Migration of Engineers

R EPORTS from Asia, Europe, and North America call into question the migration of scientists and engineers from the countries of their birth or training. This controversial issue arises in an era when new means of communication bring the world population closer together and there is unprecedented expenditure on improving the standard of living of backward nations.

Eliminating areas where migration is virtually forbidden, the movement of engineers is mainly confined to Western Europe, Northern America and most of Asia. For well over one hundred years temporary migration from the less developed countries of Asia and South America to the more highly developed countries has been encouraged. The resultant technical education and training of the migrant has, on his repatriation, been of benefit to his country.

The present problem of migration, however, is the increasing drift of trained scientists and engineers toward countries already rich, in the technological sense, and able to offer the best opportunities for technological innovation. The radio and electronic engineer is especially involved in the problem of migration because of the increasing international demand for his skill.

Moral obligation to country of birth or training is a personal matter, and short of arbitrary legislation, depends on patriotic persuasion. The more realistic fact, however, is that personal reward in the shape of opportunity for achievement together with a high standard of living are natural human desires. In the past, Great Britain and the Commonwealth have afforded such opportunities and have reaped great technical benefit therefrom, but in present circumstances migration is an economic problem in terms of realizing investment in education, training and research.

Moreover, the cost of research and development is now so great that no single country can hope to cover all future possibilities in, for example, radio and electronics. Prosperity will, therefore, depend on the technical foresight of both government and industry in determining those areas of research and development which can most profitably be exploited in manufacturing processes. A clear example is shown by the British post-war record. In spite of spending far more on research and development than any other country in Europe, Britain has had a slow growth rate in terms of total production: perhaps because of too widely diffused efforts.

Underlying these problems is the urgent need to secure more recruits to the engineering profession. The world shortage of engineers, particularly in the electronics field, demands urgent attention, for without the engineers who make the benefits of modern technology available, it is impossible even to maintain, let alone improve, the standard of living.

G. D. C.

NOTICES

France Honours British Engineers

The 'Société d'Encouragement pour la Recherche et l'Invention' honoured five members of the I.E.R.E. by the presentation of its 'Ordre du Mérite pour la Recherche et l'Invention', at a meeting held in Paris on 7th June.

This was the first time that this Society, which was founded in 1955, had awarded medals to engineers from outside France 'who have attained particular distinction by their activities or by the efficient contribution they have made to the development of research and invention'.

Details of the Awards, and the citations, are as follows:

Grande Médaille d'Or to Admiral of the Fleet the Earl Mountbatten of Burma, K.G., O.M., F.R.S. (Charter President of the I.E.R.E.), Chairman, National Electronics Council, in recognition of his support for research.

Médaille d'Or to Leslie H. Bedford, C.B.E. (Past President of the I.E.R.E.), Director of Engineering, Guided Weapons Division, B.A.C. (Operating) Ltd., in recognition of his original researches and direction of work in Television, Radar and Guided Weapons.

Médaille de Vermeil to Graham D. Clifford, C.M.G., Secretary of the I.E.R.E., in recognition of his work nationally and internationally in education and international co-operation.

Médaille d'Argent to Wing Commander Gerald E. Trevains, R.A.F., a member of the Committee of the Institution's French Section, in recognition of his work at the British Embassy in Paris in fostering British/French co-operation, and in liaison between Learned Societies.

Médaille de Bronze to Paul J. C. Prevost, Secretary of the Committee of the Institution's French Section, in recognition of his contributions to radar during the War and to the development of the *Europa* Satellite Launcher.

Formal Winding-up of the Brit. I.R.E.

The London Gazette dated 9th March 1967 gave formal notice of the final meeting of the British Institution of Radio Engineers, which was held at 8 Bedford Square, London, W.C.1, on Monday, 10th April 1967 at 11 a.m.

At the meeting the Institution's Auditor, Mr. R. H. Jenkins, Chartered Accountant, who was appointed liquidator for the winding-up at an Extraordinary General Meeting of Corporate Members of the British Institution of Radio Engineers, held on 13th December 1965 (see *Proceedings of the I.E.R.E.*, January–February 1966) reported on his actions. He stated that, in accordance with the Special Resolution passed at the E.G.M. and Clause 22 of the Royal Charter of Incorporation of the I.E.R.E., the property and moneys had been formally transferred to the Institution of Electronic and Radio Engineers.

Correction

The following correction should be made in the paper 'Low frequency sound sources: statement of problem and some possible solutions', which was published in the April 1967 issue of *The Radio and Electronic Engineer*:

Page 267, Fig. 6: caption should read:

. . . (after Clay, Liang and Wisotsky[16]).

Conference on

'The Integration of Design and Production in the Electronics Industry'

The programme arrangements for this Joint Conference to be held at the University of Nottingham from 10th to 13th July are now completed and details of the papers, including synopses and the times of presentation, are given on the following pages of this issue. The majority of the papers will be available in preprint form at the Conference; the complete Conference Proceedings will be published during the Autumn of 1967, price £6 and orders may be placed now.

Applications for registration forms for this Conference can still be made, using the form at the end of this issue or by telephone (01-580-8443, ext. 11). The charge for registration, which includes the pre-

print volume and lunch and refreshments on the three full days, is £10 for members of the sponsoring Institutions and £13 for non-members. Accommodation in a Hall of Residence is available at a charge of £10 for the nights of Monday to Wednesday, and includes the Dinner on Monday evening marking the official opening of the Conference. The charges for accommodation for shorter or longer periods than the duration of the Conference may be obtained on request. Members resident in the East Midland Section and adjacent areas may obtain tickets for this Dinner on application to the Institution; the charge is £2 12s 6d per ticket.

Joint I.E.R.E. - I.Prod.E. - I.E.E. Conference on

'The Integration of
Design and Production in the Electronics Industry'

at the University of Nottingham, 10th to 13th July, 1967

PROGRAMME AND SYNOPSES OF PAPERS

Monday Evening, 10th July, 7 for 7.30 p.m.

   Formal Opening of the Conference and Conference Dinner in the Portland Building

Tuesday, 11th July   SESSION I: DESIGN FOR PRODUCTION

9.30–10.30 a.m. *Chairman:* J. A. SARGROVE, C.Eng., M.I.Mech.E., F.I.E.E., M.I.E.R.E., M.I.Prod.E.

Looking Ahead with Hindsight

E. EASTWOOD, C.B.E., Ph.D., M.Sc., C.Eng., F.I.E.E. (*English Electric Company Ltd., London.*)

   Progress in electronics since the end of World War II has been spectacular enough but its performance might have been even more impressive if a larger proportion of all the prophecies for new devices, techniques, materials and functions which have been made during the last twenty years had indeed come to pass.

   The paper touches upon a number of these hoped-for developments and seeks to understand why their promise has not been fulfilled. Lessons learnt in this way can then be applied to present-day forecasts of the probable path of development of the electronics industry and so try to temper optimism with discretion.

The Relationship between Design and the Market in the Electronics Industry

J. SHARPE, B.Sc., C.Eng., F.I.E.E. (*E.M.I. Electronics Ltd., Hayes.*)

   There are no profit-making activities within a business. Profit comes from the interaction of design and production with the market. The problem facing designers is, therefore, to create devices which are needed by the market and which can be produced at a price acceptable to the market.

   The relationships between marketing, design and production are examined in detail and the tensions between the different functions discussed. The problems rising from co-ordination of these functions to create an efficient activity and a healthy market, which is the duty of management and, to some extent, of Government, are touched on. The constraints laid upon the designer by trends of market requirements are examined and, in particular, the need for compatibility and the extensions of the system concept to smaller units, imposed by the growth of integrated circuits. The possibility that trends such as compatibility and ergonomic design may decrease differentiation between products of individual designers is discussed, and the increasing importance of the designer as a packaging expert is highlighted.

<div align="center">10.30–11 a.m. Coffee Interval</div>

11 a.m.–12 noon   *Chairman:* J. A. SARGROVE

Design Team Management

MICHAEL FARR. (*Michael Farr (Design Integration) Ltd., London.*)

   Industrial design is defined and the need for the management by a specialized design manager established. The design manager may be an outside consultant or a staff appointment. The tasks of the design manager are considered, particularly in relation to the other executives in a manufacturing company. This relationship ought to be close if it is to be fruitful.

<div align="center">DISCUSSION ON THE MORNING'S PAPERS</div>

<div align="center">12–2 p.m. Lunch Interval</div>

Tuesday, 11th July (*cont.*)

2–3.30 p.m. *Chairman:* J. D. TUCKER, C.Eng., A.M.I.E.R.E.

Design for Production

H. J. H. WASSELL, B.Sc. (*The Marconi Company Ltd., Chelmsford.*)

This paper underlines the designer's personal responsibility for his design, and the need for him to establish a good working relationship with many experts. It suggests that modern electronic design is dominated by complex production processes and integrated assemblies, and reviews the implications of this on the design task.

The designer should become familiar with the engineering implications of elemental assembly operations, and tailor his design to suit the particular degree of automation existing in his factory. The integrated nature of equipment assembly and testing is considered and a distinction specifically made between design proving and production testing.

Some factors affecting the length of time taken for design are discussed and the specification of quality is considered.

Production training policy for designers is outlined and the need to integrate design and production engineering emphasized.

Finally the paper summarizes a designer's Code of Practice for the part of his task that is concerned with production.

Value Engineering in the Electronics Industry

P. F. THEW. (*Standard Telephones and Cables Ltd., London.*)

Value analysis is the art of pin-pointing unnecessary high costs in products. It depends for its success, frequently a 25% reduction in product cost, on the disciplined use of common sense and problem solving methods known to most engineers. However, it places emphasis on 'Function' in a rather unique way.

The term 'Value' needs to be understood by the engineer in terms of what it means to both his company and his company's customer. A successful analysis requires team participation from Marketing, Design, Production Engineering, Estimating and Purchasing. They must clearly define the function of the product, provide a number of alternative ways of fulfilling that function, cost the alternatives and choose the lowest cost method that will meet the performance required.

There are many approaches to match different circumstances and some of these will be discussed in the paper which will cover:

The meaning of value.	The team approach.
Functional analysis.	Value analysis of electronic equipment.
Function/cost relationships.	Typical examples of successful analysis.
Picking product for analysis.	

Value Engineering is the logical extension of value analysis in that it aims to design a product that will have inherent good value from the outset. Methods of ensuring that this aim is met will be discussed in the paper.

DISCUSSION

3.30–4 p.m. Tea Interval

4–5.30 p.m. *Chairman:* J. D. TUCKER

Quality Failure Loss Analysis

D. C. STONE. (*Philips, Croydon.*)

It is normal to expect that faults will occur during the assembly of apparatus, as a result of weakness of design, faulty materials, errors in processes, and human mistakes. Each fault costs money to repair, the cost appearing in the final price. Analysis shows that costs resulting from quality failures during manufacture account for at least 5% of the final price. This paper presents a method of organizing production, inspection, fault information analysis, and costing to discover and control the quality failure costs arising during assembly of television receivers.

