value of $R(u')$ can be found. Inserting this value in eqn. (9) yields the maximum efficiency in a feeder design for such beam-shapes and crossover level. Figure 17 shows a plot of maximum efficiency against side-lobe level for the 'constant + $\cos^2$' type of aperture distribution when the crossover level is maintained at $-3{\cdot}92$ dB. The feeder efficiency falls as the side-lobe level is reduced but even for a side-lobe level of -40 dB an efficiency of 74% could be achieved.

Equation (9) has been verified using the coupler design program on the computer. The spacing between the two sin u/u beams was narrowed and the value of $R(u')$ for this new spacing evaluated. Equation (9) then predicted the maximum feeder efficiency, η^2, attainable. The coupler program was then run with load powers less than and greater than $1-\eta^2$. When a design with a greater efficiency than η^2 was attempted the program failed as coupling values greater than 0 dB were produced, i.e. gain was required in the feeder matrix. When a design with an efficiency 0·1% less than maximum η^2 was attempted the program succeeded but very large coupling values were required in the second feeder arm. To achieve reasonably practical coupling values an efficiency of about 4% less than maximum was required. It could thus be expected that a practical non-orthogonal design having very low side-lobes could be made with 70% feeder efficiency.

Whilst this Section has shown what trade-offs there are between side-lobe, crossover and efficiency for individual beams, there is another alternative. When sin u/u beams spaced for orthogonality are combined in the correct ratios they can realize 'cosine' or 'constant + $\cos^2$' distribution (Appendix 3). It is thus possible to achieve a low side-lobe composite pattern by performing parallel operations on the feeder inputs. The choice is then between parallel operations on the feeder inputs using orthogonal sin u/u beams, or the relatively simpler sequential operations on the feeder inputs using non-orthogonal beams which have a reduced feeder efficiency. The practical difficulty in implementing all but the simplest parallel feeder schemes is such that the more involved design of the non-orthogonal beams is probably worthwhile.

9. Applications

As a multi-beam antenna it is felt that unless a better matched low-loss phase insert can be designed only a moderate number of beams can be produced. Otherwise the mismatch in the cross-arms from the remotest feeders will build up to an impractical level.

A suitable application of this antenna would be as an evaluation feeder for a stack of linear arrays. Switching inputs on the feeder would give elevation scan and azimuth scan could be performed by conventional rotation. If simple combinations of the feeder inputs were performed then monopulse operation could be achieved between the vertical beams, allowing height finding to be facilitated.

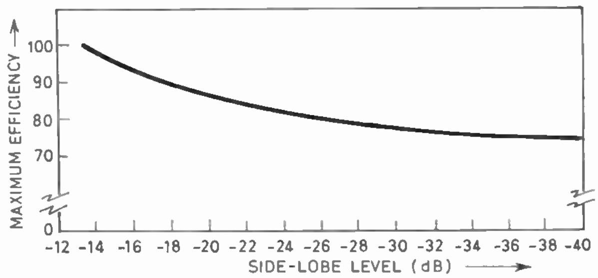

Fig. 17. Constant + $\cos^2$ distribution, crossover maintained at $-3{\cdot}92$ dB.

The structure could be used as a surveillance radar with multiple fan beams but unless only a limited angular sector needed searching, the advantage of multiple beams would not be great.

The two beam structure is a method of providing fan-beam monopulse but the applications of this beam configuration appear limited. In its favour is that the sum and difference patterns could be carefully controlled so that an optimum monopulse system could be designed. A more worthwhile arrangement would be to form a vertical stack of these monopulse feeders to produce a planar array having true monopulse null. Again the Maxson feeder structure could be used to feed the vertical stack. An advantage of this planar assembly over its optical conterpart is that the depth of the antenna could be smaller and the beam patterns could be specified quite accurately.

10. Conclusions

The constraints on the design of multiple beams from a Maxson antenna have been presented. A design procedure for calculating the coupling coefficients and phase inserts required to produce non-interacting beams from the feeder matrix has been described. This design procedure allowed for the phase-shifts associated with the directional couplers and the values of the phase inserts are the extra phase shift required to produce the linear phase fronts of the beams.

Although only limited measurements have been performed on the Maxson antenna that was built, they indicate the correctness of the above design approach. This antenna was designed to produce two orthogonally-spaced sin u/u beams.

The design constraints on orthogonal and non-orthogonal beams have been theoretically examined. Orthogonal beams cannot give a high crossover level and low side-lobes. However, non-orthogonal beams can realize these two objectives when the feeder efficiency, η^2 is less than 100%. A crossover level of

$-3{\cdot}92$ and side-lobe level of -40 dB could be achieved in a practical antenna with a feeder efficiency of 70%.

When sequential processing of the feeder inputs is required the non-orthogonal design is required. If combination processing of various inputs is performed then low side-lobes can be achieved using orthogonally spaced sin u/u beams.

Several applications of the Maxson antenna have been outlined but of these its use as a vertical feeder for a planar assembly of linear arrays and the production of a Maxson–Maxson-fed planar array structure for true monopulse appear to be the most useful.

11. Acknowledgments

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13. Appendix 1

Phase Relationships in a Directional Coupler

Consider the perfectly-matched directional coupler shown in Fig. 18(a). The voltages are the normalized voltages such that $\frac{1}{2}|V|^2$ represents the power at that point in the guide. The planes at which $\bar{V}_1$, $\bar{V}_2$, $\bar{V}_3$ are considered are far enough from the coupling holes, that only a pure H_{01} mode exists.

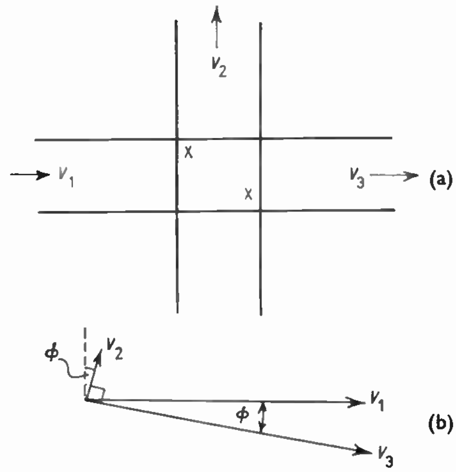

Fig. 18. Phase relationships in a directional coupler.

Now from the principle of conservation of energy, the following relation can be written.

$$|V_1|^2 = |V_2|^2 + |V_3|^2$$

Also the voltages $\bar{V}_2$, $\bar{V}_3$ can be considered to have their origin at a junction where $\bar{V}_1 = \bar{V}_2 + \bar{V}_3$.

The two relationships can only be satisfied when $\bar{V}_1$, $\bar{V}_2$, $\bar{V}_3$ comprise a right-angled phase diagram as in Fig. 18(b).

The coupled power is given by $\frac{1}{2}|V_2|^2$.

Therefore, coupling factor,

$$C = \frac{|V_2|^2}{|V_1|^2}$$

From Fig. 18(b),

$$\sin\phi = \frac{|V_2|}{|V_1|} = \sqrt{C}$$

The alternative phase diagram with V_3 phase advanced by the angle, ϕ_1 is discounted on experimental grounds.

14. Appendix 2

Aperture Power Distribution and Beam Pattern Correlation

From Figs. 19 (a) and (b) the following relations can be obtained:

$$G(\theta) = \int_{-l/2}^{+l/2} \mathrm{A}(x)\exp\left(\mathrm{j}\frac{2\pi}{\lambda}\, x \sin\theta\right)\mathrm{d}x$$

let

$$\bar{x} = 2x/l$$

$$u = \frac{\pi l}{\lambda}\sin\theta$$

$$G(u) = \int_{-1}^{+1} \mathrm{A}(\bar{x})\exp(\mathrm{j}u\bar{x})\,\mathrm{d}\bar{x}$$