Detailed information is supplied to designers, production engineers, suppliers, and factory management not only of the reject percentages, but of the related costs resulting from faults occurring during manufacture. The main causes are located: it can be shown that expenditure may be permitted in design or methods improvement to produce an improvement in quality and reliability without increasing the price; in fact, price reduction may be achieved by reducing failure costs and easing the flow of production. Experience has shown that there is correlation between assembly failures in the factory and field failures in service; reduction of factory quality failure costs also reduces field service costs.

Unit Product Cost Control

M. E. DOYLE, C.Eng., A.M.I.E.R.E. (*Raytheon Company, Lexington, Mass., U.S.A.*)

Unit Product Cost Control is a system for controlling the manufacturing cost of large electronic systems during the development phase. As described it applies to a development laboratory having annual gross sales of approximately $60,000,000 and a production facility located 50 miles away. The main products are radar and weapons systems, air traffic control systems, computer displays and peripheral equipment. Obviously a large proportion of this business is Government oriented and the system to be described is an answer to the government trend towards fixed price contracts.

If system cost is not established during the development of a system, it may be extremely difficult to reduce the cost at some later date. The prime purpose of the system is to provide an agreed frame of reference for reviewing the status of the job during the development phase. Manufacturing cost targets are established for each identifiable element of hardware and the design engineer is charged with designing a product to meet this figure. To assist him a programme of continuous configuration updating and pricing is initiated. This paper deals primarily with the mechanics of this cost reporting system.

It is proposed to flow chart the computer process and describe the interface with estimating and pricing. In addition, the manufacturing/engineering interchange of information will be described and the beneficial results resulting from this such as early and intelligent make/buy decisions, test planning, preliminary industrial engineering, critical item release and system design trade-offs. It also considers the beneficial results from a machine programme which provides outputs in a format suitable for use at each level of management. Examples of these are the generation breakdown which provides the design engineer with detailed costs on his unit, and the high dollar usage report which lists piece parts in order of dollar impact on the contract and forms the basis for value engineering projects, cost trade-off studies, limiting costing and preliminary purchasing 'shopping'. Examples of the final computer print-out, variance reports and programme management reporting will be included together with a number of examples of past experiences with it.

DISCUSSION

6.30–7.30 p.m. Dinner in Hall

SYMPOSIUM OF PAPERS ON RELIABILITY AND MAINTAINABILITY

8–9.30 p.m.   *Chairman:* S. H. PERRY, C.Eng., M.I.E.E., A.M.I.E.R.E.

The Implications of Applying Basic A.G.R.E.E. Principles to the Design and Production of Commercial Electronic Equipment

J. HARRIS and A. P. TUTHILL. (*Decca Radar Ltd., Croydon.*)

Reliability has become the most important factor in marine radar. The use of improved components and the introduction of solid-state devices have increased the possibilities of improving reliability. However, the basic philosophy of the A.G.R.E.E.† report was adopted in an effort to control quality during the design and pre-production stages. It was continued throughout the production phase as a quality control tool in order to maintain the quality standard set in the design phase. After four years experience in the use of this philosophy what is its degree of success? Has the guaranteed m.t.b.f. become a possibility? How do A.G.R.E.E. production test results (m.t.b.f.) compare with field results? What are the limitations of the procedure? The paper will discuss these questions.

† A.G.R.E.E. = Advisory Group on Reliability of Electronic Equipment.

System Engineering for Reliability and Ease of Maintenance

K. F. RANKIN. (*The Plessey Company Ltd., Automation Group, Poole.*)

The paper will discuss the lessons learned from several years of investigation into the performance of a large air traffic control data processing system. To achieve the twin requirements implicit in the title the time between faults must be made as long as possible, whilst the time required to clear a fault must be reduced to the minimum. The practice of building up some degree of system redundancy with standby equipment is also one that can be employed to ensure that the occurrence of a fault does not cause an appreciable loss of system facilities.

The experimental air traffic control system, on which the original findings were based, illustrates the feasibility of designing a high standard of reliability into data processing equipment. Extensive data on the performance of the system have since been analysed and used to further improve the reliability factor.

Tuesday, 11th July (*cont.*)

Design for Maintainability with particular reference to Domestic Radio and Television Receivers

D. W. HEIGHTMAN, C.Eng., M.I.E.R.E. (*Radio Rentals Ltd., London.*)

Anyone who has occasionally endeavoured to gain access to certain troublesome parts of a motor car will know that Design for Production does not necessarily mean Design for Maintainability! The same comment is true of electronic equipment, and particularly domestic radio and television equipment, with which the author is concerned. The main theme of this paper is that of designing good maintainability into a product from the viewpoints of:

(*a*) basic component reliability,

(*b*) readily accessible test/monitoring points,

(*c*) ready access and speedy changing of components or modules in cases of failure.

The coming of colour television, with its greater complexity, emphasizes the need for attention to these points.

DISCUSSION

Wednesday, 12th July   **SESSION II: ECONOMICAL PRODUCTION**

9.30–10.30 a.m.   *Chairman:* T. NEAL, C.Eng., M.I.Prod.E.

Handing over the Design to the Works

F. J. ADAMS, C.Eng., A.M.I.Mech.E., and S. C. DUNN, M.Sc.(Eng.), C.Eng., F.I.E.E. (*British Aircraft Corporation (Operating) Ltd., Guided Weapons Division, Stevenage.*)

Design has two aspects: deducing and recording the decisions about choice; communicating the instructions to the manufacturer. The form of the instructions as they leave the design office is such as to suit what is to be made. The works require them to be conformable to the division of skills in the shop and the loading sequence dictated by competing tasks. The translation, which takes place in the planning department is further complicated by the fact that if the product is quite new, new procedures will have to be learned. Such a situation requires supplementary instructions or at least continuous, detailed engineering supervision to smooth the introduction of the new work. Although a distinction is usually made between 'development' and 'production' a considerable number of modifications must still be incorporated into the product without unduly disturbing the progress of manufacture. The problems which arise, and some solutions, are discussed.

DISCUSSION

10.30–11 a.m. Coffee Interval

11 a.m.–12 noon   *Chairman:* T. NEAL

Batch and Flow Production: the Influence of Quantity on Cost

J. M. BARBER, B.Sc.(Eng.), C.Eng., A.M.I.Mech.E., A.M.I.Prod.E. (*The Plessey Company Ltd., Telecommunications Group, Liverpool.*)

The paper will set out to examine the relationship between batch and flow production and show through case studies the functions that have to be considered in developing from a batch to flow unit.

Batch production will be discussed under the following headings: industrial development in the U.K. with batch production; typical volume warranting batch production; evaluation of the change point to flow production—quantity and cost; factors to be considered and techniques used in developing the economics of flow production; benefits and limitations of line and flow production.

Flow production calls for consideration of the following aspects: volume requirements and marketing relationship; sales programming and production scheduling; methods—tooling—flow charts and line balancing; operation selection, training, time standards and time cycles selection; economics evaluation—volume, quantity cost.

Case studies to be presented will include telephone relay machining and assembly which demonstrate the techniques to be followed.

DISCUSSION

12–2 p.m. Lunch Interval

2–2.45 p.m.   *Chairman:* J. Aikman, B.Sc.(Eng.), C.Eng., M.I.Prod.E., M.I.E.E.

The Use of Computers in Production and Management
A. R. Triggs. (*The Plessey Company Ltd., Automation Group, Poole.*)

The concept of a completely automatic production unit managed and contolled by a computer is an attractive one—and many attempts have been made to achieve these ends but with not much success either economically or technically. In many cases the existing paper work associated with manual systems has been processed by the computer, but this deals with only part of the problem and leaves the majority of the trouble spots untouched. It will be the object of this paper to re-examine the fundamentals of management and the production process and to propose both software and hardware to facilitate man's control.

Any study involving the introduction of automatic systems to work with human operatives must involve the close study of the psychology of man's motivation from the level of top management through to the operator on the bench. Many systems have failed because the designers have not given sufficient attention to the 'man-machine' interface.

Production control embraces the conditioning and preparation of information relevant to the fulfilment of customers' orders from the time the orders are received until the goods are despatched and the processing of such information to ensure correct decision and action at the right time and place. This, by definition, must embrace all aspects of management in any production organization.

DISCUSSION

SYMPOSIUM OF SHORT PAPERS ON MODERN MATERIALS, COMPONENTS AND TECHNIQUES

2.45–3.30 p.m.   *Chairman:* J. Aikman

Materials for Large-scale Integration
P. R. Stuart, Ph.D., B.Sc., A.R.C.S., A.Inst.P. (*Inorganic Materials Unit, National Physical Laboratory, Teddington.*)

The advantages and disadvantages of each material or process will be described, both from the point of view of physical properties and of ease of preparation for use. The paper will discuss the following: semiconductor integrated circuits, thin film circuits, hybrid circuits; interconnections on and off the slice; discretionary wiring.

Passive substrates: glass; alumina; glazed alumina, beryllia.

Resistance materials: evaporated materials including cermets; sputtered materials, thick films; electroless deposits; methods for direct, localized deposition.

Dielectric materials: evaporated materials; sputtered materials; r.f. glow discharge deposition; pyrolytic deposition; thick films and glazes; polymer films.

Semiconductors: III-V compounds; II-VI compounds; mounting methods, including 'flip-chip' techniques.

Finishes and encapsulants: r.f. sputtered glazes; polymers, including polyimides; glazes and enamels; flame sprayed coatings.

Integrated Circuit Packages
J. A. Cayzer. (*Texas Instruments Ltd., Bedford.*)

The various types of integrated circuit packages are reviewed, with particular reference to their assembly into systems. Currently available integrated circuits are described and examples are quoted of their assembly on to printed circuit boards. The problems of handling the future integrated circuits (arrays and large scale integration) are outlined.

3.30–4 p.m. Tea Interval

4–5 p.m.   *Chairman:* J. Aikman

Plastics in the Electronics Industry
A. S. Matthews, B.Sc., Ph.D. (*K. C. Ellis Ltd., Stockton-on-Tees.*)

Recent progress in developments of all types of plastics materials is revised with special reference to usefulness in the type of application required by the electronics industry. This industry is one of those which is well aware of the potential valuable use of plastics materials and certain more recently introduced materials are described in some detail with examples of their application. Examples of various rationalized designs made possible by the use of plastics materials are quoted and illustrated. These range from the simple use of insulating properties combined with mechanical functions, to the use of virtually 'all plastics' in electro-mechanical applications.

Wednesday, 12th July (*cont.*)

Paint Finishes

R. A. FIDLER. (*Robert Ingham Clark, Division of Pinchin Johnson and Associates Ltd., London.*)

Many products are manufactured without consideration being given to the painting process until it becomes necessary, whereas appreciation at the design stage offers production and economic advantages. One reason for this oversight is the engineer's small knowledge of paints and a brief explanation is, therefore, necessary.

In simple terms, paint is comprised of pigment, media and thinners with each having a specific role to perform, but the media being the dominating factor as to whether the paint will air dry, stove and how it can be applied. Each method of application has its own particular advantages which need to be explained, and it is at the design stage that provision must be made so that the article to be cleaned and painted can be efficiently handled and processed.

By a greater understanding of the methods of application it is then possible to estimate the cost to paint a given area with the ultimate aim of obtaining the cheapest but best finish, and the basis of such calculations will be explained.

Note: Other short papers are under consideration for inclusion in this symposium.

Thursday, 13th July SESSION III: FUTURE DEVELOPMENTS

SYMPOSIUM OF PAPERS ON AUTOMATIC ASSEMBLING AND TESTING

9.30–10.30 a.m. *Chairman:* A. E. MAVER, C.Eng., M.I.Prod.E.

Mechanical Assembling

W. V. TIPPING, C.Eng., M.I.Mech.E., M.I.Prod.E. (*Formerly with Roevac Automation Ltd., now with Rolls-Royce Ltd.*)

A brief history of assembly by machinery will be followed by an explanation of the basic problems involved. These are product and component design to suit the method. Component manufacturing consistency. Development risks involved with special purpose machinery. The need of communication between product design, equipment design and the production shopfloor.

Component and product design for mechanical assembly will be dealt with, giving the basic rules to follow. Some case studies biased towards the small electrical field will be given.

Principles of conveyor systems, parts placing, feeding of parts, orientation, etc., will be explained. Standards are being evolved to help to reduce cost, delivery and risk.

An appendix describes the conduct of a feasibility study and a hypothetical case study showing the elementary techniques of assembly line balancing is applied to finding the economic optimum for the project.

Automatic Assembling on Printed Wiring Boards

Ir. H. C. HAVERKORN VAN RIJSEWIJK. (*N. V. Philips Gloeilampenfabrieken, Eindhoven.*)

Of the assembly process of radio and television receivers only the assembly of the components on to the printed wiring boards and the soldering process have been subjected to mechanization.

Assembly of components includes six basic functions:

1. Component transport to insertion head.
2. Component preparation.
3. Printed wiring board transport to insertion head.
4. Printed wiring board alignment.
5. Component insertion into board.
6. Component fixing to board.

All assembly machines can be classified in three basic types, the insertion line, the multiple insertion machine and the programmed insertion machine, each executing the functions in different ways.

An analysis of existing assembly machines shows they do save labour but not enough costs for projects of this magnitude. Further analysis shows that components can be more cheaply prepared (function 2) and supplied with means for fixing to the printed wiring board (function 6) by the component manufacturer than by the set-maker's assembly machine.

Insertion machines based on these components can be extremely cheap and have considerable output. Printed wiring board transport and alignment (functions 3 and 4) can now be done by hand since it occurs only once per board assembly. The specification of these components, truly fit for mechanization, is discussed and the importance of standardization underlined. The key to cheap assembly of printed wiring boards is in the hands of the component manufacturer.

DISCUSSION

10.30–11 a.m. Coffee Interval

11 a.m.–12 noon   *Chairman:* J. A. SARGROVE

The Application of Automation to Assembly and Testing in Computer Manufacture

A. M. F. COLVIN, B.Sc., A.C.G.1. (*Elliott-Automation Computers Ltd., Borehamwood.*)

This paper will outline the problems relating to production testing of digital computers. It will indicate that these fall into two main areas, namely, elimination of faults in the printed-circuit logic cards, and the clearing of wiring errors in the complex of inter-wiring between these logic cards.

In the first case, a fully automated computer-based system is used to test the logic cards at a very high speed. In the second case a semi-automatic technique is applied in which the wiring operation is inhibited should the operator attempt to introduce any incorrect wire link into the system.

The theory and practice of these two systems will be discussed.

Automatic Testing of Electronic Equipment for Aircraft

J. W. ANSTEAD. (*Smiths Industries Ltd., Aviation Division, Cheltenham.*)

This paper discusses applications where automatic testing can be a requirement, some of the advantages and disadvantages to both manufacturer and customer, and the demands its use makes upon the aircraft electronic equipment designer.

Some of the problems which exist at present are due to the lack of consideration given to the use of automatic testing by the equipment designer. These can be overcome in the future if design, production and test methods are organized to take proper account of them. The constraints that these requirements place on the design processes are discussed.

The problems encountered by the airlines in maintenance work on the equipment are similar to those of the manufacturers, but the former have an additional major problem associated with inaccurate fault diagnosis in complex systems in the aircraft. This has encouraged them to go ahead and buy general-purpose automatic test machines while most manufacturers are still hesitant. Many units removed from aircraft are found to be serviceable, and automatic test centres are being established at hangar-floor level to filter the good from the bad units, so that serviceable equipment can be returned to stores with the minimum of delay. Airlines have found that they are able to repair a high proportion of electronic units at this stage if easily replaceable modules have been used in the construction. The overall effect of this philosophy is a dramatic reduction in spares holdings and in the load on the main repair workshops.

It is likely that there will be a trend towards the inclusion of automatic built-in test features in future aircraft electronic systems to improve fault diagnosis. This will become feasible with the increasing use of digital techniques in aircraft, but it does present weight, cost and integrity problems.

The author concludes that the use of automatic testing has a generally beneficial effect on equipment design and production as it emphasizes the need for improved maintainability. The use of standardized, interchangeable modules reduces design, production, documentation, and maintenance costs, apart from the direct savings in time and skilled labour which result from automatic testing.

DISCUSSION

12–2 p.m. Lunch Interval

2–3 p.m.   *Chairman:* H. V. BECK, M.A., B.Sc., C.Eng., F.I.E.E.

The Changing Pattern of Equipment Design

R. P. TOWELL, C.Eng., A.M.1.E.R.E. (*E.M.I. Electronics Ltd., Microelectronics Division, Hayes.*)

The widespread introduction of integrated circuits into the industry will bring about standardization of one sort or another in equipment design undreamed of five or ten years ago. A decrease in overall timescale will result, achieved mainly as the result of a decrease in development time. This will be due primarily to the fact that so much of the equipment will be pre-developed before the experimental phase of the project has started.

Assuming that the interconnection and packaging design methods and processes are matched to silicon integrated circuits in their state of readiness the comparable activities to main development will be carried out in the first phase rather than the second phase of a project. This implies an awareness and capability in the experimental phase pointed directly at production.

The paper will consider the effect on the system designer, circuit designer and drawing office of such implications, and describe the facilities of a design office which are matched specifically to this new requirement.

Thursday, 13th July (*cont.*)

The Impact of Integrated Electronic Circuits on the Company

S. S. FORTE, B.Sc., Ph.D., C.Eng., M.I.E.E. (*The Marconi Company Ltd., Microelectronics Division, Witham.*)

> The paper will discuss some of the ways in which established practice within a manufacturing organization must be re-orientated if the full benefits of current developments in integrated circuits are to be realized.

DISCUSSION

3–3.30 p.m.   *Chairman:* M. JAMES, Dip.El., C.Eng., M.I.E.R.E.

Some Applications of Computers in Electronics Design

N. E. WISEMAN, M.A., M.S., B.Sc.(Eng.). (*University Mathematical Laboratory, Cambridge.*)

> Design is seen as a complex activity requiring intellectual effort at several levels. The highest level is what we, rather loosely, call creative and corresponds with the selection ('invention') of potential schemes. There follows an analytical activity ('how does it work') and finally a clerical operation ('how good is it?') which yields quantitative data on which the acceptability of the scheme is judged. Interactions occur between activities at the several levels until a satisfactory solution is achieved (or hope is given up of ever finding one). Some of the operations, primarily from the lower intellectual levels, may be handled by computer methods but very few applications are known in which the entire design process has been performed by the computer. Typical applications at the present time necessitate constant interplay between man and computer and this requirement for a man-machine partnership raises special problems in hardware and software departments.

> The paper identifies the roles of man and computer in typical design applications and outlines some of the special hardware, software and system techniques which have been developed. Particular C.A.D. topics in circuit design, logic design, placement and conductor layout are discussed. Finally, the generality of total systems and the interfaces between the C.A.D. office and other plant activities are considered.

3.30–4 p.m. Tea Interval

4–4.30 p.m.

Computer Techniques in the Drawing Office

C. ALLEN, A.H.-W.U., C.Eng., M.I.Mech.E. (*Ferranti Ltd., Edinburgh.*)

> Many manhours of highly skilled, and also scarce, effort is used in the drawing office for the later stages of engineering design and in the preparation of manufacturing information. Some of this effort is used to carry out routine calculations searching for alternatives to reduce the cost, ensuring that standards are used, reducing variety and preparing working drawings and text. Similarly in the experimental workshops and in production, skilled effort is required to interpret this data and, quite often, change it into another form, this being particularly so when dealing with a production organization using computer control and machine shops equipped with numerically controlled machine tools. This paper deals with some methods of saving effort and time and improving the passage of data across the design/production interface. Consideration is also given to the automatic preparation of detail drawings and other documents necessary for the control and modification of a product. The methods described can be applied over the next few years even in the smaller organization, there being no need to be 'on-line' to the computer. Some details of hardware and software are mentioned. It is believed that the use of the aids described will, in addition to the immediate savings, allow techniques to be evolved, which will help in the introduction of the more advanced c.r.t. displays operating 'on-line' on a time-shared basis.

DISCUSSION

4.30–5.30 p.m.   *Chairman:* J. A. SARGROVE

New Vistas in Design

I. M. ROSS, M.A., B.Sc., C.Eng., F.I.E.E., F.Inst.P. (*Consultant.*)

> Technological investment, if it is to be viable, must not only take into account the local situation and answer to the cut and thrust of immediate forces but, like the Channel swimmer, be aware also of the deeper more insidious currents, understand how and where they are moving, and how therefore to exploit or avoid their influence. This paper looks at the nature of design and its methodology, the way things are changing, and the evolution also of user needs, as an aid to longer-term policy making.

Closing Address—J. LANGHAM THOMPSON, C.Eng., F.I.E.E., M.I.E.R.E. (*Past President, I.E.R.E.*)

A Hot Electron, Cold Cathode, Emitter

By

R. R. VERDERBER,
B.Sc., M.Sc.†

AND

J. G. SIMMONS,
B.Sc., Ph.D., F.Inst.P.‡

Summary: A hot electron, thin-film, cold-cathode device is described which has an emission efficiency much higher than previously reported for similar cathodes. Measurements of the emission current and distribution at various voltages indicate the importance of a coherent scattering process, which if not considered in the design of the cathode will severely limit the emission efficiency. The important conditions for a more efficient cold cathode are specified and discussed with respect to previous cold cathodes designed to surmount other loss processes which are shown to be of secondary importance to coherent scattering.