Define

$$R(u') = \int_{-\infty}^{+\infty} G\left(u+\frac{u'}{2}\right).G^*\left(u-\frac{u'}{2}\right)\mathrm{d}u$$

$$G\left(u+\frac{u'}{2}\right) = \int_{-1}^{+1} \mathrm{A}(\bar{x})\exp\left\{\mathrm{j}\left(u+\frac{u'}{2}\right)\bar{x}\right\}\mathrm{d}\bar{x}$$

$$G^*\left(u-\frac{u'}{2}\right) = \int_{-1}^{+1} \mathrm{A}^*(\bar{y})\exp\left\{-\mathrm{j}\left(u-\frac{u'}{2}\right)\bar{y}\right\}\mathrm{d}\bar{y}$$

Therefore,

$$R(u') = \int_{-\infty}^{+\infty}\left[\int_{-1}^{+1} \mathrm{A}(\bar{x})\exp\left\{\mathrm{j}\left(u+\frac{u'}{2}\right)\bar{x}\right\}\mathrm{d}\bar{x}\times\right.$$

$$\left.\times\int_{-1}^{+1} \mathrm{A}^*(\bar{y})\exp\left\{-\mathrm{j}\left(u-\frac{u'}{2}\right)\bar{y}\right\}\mathrm{d}\bar{y}\right]\mathrm{d}u$$

$$= \int_{-1}^{+1} \mathrm{A}(\bar{x})\exp\left(\mathrm{j}\frac{u'}{2}\bar{x}\right)\mathrm{d}\bar{x}\,.\,\times$$

$$\times\int_{-1}^{+1} \mathrm{A}^*(\bar{y})\exp\left(\mathrm{j}\frac{u'}{2}\bar{y}\right)\mathrm{d}\bar{y}\,.\int_{-\infty}^{+\infty}\exp\{\mathrm{j}u(\bar{x}-\bar{y})\}\,\mathrm{d}u$$

$$= \int_{-1}^{+1} \mathrm{A}(\bar{x})\exp\left(\mathrm{j}\frac{u'}{2}\bar{x}\right)\mathrm{d}\bar{x}\times$$

$$\times\int_{-1}^{+1} \mathrm{A}^*(\bar{y})\exp\left(\mathrm{j}\frac{u'}{2}\bar{y}\right)\mathrm{d}\bar{y}\cdot\delta(\bar{x}-\bar{y})$$

which, from the properties of the delta function, yields

$$R(u') = \int_{-1}^{+1} \mathrm{A}^2(\bar{x})\exp(\mathrm{j}u'\bar{x})\,\mathrm{d}\bar{x}$$

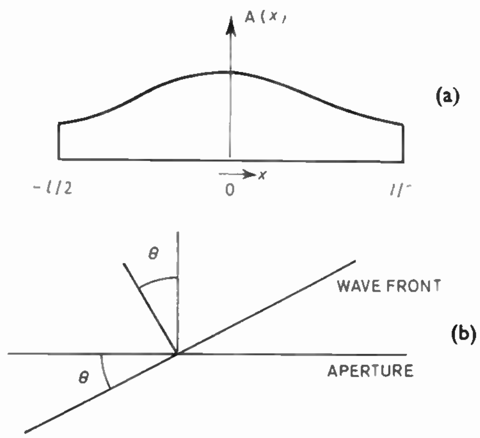

Fig. 19

15. Appendix 3
Aperture Distributions and Beam Patterns

15.1. *Cosine Distribution*

$$\mathrm{A}(\bar{x}) = \cos\frac{\pi}{2}\bar{x}$$

$$= \left\{\exp\left(\mathrm{j}\frac{\pi}{2}\bar{x}\right)+\exp\left(-\mathrm{j}\frac{\pi}{2}\bar{x}\right)\right\}\Big/2$$

$$G(u) = \int_{-1}^{+1} \mathrm{A}(\bar{x})\exp(\mathrm{j}u\bar{x})\,\mathrm{d}\bar{x}$$

$$= \tfrac{1}{2}\int_{-1}^{+1}\left\{\exp\left(\mathrm{j}\frac{\pi}{2}\bar{x}\right)+\exp\left(-\mathrm{j}\frac{\pi}{2}\bar{x}\right)\right\}\exp(\mathrm{j}u\bar{x})\,\mathrm{d}\bar{x}$$