1. Introduction

The principle involved in a thin-film, metal-insulator-metal, cold cathode is that of injecting hot electrons from the lower electrode into the upper electrode such that the electrons have sufficient energy to surmount the upper electrode, vacuum-interface potential barrier; that is, the hot electron must have greater energy than the outer electrode vacuum level, as shown in Fig. 1. Also, the upper electrode must be thin enough to be permeable to the electrons (less than the electronic mean free path, which in metal is usually about 200 to 300 Å). Previous thin-film cold cathodes[1-7] have attempted to utilize the tunnel effect as a means of transferring the electron through the insulator from one electrode to the other electrode, because in this process the electron energy is conserved, and, hence, the (hot) electrons are presumably injected into the second electrode with energy equal to eV, where V is the bias voltage.

This paper is concerned with describing the properties of a thin-film, Al-SiO-Au, hot electron, cold cathode in which the insulator is several hundred angstroms thick. The emission current has been measured as a function of the device voltage, and the spatial distribution of the emitted electrons has been observed by viewing the optical image arising when these electrons are incident upon a phosphor-coated anode. The results are discussed in terms of (i) a model which describes the origin of the emitted electrons and (ii) a coherent scattering process, which must be considered in the design of the cathode if good emission efficiencies are to be obtained. The importance of the coherent scattering process is compared to other mechanisms that have been proposed to account for low emission efficiency from hot electron, cold cathodes.

2. Fabrication

The cold cathode consists of a sandwich structure of vacuum-deposited films of Al, SiO and Au. The device is deposited on a cleaned 1 in × 3 in microscope slide in a configuration shown in Fig. 2. The

Fig. 1. Energy diagram of a metal-insulator-metal hot electron cold cathode.

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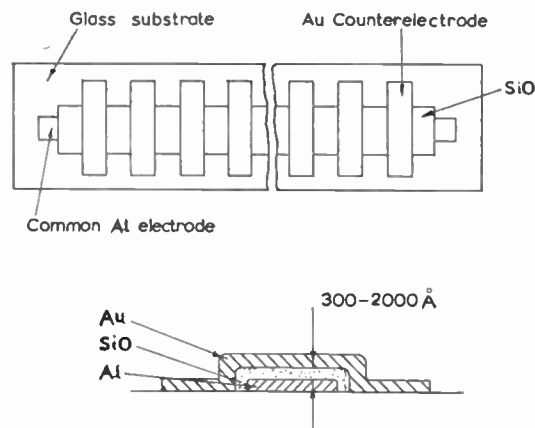
Fig. 2. Device configuration.

common electrode is an Al stripe ⅛ in wide and 2000 Å thick. The thickness of the SiO layer is not critical, and emitting devices have been made using SiO films of thicknesses ranging from 200 Å to 2000 Å. The thickness of the SiO is predetermined by thermally evaporating to completion a weighed amount of SiO from a molybdenum boat or by monitoring the thickness during deposition using the quartz crystal technique, the average rate of deposition of the SiO being 4 Å/s. Ten counter electrodes of Au, ⅛ in wide and 1000 Å thick, are then deposited on the SiO film; thus, ten devices are fabricated on each substrate having a device area of 0·1 cm². The devices are made without breaking vacuum, and all films are deposited at a pressure of 10^{-6} torr.

3. Device Characteristics

3.1. Forming

The impedance of the virgin device is typical of a good thin film capacitor, as is to be expected, and in this state no electron emission is observable. The devices are then 'formed' in a vacuum of 10^{-2} torr at ambient temperature by applying 10 V across the device with the Au top electrode biased positively. Accumulated data suggest the forming process is due to injected positive ions, emanating from the Au electrode.[8]

The forming process results in a large decrease in the resistance of the sample, the current-voltage characteristic of which is illustrated in Fig. 3. The device $I-V$ characteristic can be conveniently divided up into four voltages where there are pronounced changes in the functional dependence of the current range on voltage: (i) 0 to 2·5 V—$I \propto \sinh KV$ where K is a constant, (ii) 2·5 to 3 V—deviation from the $\sinh KV$ dependence and the maximum current at 3 V, (iii) 3 to 7 V—negative resistance, (iv) above 7 V—current again increases with voltage. From the thermal and isothermal current-voltage characteristics it has been shown[8] that the current is transported through the device by a tunnel hopping process via the Au ions in the insulator. This conduction process is fairly complex, so we make use only of the points that are of direct interest to us in discussing the emission process, as any detailed discussion would be out of place here.

3.2. Emission Current

The devices were placed in a vacuum chamber at 10^{-5} torr, and the emitted electrons were collected by a metal anode positioned 2 cm above the device and biased 100 V positive with respect to the cathode. The emission current was recorded for various voltages across the device and the results are plotted in Fig. 3, for both increasing and decreasing voltage. It will be observed that the emission current exhibits

four distinct voltage ranges corresponding to the circulating current characteristic: (i) 0 to 2·5 V—no ission, (ii) 2·5 to 3 V—emission appears and reases exponentially, (iii) 3 to 7 V—emission increases only slightly and is noisy, (iv) above 7 V—a sharp increase in the emission current. Since the two sets of voltage ranges are similar, it appears that a common mechanism is responsible for the functional changes in both the emission and circulating current-voltage characteristics. This is indeed the case, for, as we shall see below, the band model built up from the circulating $I-V$ data is compatible with the emission $I-V$ characteristic. Figure 3 also illustrates an additional noteworthy characteristic which is the observation of an emission current for voltage biases less than the work function of Au (4·82 V).

The emission process was shown to be due to hot electron emission rather thermally stimulated on two counts: (i) the emission current in Fig. 3 is observed to increase in the range 3–7 V, although the power dissipated in the device is actually decreasing, and (ii) on reversing the polarity of bias across the cathode no emission is observed, even though the device circulating $I-V$ characteristic is symmetrical with polarity of bias, i.e. the same power is being generated in the device independent of polarity.
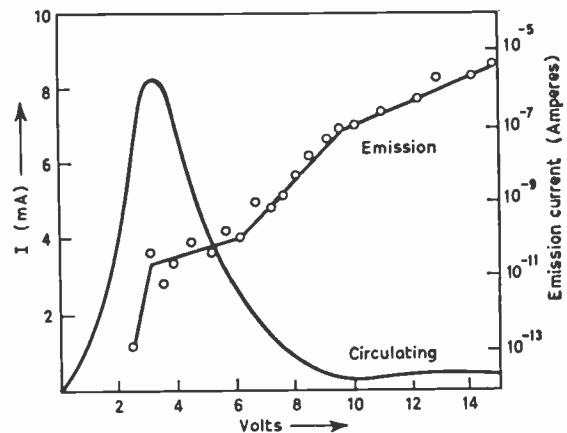
Fig. 3. Curves of the circulating current and emission current for different bias voltages.

Examination of the cathodes always revealed pinholes in the upper electrodes, and by forming a magnified image of the emitting surface of the cathode at a phosphor screen (see Section 3.3) by means of an electrostatic lens, it was possible to show that the pinhole edges were the source of the emitted electrons.

3.3. Distribution of Emitted Electrons

The distribution of the emitted electrons was observed by replacing the metal anode with a con-

ducting glass anode coated with a zinc sulphide phosphor. The phosphor-coated anode was positioned parallel to the plane of the cathode and biased at 1 to 2 kV positive with respect to the cathode in order that the emitted electrons would acquire sufficient energy to cause the phosphor to glow.

With the anode positioned a few centimetres from the cathode and the device biased less than 8 V, the emission pattern consisted of several small spots within the 0·1 cm² area of/the device, which indicated that the electrons were being emitted from small localized regions. If the bias voltage was increased above 7 V, an image much larger than the cathode was observed. The particular interesting aspect of the image was the amount of structure exhibited in it, as shown in the photograph Fig. 4. Imposed on a faintly illuminated background are many small brightly illuminated arcs all of the same radii, none of which subtends an arc greater than 90°.

All the devices displayed similar images on the screen. It has been shown that these patterns are caused by the coherent scattering of the hot electrons injected into the gold electrode, followed by their

subsequent escape from exposed *edges* of the electrode, such as the circumference of pinholes.[9] Briefly, what this means is that the electrons entering the upper electrode undergo very low energy electron diffraction and are scattered into a hollow cone of diffraction of half angle 2θ (see Fig. 5). The angle (2θ)
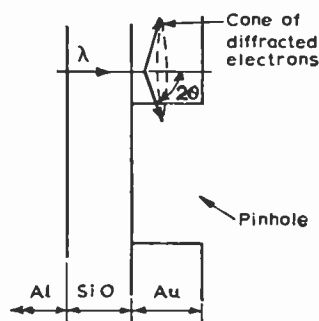
Fig. 5. Schematic diagram of the cathode, in the vicinity of a pinhole, illustrating the path of the diffracted electrons.

through which these electrons are scattered in the upper electrode is related to the bias voltage, V_b, and the material constants by the Bragg equation

$$\sin \theta = \frac{\lambda}{2d_{hkl}}, \qquad \ldots\ldots(1)$$

where d_{hkl} is the distance between the diffracting crystal planes and λ is the wavelength of the electron in the top electrode expressed by:

$$\lambda = \frac{12\cdot27}{(V+\eta)^{\frac{1}{2}}} \qquad \ldots\ldots(2)$$

where η is the Fermi energy of the top metal electrode. For the voltage biases used herein, it can be shown that the component of electron momentum parallel to the plane of the film is much greater than the perpendicular component, which means that the electron can only escape at an exposed edge of the electrode.[9]

4. Origin of Emitted Electrons

4.1. *Energy Considerations*

Figure 6 is an energy diagram of a formed device with the gold electrode biased positively V_b volts. This energy band scheme has been constructed from accumulated experimental data on the circulating current characteristic.[8] The diagram shows allowed states, resulting from the injected ions, in the forbidden gap of the SiO which extend up to an energy E_i below the conduction band, with no possible electronic states between E_i and the bottom of the conduction band.
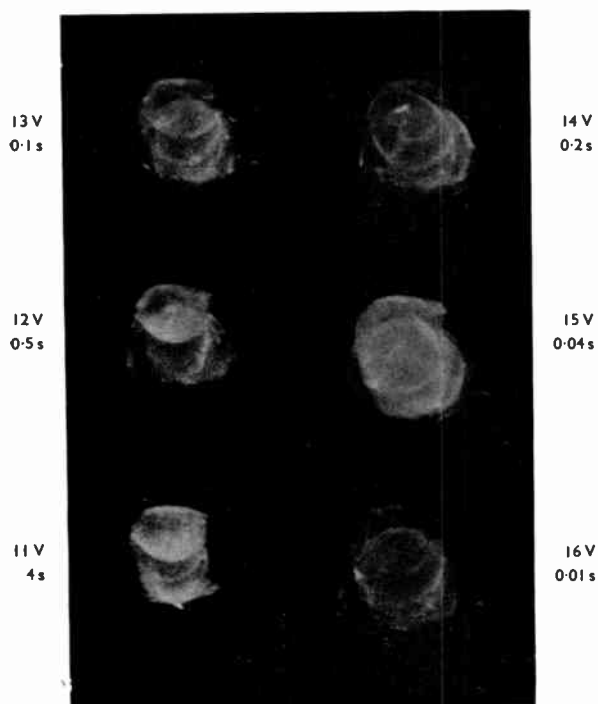
Fig. 4. Full size photographs of the emission pattern from cold cathode devices for different voltages, V_b, across the device (cathode-screen distance = 4·42 cm, cathode-anode potential = 2·0 kV). Exposure times are given alongside photographs, showing, qualitatively, that the emission is increasing rapidly with V_b.
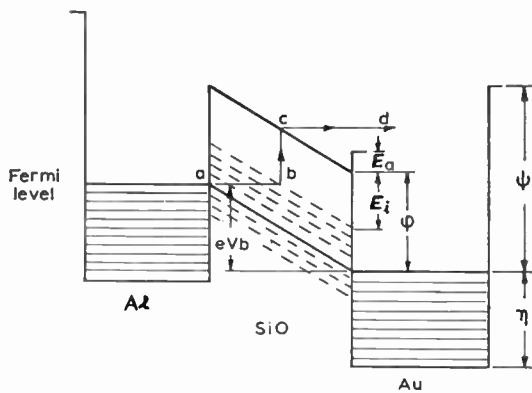
Fig. 6. Energy diagram illustrating the transport process of the emitted electron for V_b between 2·5 and 7 V.

The path abcd represents the path through the insulator of electrons which are injected into the insulator from the electrode with an energy equal to the Fermi energy. For $V_b < 2·5$ V the electrons can pass through the insulator, but for $V_b > 2·5$ V the electrons can only reach the point b in the insulator. For an electron at b there is a probability that the electrons can be excited into the conduction band by acquiring an energy E_i from the lattice. These electrons are then accelerated in the conduction band of the insulator, and those which do not undergo any losses in the insulator will enter the upper electrode with an energy $eV_b + E_i$ above the Fermi of the electrode, rather than just eV_b, as would be the case if they had tunnelled right through the insulator. If these electrons can now reach an exposed edge of the electrode without loss of energy they can escape providing

$$eV_b + E_i > \psi \qquad \ldots\ldots(3)$$

The particularly noteworthy point of eqn. (3) is that E_i is of the order of at least 1 eV,[8] which means that for voltage biases substantially less than ψ/e the electrons can still escape, thus resolving the apparent anomalous emission occurring at voltages less than 4·8 V (the work function of the upper electrode expressed in volts), see Fig. 3.

4.2. Scattering Processes

The structure in the emission $I-V$ characteristic of Fig. 3, i.e. the rapid rise in emission current above 8 V, is due to electrons undergoing multiple scattering below 8 V and only a single scatter above 8 V. Using $V_b < 8$ V, $\eta = 5·5$ eV and d_{111} (the most intense diffracting plane of Au) $= 2·35$ Å in eqn. (1), we find $\sin \theta = 0·71$ and, hence, $2\theta > 90°$. Thus for $V_b < 8$ V the electrons are scattered back towards the insulator and would not normally escape from the electrode, unless they underwent further scattering

(electron-electron, electron-phonon) to reach an electrode edge. These additional scattering processes lead to substantial energy losses, and the process is obviously an inefficient means of directing electron towards an edge. For $V_b > 8$ V, the scattering process is coherent only, and the energy is conserved in this process; thus, providing an electron scattered in this manner is within a distance not greater than the electron mean free path from the edge, it will reach the edge without loss of energy and, hence, escape into vacuum.

4.3. The Emitted Current

The current traversing the path abcd is given by:

$$I_e = eN_i \exp[-(E_i - \Delta E)/kT] \qquad \ldots\ldots(4)$$

where N_i is the number of electrons arriving at b per unit time, k is Boltzmann's constant, T is the temperature in deg K, and ΔE is the Schottky lowering of the barrier E_i. N_i is of course an increasing function of voltage bias, as is ΔE, which is given by:

$$\Delta E = \left(\frac{14·6}{Ks} V\right)^{\frac{1}{2}} \qquad \ldots\ldots(5)$$

when kT is expressed in electron volts, and s in angstroms; K is the dielectric constant of the insulator. In actual fact, only a fraction of the current I_e will arrive at the upper electrode with no loss of energy, because of electron-phonon interactions in the insulator conduction band. It is this fraction of the electrons which emerge from the insulator as the measurable emission current. It is clear from eqn. (4), then, why the emission current continues to increase in the voltage range from 3 to 7 V, even though the circulating current is decreasing.

5. Comments on other Devices and Loss Mechanisms

Previous studies in thin-film, hot electron cathodes have been carried out on devices which depend on electrode-electrode tunnelling as the means of electron transport between electrodes.[1] Although large emission efficiencies were expected from these devices, the best of those reported have not exceeded more than approximately 10^{-4}. Energy losses in the top electrode can be apparently eliminated as the major cause of this discrepancy because of the work of Cohen.[4] He used cathodes which had a top electrode of only 25 Å thickness and obtained very low efficiencies, yet measurements of the attenuation length of electrons[10, 11] show that the electrons should be able to travel a few hundred angstroms in metals with little loss of energy.

Collins and Davies[6] concluded from energy distribution measurements of the emitted electrons that energy losses in the conduction band of the insulator were responsible for the low emission efficiencies.

However, the efficiencies of our devices are 100 times greater than reported for tunnel cathodes, yet our insulator and electrodes are typically *ten times thicker* than those used by Collins and Davies.

We feel that the important difference in the present device and other cold cathodes is that the thicker insulator in our device permits one to operate at a higher voltage level. In the ultra-thin insulator devices high voltages cannot be applied because of the problem of dielectric breakdown; for example, a voltage bias of only 6 V induces a field of 2×10^7 V/cm in a typical tunnel cathode thickness of 30 Å, and this exceeds the dielectric breakdown strength of most insulators. It follows then, even if the electrons in these devices did not lose any energy whatsoever, because of the low voltage level of operation and coherent scattering in the electrode, they will be diffracted back into the insulator and, hence, their escape inhibited. Thus the present study indicates that the failure to collect the coherently scattered electrons in the top electrode has been a major cause of the low emission efficiencies observed in the past.[1]

6. Design Considerations

The emission efficiency of the device can be improved in two ways. First, by using an electrode which has a large edge to area ratio, such as a grid structure. Second, by using a thicker insulator in order that the device can work at a sufficiently high voltage level, so that the scattering angle (2θ) is reduced and, hence, the forward component of momentum of the electron in the top electrode increased sufficiently to overcome the surface potential of the electrode. Thus we are not relying on a relatively small pinhole edge surface to provide the emission current, but rather the whole surface of the electrode. It will be apparent that a very thin electrode is required in this case.

7. Conclusions

The study of emission currents from Al-SiO-Au devices show the electrons are emitted from pinhole edges in the top electrode. At higher voltages the very large increase in the emission current is due to the collection of electrons that have been injected into the gold electrode, coherently scattered at large angles, and finally emitted from the gold. Emission efficiencies as large as 1% have been measured which is an increase of two orders of magnitude above those which have been reported for ultra-thin hot electron cold cathodes, and we have concluded that the low efficiency of previous thin-film cold cathodes is due to the failure to collect coherent scattered electrons in the upper electrode. It is expected that, by suitably designing the geometry of the device, the emission efficiency of the device will be improved.

8. Acknowledgment

9. References

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Paris Components Show–April 1967

The value of components exhibitions to the electronics engineer, particularly if he is concerned with the design of equipment, can hardly be overestimated. On these occasions opportunity is given to see and compare the diverse designs of active and passive components, and mechanical items, and to discuss their features with the makers' engineers on 'neutral ground'; if the exhibition is an international one the range of items is correspondingly wider. The 'Salon des Composants Electroniques' held annually in Paris draws together exhibitors and visitors from all over the world and each year sees greater numbers of both. This year attendance was nearly 150,000 and there were altogether 708 exhibitors. The number of exhibitors in 1966, 775, included 128 instrument firms who were excluded this year; allowing for this restriction, due to the 'Mesucora' Exhibition being held the following week, increased numbers of exhibitors came from France (339, an increase of 14), the U.S.A. (138, +19) and Great Britain (75, +9). An idea of the international scope of this exhibition may be gained from the fact that there were 369 exhibitors from outside France; these were drawn from 14 countries.

Technologically, the most striking feature of the Salon was the emphasis by many manufacturers on the integrated circuit in its various guises, solid-state and thin film. American companies and their European subsidiaries were, of course, prominent in this field, but British and French firms showed developments of considerable interest. It is worth mentioning that in both countries the respective governments are discussing with industry the desirability of rationalizing integrated circuit development and production to meet U.S. competition.

Colour television is the main development now pending in the domestic equipment sector, but it was not as prominent as might have been expected, at least as far as new developments were concerned. Tuners and scanning components were certainly in evidence, but the most interesting recent French development, namely a new tri-colour display tube, developed by Cie. Française de Television, was not in fact shown at the Salon. In this tube beam modulation energy and deflection energy are lower than are required in the shadow-mask type—low enough indeed for transistor circuits to be used without difficulty.

For the third successive year the I.E.R.E. took part in the Salon and at the invitation of the organizers displayed *The Radio and Electronic Engineer* and other publications, such as Conference Proceedings, to many interested visitors. The Institution's stand, shown below, was in the charge of Mr. W. G. J. Nixon (Member); Mr. Nixon (left) was joined in this photograph by Dr. A. V. J. Martin (centre), a founder member of the French Section of the I.E.R.E., and Mr. F. W. Sharp (right), the Institution's Assistant Editor.

I.E.R.E. Stand at Paris Components Exhibition, April 1967.
Left to right: W. G. J. Nixon (*Member*), Dr. A. V. J. Martin (*Associate Member, French Section*) and F. W. Sharp (*Associate Member*), Assistant Editor *The Radio and Electronic Engineer*.

Digital Read-out Echo Sounder

By

C. H. COOKE

(*Graduate*)†

Summary: The trend towards automation in hydrographic survey and oceanography calls for automatic read-out of depth from an echo sounder. A system of signal selection is described which includes an automatic time gate synchronized with the sea-bed echo. Automatic gain control is also used to maintain optimum amplitude discrimination throughout varying conditions. A recording echo sounder forms an integral part of the apparatus and is used to check the signal selection by identification of the selected echo.

1. Introduction

For many years the standard method of display for interpretation of pulsed echo-sounding signals has been the chart recorder. There has also been a demand for a depth-measuring system producing readings in purely electrical form. In the past, this demand has mainly resulted from the need to provide depth repeaters in different parts of a vessel for navigation purposes. In recent years the problem has received renewed attention due to the trend towards automation in oceanography and hydrographic survey.