$$= \tfrac{1}{2}\left[\frac{\exp\left\{\mathrm{j}\left(u+\frac{\pi}{2}\right)\bar{x}\right\}}{\mathrm{j}\left(u+\frac{\pi}{2}\right)}\right]_{-1}^{+1} +$$

$$+\tfrac{1}{2}\left[\frac{\exp\left\{\mathrm{j}\left(u-\frac{\pi}{2}\right)\bar{x}\right\}}{\mathrm{j}\left(u-\frac{u}{2}\right)}\right]_{-1}^{+1}$$

$$= \frac{\sin\left(u+\frac{\pi}{2}\right)}{u+\frac{\pi}{2}}+\frac{\sin\left(u-\frac{\pi}{2}\right)}{u-\frac{\pi}{2}}$$

15.2. *Constant* $+\cos^2$ *Distribution*

$$\mathrm{A}(\bar{x}) = (1-a)+a\cos^2(\pi\bar{x}/2)$$

$$= 1-a+\frac{a}{2}\{1+\cos(\pi\bar{x})\}$$

$$= \left(1-\frac{a}{2}\right)+\frac{a}{4}\{\exp(\mathrm{j}\pi\bar{x})+\exp(-\mathrm{j}\pi\bar{x})\}$$

$$G(u) = \int_{-1}^{+1} \mathrm{A}(\bar{x})\exp(\mathrm{j}u\bar{x})\,\mathrm{d}\bar{x}$$

$$= \int_{-1}^{+1}\left[\left(1-\frac{a}{2}\right)+\frac{a}{4}\{\exp(\mathrm{j}\pi\bar{x})+\right.$$

$$\left.+\exp(-\mathrm{j}\pi\bar{x})\}\right]\exp(\mathrm{j}u\bar{x})\,\mathrm{d}\bar{x}$$

$$= \left(1-\frac{a}{2}\right)\frac{\sin u}{u}+\frac{a}{2}\left\{\frac{\sin(u+\pi)}{u+\pi}+\frac{\sin(u-\pi)}{u-\pi}\right\}$$

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The Nucleonic Instrumentation Conference and Exhibition

READING UNIVERSITY

23rd to 25th September 1968

The Conference was organized by the Institution of Electrical Engineers, in collaboration with the British Nuclear Energy Society, the Institution of Electronic and Radio Engineers, the Institute of Physics and The Physical Society, and the Scientific Instrument Manufacturers' Association. The I.E.R.E. was represented on the Conference Organizing Committee by Messrs. R. J. Cox and F. E. Whiteway. Mr. Cox, who has recently been appointed Chairman of the Institution's Programme and Papers Committee, also represented the British Nuclear Energy Society and this report by him has appeared in the January issue of the Society's Journal.

Several conferences each year take place in various parts of the world on Nuclear Electronics, but very rarely do they seem to be held in Britain. The conference held at Reading University in September was therefore particularly welcome. It was also one of the first conferences organized by learned societies to be run in collaboration with a commercial exhibition. This combination is frequent in other countries, particularly the U.S.A., but is quite new over here. It proved to be popular as most of the devices discussed at the Conference were on display—and working—on the Exhibition stands. (It was also successful in the sense that the organizers appeared to have almost suppressed the blatant sales lecture.) The Exhibition clearly helped towards keeping the attention of the Conference focused on what could be achieved on a production basis and away from advanced ideas that may never reach fruition.

The papers tended to present a mosaic of the field of Nuclear Instrumentation rather than covering the whole field uniformly. A few papers are remembered in vivid colours but many just presented a grey background.

The Introductory Address by Dr. S. C. Curran presented a nostalgic but humorous review of nucleonic electronics from the days in 1935 when it was a struggle to make a Geiger counter that worked.