The outputs of many of the instruments used in these fields may be presented readily in digital form. Survey teams are making increasing use of radio navigation systems and the fix readings from such systems may be logged automatically. A considerable saving in time would result if depth readings were available in digital form to be logged together with the fix. A digital read-out sounder is also a necessary link in complete systems for data collection in oceanography.

A development programme was commenced in August 1964 to examine the problems associated with reliable digital read-out of depth, and to design an experimental equipment for evaluation. The work has been mainly concerned with shallow-water sounding but is generally applicable to medium depths also.

The basic simple concept is to measure the time between transmission and the reception of the sea-bed echo. Time interval measurement in digital form is well established, but the problem is complicated by the nature of the received echoes since these signals may be received from many other reflectors besides the sea bed. Some of these reflectors, e.g. masses of aerated water at short range, may cause masking of the sea-bed echo. For these reasons a process of signal selection must be employed and the success of automatic echo sounding depends on this part of the system.

† Kelvin Hughes, a division of Smiths Industries Ltd., Ilford, Essex.

2. Signal Selection

Due to the nature of echo-sounding signals, chart recorders have been used mainly in the past. In such recorders a great deal of information is provided by trace-to-trace correlation[1] of the signals which enables the operator to discern the sea-bed echo in the most adverse conditions, e.g. when the signals could not be distinguished one from another on the basis of amplitude alone. Even under reasonable conditions the operator may unconsciously use a complex process of logic in his reading of the chart.

Whilst a great deal can be done to obtain high reliability in an electronic method of signal selection, a check on the operation of the system is of considerable advantage. The retention of a recording echo-sounder to work in parallel with the automatic system, enables the echo signal which has been selected to be identified on the chart.

The sea-bed echo possesses distinguishing features over other signals for most of the time. In the main, it will be the largest echo-signal received in any one sounding interval from the point of view of both amplitude and duration. Also, the rise time of the leading edge will generally be less than for other echoes. Perhaps the most important feature is that from one sounding to the next, the time relationship with respect to the transmission will not alter significantly when considered as a percentage of the total transmission interval. This fact, together with amplitude selection, forms the basis of the signal selection system used in the experimental equipment.

2.1 Amplitude Selection

Figure 1 shows the general arrangement of the signal-selection system. Received echo-signals are amplified and applied to an amplitude gate via a detector and filter circuit. The amplitude gate is formed by a Schmitt trigger circuit and in the experimental equipment, adjustment of the threshold level is not provided. The output of the gate is a d.c. pulse
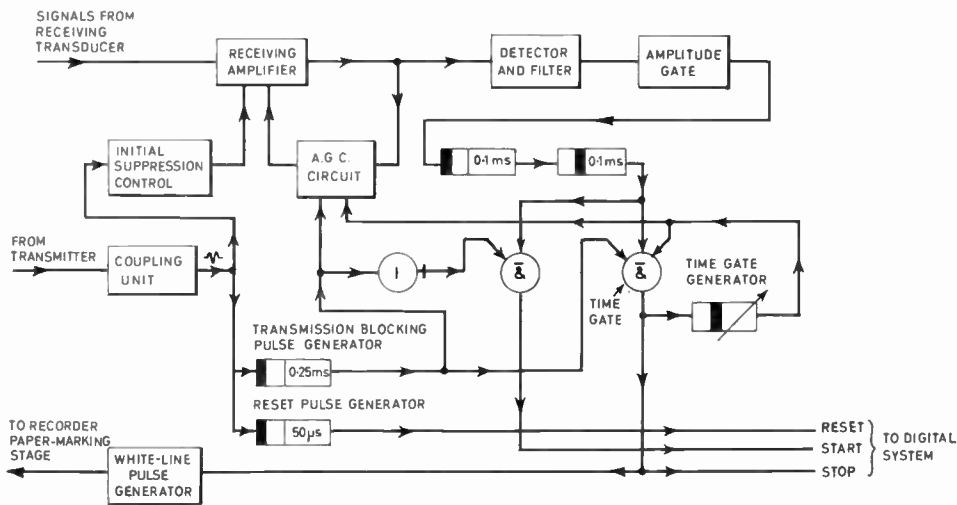
Fig. 1. Signal selection system.

of the same duration as the input signal. The leading edge of this pulse triggers a monostable circuit which, after a delay period of the order $0 \cdot 1$ ms, triggers a similar circuit. The output from the second circuit is again a pulse of $0 \cdot 1$ ms duration which corresponds to an echo-signal capable of operating the amplitude gate, but delayed with respect to the leading edge of that signal by the first monostable device.

Assuming that the gain of the receiving amplifier is set correctly, output pulses will be produced by:

(a) the transmission break-through

(b) the sea-bed echo, and

(c) the echoes closely comparable in amplitude with the sea-bed echo.

The effect of transmission break-through is removed from the subsequent parts of the system by a blocking pulse, generated by a monostable circuit, triggered from the transmitter.

2.2 Automatic Gain Control

To ensure that optimum performance is obtained from amplitude selection of the sea-bed echo, it is important that the amplitude relationship of the signals is not distorted by the receiving system. The gain of the receiving amplifier should provide just sufficient signal at the output to reliably operate the amplitude gate, allowing for the normal 'ping-to-ping' fluctuations which occur under any given conditions. However, the average level of the echo signals may vary considerably in a short time and to avoid the necessity for frequent gain adjustments, automatic gain control[2] has been incorporated in the receiving amplifier. Automatic gain control systems designed for use on c.w. signals are not generally successful

when applied to pulsed echo sounding since a long discharge time-constant is necessary to store the a.g.c. level from one 'ping' to the next. This makes the system very slow to respond to a falling signal level. The a.g.c. should have a fast and equal response to increasing or decreasing signals and this has been obtained using a pulse counting technique. The general arrangement of the a.g.c. is shown in Fig. 2, and associated waveforms in Fig. 3. In operation the gain is reduced by a fixed amount when echo signals during any sounding exceed a pre-set level at the output of the amplifier. An increase by a similar amount is applied if they do not. The a.g.c. pre-set level is defined by a second amplitude gate which, when operated by an echo signal, causes a monostable pulse generator to be triggered. The threshold of this gate is considerably higher than that of the signal selection circuit. The duration of a pulse from the generator is approximately 80% of the transmission interval period. It is applied to an integrator circuit which produces an output level representing the number of such pulses produced over a period of several soundings (see Fig. 3, lines 4 and 5). This level is used to control the gain of the receiving amplifier.

Transmission break-through is prevented from operating the pulse generator by the use of a short delay and a suppression gate placed between it and the amplitude gate. As d.c. pulses only are involved at this stage, logic circuits are used. A NAND circuit forms the suppression gate, which is closed by the transmission blocking pulse.

Two factors which must be considered are the rate of response of the a.g.c. and the separation in terms of signal level between the thresholds of the a.g.c. and signal selection amplitude gates. The effect of the
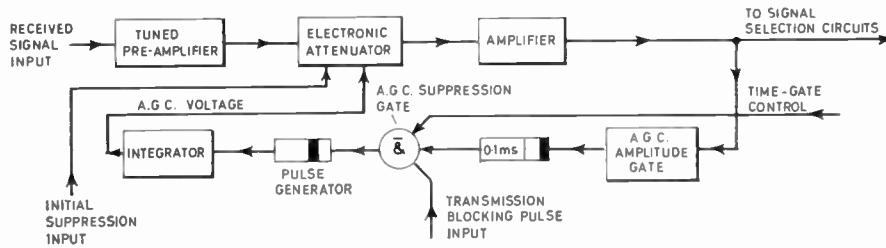
Fig. 2. Automatic gain control system.
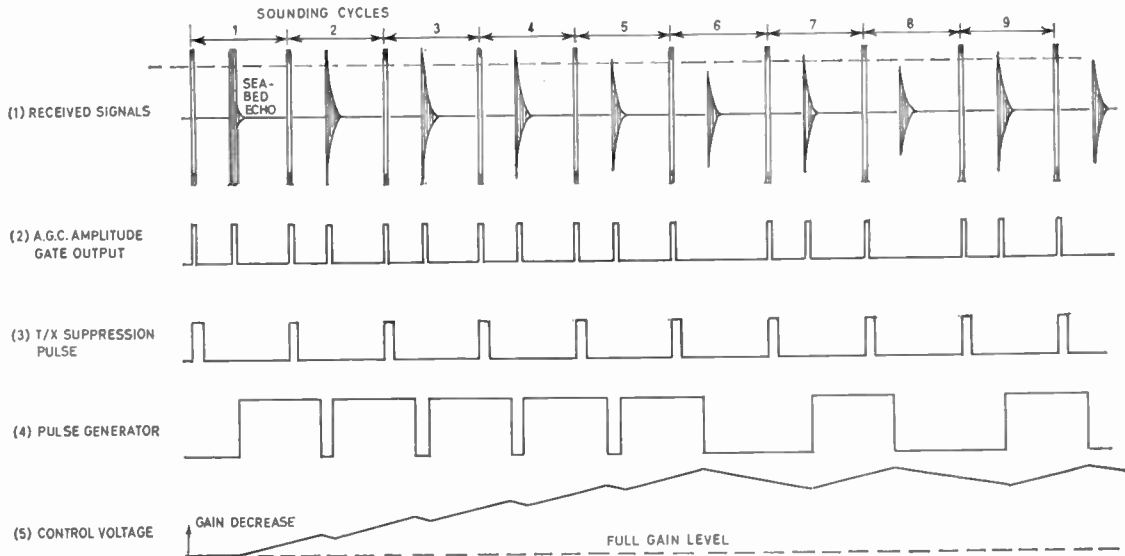
Fig. 3. Automatic gain control waveforms.

a.g.c. circuit is to cause the signal level to hunt around the threshold of the a.g.c. amplitude gate. The hunting amplitude increases as the a.g.c. response time is decreased and the effect must be added to the normal 'ping-to-ping' variations in signal when determining the separation of the two gates. In shallow water with a high sounding rate, an a.g.c. response of 2–3 dB per sounding interval has been found satisfactory, the threshold separation in this case being approximately 15 dB. For medium depths, a response of approximately 6–8 dB per ping is required with a separation of up to 20 dB.

In order to achieve a reasonable degree of linearity between the control voltage and the receiver gain, the active stages of the amplifier are left uncontrolled, and the signals are passed through an attenuator constructed of a network of resistors and diodes. The degree of attenuation is controlled by a forward bias current flowing in the diodes. Using this arrangement as the gain-control element in the experimental equipment, an almost linear control has been obtained over a range of approximately 60 dB. A small amount of swept gain, operative over the first portion of the sounding period, provides initial suppression of the

receiver. This is necessary when the echo-signal level is low and consequently the gain high, in order to suppress close-range reverberations including those from aeration. The swept gain and a.g.c. voltages are fed to the attenuator network in parallel and the greater of these two has the controlling effect.