The Industrial session consisted of three papers, all notable for the almost incredible nature of their technique. The first, by W. J. Hardwick and B. J. Laundy, described a β-ray backscatter thickness gauge for monitoring the sedimentation of fine particles (0·1 to 2 µm) in a liquid in a centrifuge and thus measuring their size. The source rotates with the centrifuge tube but the cylindrical plastic scintillator and photo-multiplier remain stationary outside its path of rotation. It is amazing that the scintillations on the remote side of the cylinder can reach the photomultiplier without suffering unacceptable attenuation. The authors presented experimental data that showed the effect to be quite small. The second paper, by R. L. Otlet, on 'Low-level tritium measurements for hydrological applications' depends on the enormous increase of tritium level in rain since hydrogen bomb tests started, enabling pre-1954 water to be distinguished from 'new' water. The third paper, by D. Salt, described equipment used by Rolls-Royce for tracking down wear troubles in bearings of aircraft engines. A small source imbedded in the bearing cage is viewed by a scintillation counter outside the engine. As the cage rotates it produces a cyclic variation in the counting-rate which can be used as a tachometer. Comparison with the conventional shaft tachometer gives a measure of the bearing cage 'slip' relative to the shaft. This is a technique that obviously required a lot of development effort and expensive hardware, but it is difficult to conceive any other method that would give the same results and clearly the need justifies the expense.

The first paper of the next session—'Reactor Instrumentation'—was by R. Dehn and presented a survey of methods of measuring temperature in reactors. It concentrated more on the problems of irradiation experiments, particularly post-mortem techniques, rather than on-line methods for power reactors. One very interesting method involves using what the author calls a 'kryptonate', which is described as a solid into which Kr^{85} has been implanted (bombarded under high-energy, not just soaked in the gas). Apparently when such a material is heated, a fraction of the Kr^{85} is released, the fraction being a function of temperature. A subsequent re-heating does not release any more krypton until the original temperature is passed. In a power reactor the fuel cladding is automatically impregnated with Kr^{85} by the fission process. At Dounreay they are using the technique to establish the temperature a sample of fuel cladding was experiencing when it was in the reactor. This is an intriguing method as it involves no special provisions in-pile and merely involves heating a sample of cladding metal in the laboratory and observing the temperature at which the krypton starts to be released.

The next three papers, by authors from A.E.E., Winfrith, The Plessey Company, Twentieth Century Electronics and System Computers, described the high temperature (550°C) radiation detectors and their associated electronic equipment that have been designed to be used in-pile for the control of the AGR and PFR types of power reactor, although they clearly have much wider applications. The design of the familiar pulse-counting fission chamber instrument has now reached the stage where the fission chamber can be left in the reactor in a flux of about 10^{11} neutrons/cm² at full-power for prolonged period and still be capable of counting low fluxes immediately after shut-down. This clearly enables the expensive chamber retraction machinery to be eliminated. All the radiation detectors described are available with outer cases which are an essential part of a design technique that the authors claim allows electrical interference to be reduced to a very low level. The logarithmic d.c. amplifier and periodmeter measures from a few times 10^{-11} A up to a few milliamperes which is the level that the chambers produce at full power at the in-pile sites used.

A paper by E. P. Fowler and A. Metcalf describes a new method of measuring neutron flux in reactors. The authors call it a 'Campbell system' because of its reliance on a theorem of Campbell published in 1909 related to statistical processes but it is also known as 'statistical fluctuation analysis', and sometimes, erroneously, as 'noise analysis'. Instead of measuring the mean current from an ionization chamber or counting individual neutron pulses, the technique depends upon the measurement of the statistical fluctuations of the current (usually the r.m.s. value). This produces an electrical signal proportional to the square-root of the neutron flux, so that a $10^8 : 1$ range of flux produces a $10^4 : 1$ range of signal, thereby easing the electronic problems. More importantly, the technique weights the signal in favour of the larger pulses, so that there is a substantial improvement in neutron/gamma ratio and reduction of the effects of cable leakage—both are important for high temperature in-pile applications. It is clear that the technique has more parameters to be specified than the familiar mean-current method, so that optimizing the design for a particular application is more complicated. The authors described a most ingenious circuit that provides the r.m.s. value of the signal on a logarithmic scale.

Three authors from Elliott Process Automation described the irradiated ball flux scanning system that has been developed for Dungeness 'B'. A column of stainless steel balls about $\frac{1}{16}$ inch in diameter is blown into a tube in the reactor and irradiated. After a suitable time the balls are blown out and the activity of small batches of balls successively measured in an ionization chamber. The method is analogous to the irradiated wire flux scanning method but obviously is a lot easier to use. The paper and the working exhibit in the exhibition displayed the thoroughness with which the practical problems have been tackled and solved.

A paper by G. Knill of Harwell gave an up-to-date account of the current state of development of the primary emission-neutron-activation of 'Hilborn' detectors used for flux-scanning purposes. The paper does not evaluate the usefulness of these devices under the high temperature conditions that might be met in power reactors but it does give useful new information on their burn-up characteristics and on the effect of the resonance in the rhodium neutron cross-section.

E. P. Fowler presented a review of the current 'state of the art' in low-level d.c. amplifiers, concentrating on the use of the junction field effect transistor (f.e.t.), the insulated gate field effect transistor (the m.o.s.t.) and the transistor as a logarithmic element. It is clear that these semiconductor devices have improved enormously during the past few years, in fact to the stage where they are now superior to electrometer valves for virtually every application.

The last of the nine papers on reactor instrumentation was by R. E. Whyard, J. D. Evans and R. M. Ellis and it described a method of using the m.o.s.t. type of integrated circuit for reactor safety circuits. A large number of logic elements can be produced on a tiny chip of silicon (about 1·8 mm square) so that quite complicated systems can be built from a comparatively small number of packages. By using a form of 'pulse coding' logic, the probability of system failure can be made extremely small. The method is clearly very attractive when it can be linked to a computer to initiate the automatic checking routines and for providing the necessary information to the operator.

Most of the next session was spent on discussing a batch of papers describing the use of small computers in conjunction with research experiments at a number of the Atomic Energy Authority establishments. When these tiny computers are used for a single straightforward experiment they are easy to program and are of great value. But as soon as 'double-word length working' is needed or 'time-sharing' is required to deal with a number of experiments simultaneously, then the amount of programming effort increases enormously and so does the complexity of the interface hardware. One suspects that in many cases a lot of money and time could have been saved by using a more comprehensive computer in the first place. The difficulties of using a FORTRAN or other high-level language compiler means that the standard of programming needs to be of a very high level, adding further to the software costs.

A paper by H. Bisby of Harwell on a system of mechanical and electrical standards for computer interface hardware, called CAMAC, produced the most heated discussions of the Conference. There is already a considerable degree of international agreement to these standards from the U.S.A. and European countries in the nuclear physics field. But there was a wide degree of confusion between the standards themselves and Harwell's physical interpretations of them (the 7000 series). Clearly there is merit in these standardization proposals and at first sight they would appear to be suitable for industrial process control applications as well as the present field of nuclear physics research.

In the session on 'Physics Research' there were three papers on semiconductor radiation detectors. The first described work at Aldermaston and Los Alamos on the use of surface barrier silicon diodes in the current mode for measurements during the flash of an atomic weapon explosion. The other two papers from Harwell and Nuclear Enterprises Ltd. dealt with lithium-drifted germanium detectors which are now so enormously important for γ-spectroscopy. They dealt particularly with the problem of making large volume (around 50 cm^3) detectors and of finding out whether a particular batch of germanium will produce good detectors or not before they are actually made.

M. C. B. Russell and D. B. J. Smith described an ingenious digital technique for controlling the speed and phase of the high-speed rotors of a neutron time-of-flight spectrometer. The complicated nature of the technique is understandable when one realizes that it achieves a speed accuracy of the almost incredible figure of about 1 part in 10^9.

The final session of the Conference was devoted to Health Physics and Medical Applications. A paper by T. E. Burlin and B. W. Hancock dealt with the problem of assessing the dose to a layer of soft tissue close to the surface of a bone when irradiated with γ-rays. They constructed an ionization chamber with one electrode of graphite which is an adequate simulation of soft tissue and the other electrode of aluminium which simulates bone. By operating the chamber at a potential less than 30 V, the current measured when irradiated by γ-rays is a measure of the low-energy electrons emitted from the electrodes.