The a.g.c. will adjust the overall gain of the receiver to keep the amplitude of the largest echo-signal constant. As already mentioned, other echo signals are occasionally larger than the sea-bed echo. Under these conditions, the a.g.c. is not helpful unless some further control is applied to it. This point is taken up again in the next paragraph.

2.3. Time-gating

Reliable selection of the sea-bed echo can only be achieved if some form of time-gating is used to lock-out signals which originate from short-range and mid-water targets. The output from the amplitude selection system is in the form of d.c. pulses 0·1 ms in duration. Logic circuits are used for all subsequent operations and a NAND circuit is employed as the time-gate.

If the approximate depth to be measured is known, the time-gate can be controlled by a pulse from a
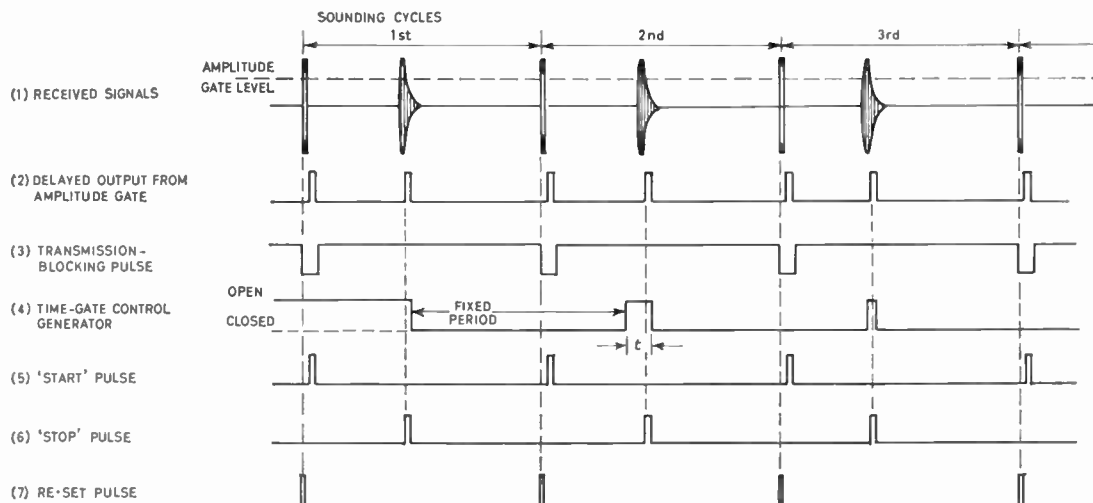
Fig. 4. Signal selection waveforms.

mono-stable circuit which is triggered at transmission. The gate is held closed for the duration of this pulse which is adjusted manually to suit variations in depth. Such a system requires constant attention by an operator and must either be accurately calibrated or provide an indication on the chart recorder of the pulse duration.[3]

An improved arrangement, which was originally developed for use in a sea-bed locked display for fishing,[4, 5] has been employed in the experimental equipment. The system, shown in Fig. 1, relies on the repetitive nature of the echo signals and on the fact that the change in depth from one sounding to the next will not be a significant percentage of the total cycle time. The time-gate is again controlled by the output of a mono-stable pulse generator which, when triggered, closes the gate for the duration of the pulse. This circuit, however, is only triggered by an output from the time-gate. The gate is therefore closed by this control waveform after the passage through it of a pulse from the amplitude gate. The duration of the time gating pulse is fixed at a little less than the periodic time of the sounding cycle. The action of the time gate is therefore to accept only one pulse in a sounding period and to bias the system heavily in favour of a signal from the same echo source in the next. Since the sea bed usually produces the most consistent echo, the time gate synchronizes with this signal and will remain so for as long as it appears regularly in each sounding, re-triggering the control generator each time. Figure 4 shows the signal-selection waveforms with the action of the time gate control generator on line 4. Line 3 shows the trans-mission blocking pulse which is applied as an additional

control to the time-gate. Although the gate will normally be closed at this time due to the control generator, it will not be so in the first sounding cycle. This situation will also exist if, due to a sudden loss of the echo-signal expected by the time gate, the control generator is not triggered during a sounding.

The time t, is the difference between the length of the time gating control pulse and the periodic time of the soundings. It must be sufficient to allow for the maximum change of depth expected between one sounding and the next. In soundings 1 and 2 of Fig. 4, the depth is unchanged but in the third a change of the maximum amount allowed for, is shown. In shallow water where the sounding pulse repetition frequency may be of the order 5 to 10 per second, changes in depth between soundings are unlikely to exceed 1 ft. Therefore the interval t can be 0·42ms. In practice the stability of the monostable time gate generator and the precision of the p.r.f., limit the minimum time to the order of 1–2 ms.

Very large mid-water reflectors sometimes cause masking of the sea-bed echo. This situation is aggra-vated by the a.g.c. system if the level of the interfering signal is greater than that of the sea-bed echo. Under these conditions false signal selection may occur. To minimize the effect of such forms of interference the time gate waveform can be used to provide an additional control to the a.g.c. suppression gate as shown in Fig. 2. In this way the amplifier gain is only adjusted by signals occurring whilst the time-gate is open. This arrangement greatly improves the operation providing the sea-bed echo continues to appear consistently (even if reduced in amplitude) when interference occurs.

2.4. *Start and Stop Pulses*

The object of the signal-selection system is to produce pulses corresponding to the transmission of the acoustic signal and the reception of the sea-bed echo. These may be termed 'start' and 'stop' pulses respectively and are used to control the operation of a digital counter.

The stop pulse is obtained from the output of the time gate, but the pulse corresponding to transmission is blocked in this channel. It is acquired by applying the output from the amplitude gate to a second NAND circuit in parallel with that used for the time-gate. As shown in Fig. 1, this gate is controlled by an inverted form of the waveform from the transmission blocking circuit. Thus it is only open during transmission and selects the appropriate pulse from the amplitude gate.

A reset pulse is also required by the digital system in advance of the start pulse. Both start and stop pulses are delayed slightly from the corresponding acoustic signals, therefore the reset pulse is conveniently obtained from another monostable device, triggered at transmission.

3. Signal Selection Check

The complete equipment includes a conventional recording echo-sounder which is used to check the signal selection process.[6] The method is to interrupt for a short time the recording on the chart of the echo which has been selected.

The stop pulse appearing at the output of the time gate is used to trigger a circuit which blocks the signals being fed to the paper-marking amplifier of the recorder for a short time. Since the stop pulse is delayed slightly from the actual echo, the effect on the chart is to produce a white gap in the echo recording following the leading edge. With repeated selection of the same signal in each sounding cycle these gaps

build up into a 'white line' drawn in the recorded echo just after the leading edge. An example of this is shown in Fig. 5.

Due to the action of the time gate which allows only one stop pulse to be produced in a sounding cycle, the white gap can only be placed on one recorded echo in any one cycle. Therefore, if after a period of correct selection a failure takes place it is clearly shown on the chart by an interruption of the white line at that point.

Correct operation of the equipment is thus indicated on the recorder chart by an unbroken 'white line' following the leading edge of the sea-bed echo. A glance at the chart is all that is necessary to check the operation of the signal selection circuits. Time marks added to the chart may be used for correlation with a 'print-out' of the depth data. In this way, a simple and continuous check on the operation of the signal selection is obtained. Also, since a permanent record is produced, the checking operation may take place at any convenient time after the results have been obtained, and then only in regions of doubt.

4. Digital System

The function of the digital system is to produce depth readings by measuring the time interval between the start and stop pulses. Count pulses are applied to a counter via a gate which allows them to pass during the period between start and stop pulses. In this way the reading in the counter represents the depth in units determined by the periodic time of these pulses.

4.1. *Generation of Count Pulses*

The arrangement used to generate the count pulses will vary according to the accuracy required of the readings, and the type of work involved. The experimental equipment uses an LC oscillator which may be switched to provide depth readings in either feet or

Fig. 5.

Signal selection check using the recorder chart to identify the selected echo.
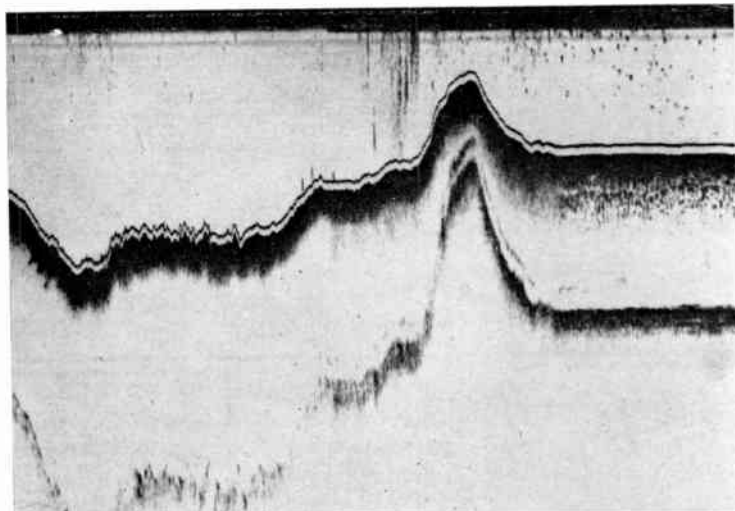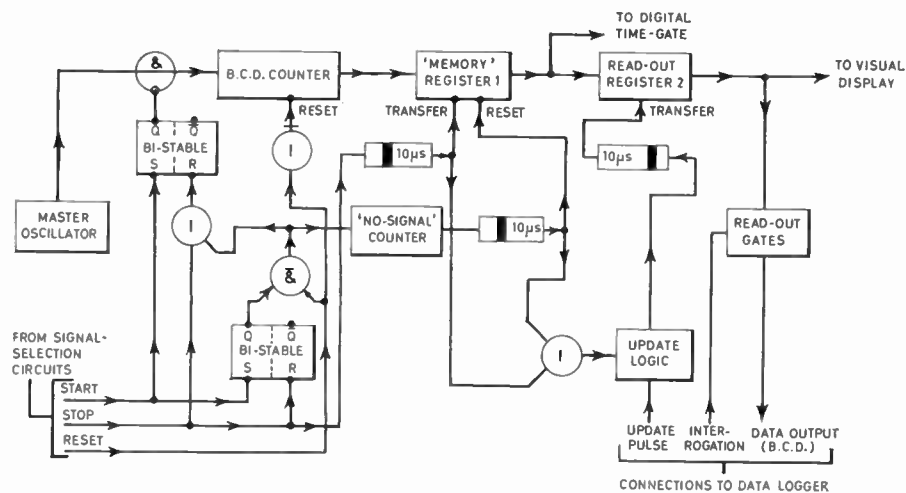
Fig. 6. Digital measurement system.

metres. An adjustment is also provided to allow for departures from the figure for the speed of sound in water which is normally taken at 4920 ft/s (1500 m/s). However, such a circuit is not ideal since two schools of thought exist on the requirements for this part of the system. One favours the use of a fixed frequency for the count pulses with adjustments to allow for sounding velocity variations being made to the readings at a later time. With this arrangement a crystal oscillator can be used to generate the count pulses thus providing a basis of measurement with much greater accuracy than is normally possible with a chart recorder.