Solid-state radiation detectors have had an enormous impact on various areas of nucleonics—particularly γ-spectroscopy. Although the lithium-drifted germanium detectors operated at liquid nitrogen temperatures have applications in the medical laboratory, they have obvious practical difficulties for experiments *in vivo*. A paper by W. Abson, F. L. Allsworth and P. G. Salmon, of Harwell, gave an excellent review of the current state of development of solid-state detectors for medical applications. All the detectors described are of silicon and most are intended to be used in the pulse counting mode but a few are used for dose or dose-rate measurements by employing direct-current amplifier techniques. The pulse-counting mode requires a pulse amplifier near the counter and the size of present designs is obviously an embarrassment. Here is an application that surely demands the use of an integrated circuit that can be built into the detector.

Two papers from Harwell dealt with proportional counters used for difficult health physics requirements. One of them, by B. T. Taylor, described a method of measuring the Pu^{239} content of a human lung by detecting from outside the chest the X-rays in the 10–20 keV region emitted following 4% of the α disintegrations of Pu^{239}. Not only are the pulses small in size but they are very few in number. The counter is large (about $35 \times 35 \times 12{\cdot}5$ cm) containing three main counting volumes but it also has built-in anti-coincidence counters, which, together with a pulse shape and pulse amplitude selection technique, enables the background counting-rate to be reduced to acceptable levels. The other paper, by H. J. Delafield *et al.*, described a proportional counter for measuring biological dose for neutrons from about 0·025 eV to 17 MeV. It is intended to be used held close to a human body so that neutrons of energies below 10 keV are reflected by the body and detected by the 580 keV protons generated by the $^{14}N\ (n,p)^{14}\ C$ reaction in the walls of the counter. Pulses are again small (about 2×10^4 ion pairs) and the sensitivity is only about 10 counts per millirem.

R. J. COX

The papers are published in a 240-*page book*, 'Nucleonic Instrumentation, I.E.E. Conference Publication No. 47', *and copies are available from the I.E.E., price* £2 5*s. for members of the sponsoring bodies of the Conference, and* £3 10*s. for non-members.*

Conference on Industrial Ultrasonics

Ultrasonics is now an established technique in nearly every industry and this progress has been only made possible by the use of electronic techniques. It is therefore appropriate that the Institution of Electronic and Radio Engineers be joined by the Institution of Electrical Engineers, the Institute of Physics and the Physical Society, the Institute of Electrical and Electronics Engineers (United Kingdom and Republic of Ireland Section), the British Acoustical Society and the Non-Destructive Testing Society of Great Britain, in organizing a three-day conference. This will be held at Loughborough University of Technology from Tuesday, 23rd to Thursday, 25th September, 1969.

The scope of the Conference will be concerned with industrial and research applications of ultrasonics. Papers will review existing techniques and the latest developments or describe research and development which will lead to future applications. One session will be devoted to applications of ultrasonics to the radio and electronics industry, such as welding of microcircuits, cleaning of components, manufacture of plastic components using welding techniques, ultrasonics in machining, etc.

Generally, papers should describe novel developments or significant improvements in design in any of the following fields:

Flaw detection and thickness gauging; Obstacle detection; Position measurement; Leak detection; High power applications; Measurement of physical properties, e.g. temperature, absorption, viscosity, elasticity; Telemetry.

Specialized applications in the fields of sonar and medicine will be excluded as these have been the subject of recent conferences.

The Organizing Committee, which is under the Chairmanship of Professor J. W. R. Griffiths and includes representatives of the sponsoring bodies drawn from Universities, Government Research Establishments and Industry, invites synopses of proposed contributions (about 200 words in length) which should be sent as soon as possible, preferably not later than 30th April 1969 to: The Secretary, Organizing Committee for the Conference on Industrial Ultrasonics, The Institution of Electronic and Radio Engineers, 9 Bedford Square, London, W.C.1.

As well as full length papers of up to 6000 words, the Committee will welcome shorter contributions in the region of 1500 to 2000 words in length. Papers in final form will be required by 30th June 1969.

Further information and registration forms for the Conference will be available in due course from the Conference Registrar at the I.E.R.E.