The second approach is to calibrate the equipment at intervals during use. With chart recorders this is usually carried out by suspending a target under the transducers at a known distance. The stylus velocity of the recorder is then adjusted to produce the echo signal from this target at the correct depth reading on the chart. For digital read-out the count pulses may be derived from a resolver connected to the stylus driving system. The frequency of the pulses will then vary with the stylus velocity thus calibrating both the recorder and the digital read-out simultaneously.

4.2 Digital Measurement System

The digital system used for most of the experimental work consisted of a three decade, binary coded decimal (b.c.d.) counter. The readings from this counter were made available in parallel form, one decade at a time to a data logging system. The counter reading was held until a reset pulse was applied from the data logger. A new reading was then produced from the next complete sounding cycle which followed. Hence, although the signal-selection system

produced start and stop pulses for the counter from each sounding cycle, a digital reading was only produced when required by the output system. It was found from extensive trials of the equipment that the most common cause of failure was intermittent or partial aeration of the transducers. This caused a sudden loss of signal for one or two cycles of sounding; consequently no stop pulse was produced at this time. The echo-sounder chart was still clearly readable under these conditions but the digital readings contained occasional errors due to a bad sounding. Since the loss of signal usually lasted for only one or two soundings, correct signal selection took place very close to the desired reading but failed on the actual cycle of measurement. The system was therefore wasteful of information.

An improved system is at present in use, and the arrangement is shown in Fig. 6. It consists of a b.c.d. counter as before, followed by two registers. The counter produces a digital reading for each sounding cycle. It is controlled by the start, stop and reset pulses obtained from the signal-selection circuits. The registers are fed in parallel form. Register No. 1 receives its information from the counter and register No. 2 from register No. 1. A common clock pulse to all 'bits' in a register is used to effect the transfer of information into it.

Readings from the counter are transferred to register No. 1 a short time after the stop pulse has caused the counting action to cease. Register No. 1 therefore provides a memory for the latest available reading until a new count is completed. If no stop pulse occurs during a sounding, transfer to the register does not take place and the old reading is retained. The reset pulse is now used to stop the counting action

(ready for the next sounding cycle) as well as resetting the counter. This situation is allowed to continue for a few soundings by arranging a circuit to count the number of successive soundings which have passed without a stop pulse being produced. When this count reaches a predetermined figure, a pulse is produced which resets register No. 1 to zero, thus indicating that no reliable information is available. The number of sounding cycles for which an old reading remains useful information depends upon the type of sounding work involved. For shallow-water high p.r.f. sounding, a figure of five would appear to be acceptable.

Register No. 2 is used to feed the depth information to the read-out devices. It is up-dated from register No. 1 when a new reading is required. An interlock circuit between the transfer pulse lines of the two registers ensures that ambiguous readings do not arise due to both registers up-dating together.

5. Read-out

The depth information stored in register No. 2 is available for interrogation by external equipment by means of output gates. The decades may be read one at a time and the binary coded decimal information appears at the output on four lines in parallel form. A visual display is incorporated in order to monitor the readings.

6. Digital Checking

A check on the signal selection has been described in Sect. 3 which involves the use of the recording echo-sounder and indicates the selected echo on the chart in the form of a 'white line' in the echo tail. This system is simple to read and produce but in certain circumstances it can only indicate that the digital readings are suspect but not definitely wrong. For instance in shallow water work, the sounding rate may be up to fifty times higher than the print-out rate, therefore, occasional failures of the signal selection may not reach the digital output. It is then useful to indicate on the echo-sounder chart the digital reading which is present in register No. 2 for printing and display purposes.[6] To produce this check, a second counter is needed which is fed continuously with the same pulses that are used to produce the depth readings. The state of this circuit is compared continuously with the reading in register No. 2. When these readings coincide, a pulse is produced which is amplified and used to make a small black mark on the echo-sounder chart. If the check counter is reset to zero at the commencement of a sounding, the coincidence pulse will indicate the digital reading against the recorder scale. Since the actual sea-bed recording will be in this position, the checking signal must be offset in range so that it may

be readily seen on the chart and this can be achieved by pre-setting the check counter to a fixed value instead of re-setting to zero.

The check signal will appear in the same place on the chart for as long as the digital output remains the same. Therefore as the readings change, a staircase effect is produced which follows the sea-bed recording but spaced a little before it.

7. Digital Time-gate

The sea-bed locked time-gate control described in Sect. 2.3 is very effective in use but only provides 'ping-to-ping' gating. A sudden loss of the expected signal will mean that no time-gating is applied during the next sounding and the system becomes liable to select any large close range signals. In these conditions a time-gate system which will persist for a short time in the absence of a signal is desirable.

In Fig. 7, a digital time-gate system in which the time-gate is controlled by a J-K flip-flop, is shown. The basis of the system is a b.c.d. digital counter and a coincidence gate. The counter is fed with pulses from the master oscillator (see Fig. 6) and its output is fed to the coincidence gate together with the output from register No. 1. This register stores the depth reading from the last sounding while a new one is being produced. When the reading in the time-gate counter is the same as that present in register No. 1, a pulse is produced from the output of the coincidence gate which is used to switch the J-K flip-flop to the gate 'open' condition. In order for the time-gate to be opened a short time before the arrival of the expected echo signal, the time-gate counter is preset by a small amount at the commencement of each sounding. The value of preset applied can be made easily adjustable and therefore provides a control over the time-gate 'open' period.
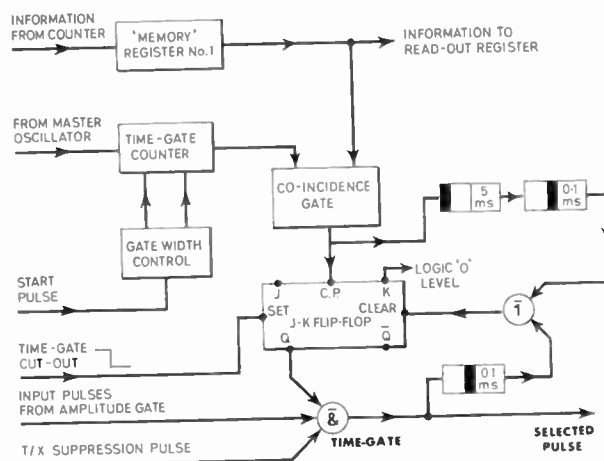
Fig. 7. Digital time-gate system.

The J-K flip-flop is switched to the gate 'closed' condition by the arrival of an echo-signal pulse from the amplitude-gate. If, however, this signal does not appear the gate may be closed after a period of a few milliseconds has elapsed from the moment of opening. A monostable device triggered by the output pulse from the coincidence gate is used to produce this effect. The time-gate will now continue to 'open' and 'close' at the same position in each sounding for as long as the reading in register No. 1 remains. If, due to the continued absence of the required echo-signal, register No. 1 is returned to zero, the time-gate control is cut out in order to allow the system to re-locate the sea-bed echo.

8. Conclusions

Extensive trials of the experimental equipment have been carried out under normal survey weather conditions. The equipment used the simple digital system described in the first part of Sect. 4.2. An analysis of the results showed that 97.7% of the logged readings were correct. The failure rate was investigated further, and of the total 2·3%, 0·55% proved to be due to selection of an incorrect echo, and 1·75% resulted from no echo of sufficient size being received. In an exercise designed to demonstrate the reliability of the system, certain areas of the sea-bed profile were reconstructed from the logged depth readings, on the same scale as the recorder chart. The two were then compared and Fig. 8 shows an example of the result. In the figure a photographic process has been used to produce a transparent overlay of the reconstructed profile (shown by the thin black line) which is seen to follow closely the outline of the recorded sea-bed echo. Toward the left-hand side of the figure a slight discrepancy in the two outlines has arisen due to difficulty in exactly reproducing the scale of the recorder chart. The depth readings have been logged at five second intervals.

An improved version of the experimental equipment has been designed and is at present in course of construction. This will form the basis of equipment which should prove to be of great value.

9. Acknowledgments

A considerable part of the trials was carried out in Holland with the co-operation of the Rijkswaterstaat who produced the original of Fig. 8 using their computer facilities at The Hague. The depth readings during these trials were logged by equipment, supplied
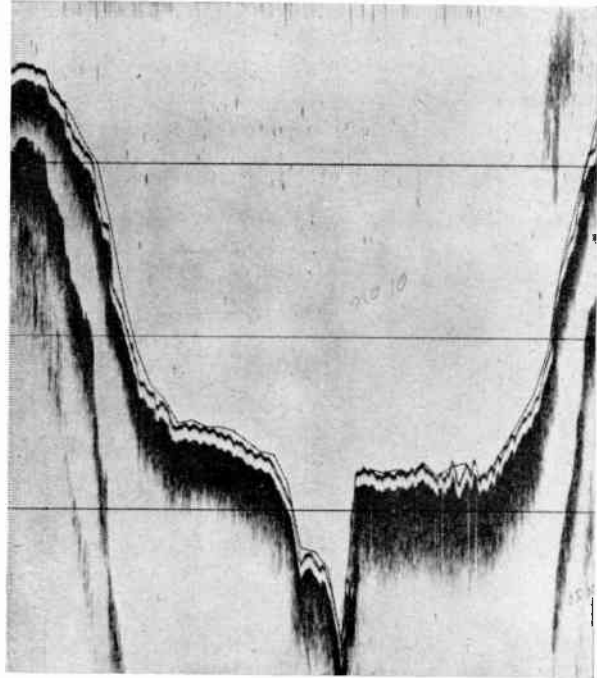
Fig. 8. A comparison between the chart recording of the echo signals and a reconstruction of the sea-bed profile from the logged depth readings.

by Decca Navigator Co. Ltd., which also recorded radio navigation fixes and time.

The author would like to thank the directors of Kelvin Hughes, a division of Smiths Industries Ltd., for permission to publish this paper and also to Mr. W. Halliday, Mr. G. Stow, Dr. R. W. G. Haslett and Mr. P. R. Hopkin for their valuable help and advice during its preparation, and throughout the project.

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Simultaneous Use of Sideways-looking Sonar, Strata Recorder and Echo Sounder

By

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AND

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Summary: In geological survey, the simultaneous operation of these three equipments on one vessel can have considerable merit.

The sideways-looking sonar (known as the 'Transit Sonar') has a range of 600 yd and gives an approximately true plan of the sea bed on dry paper 6 in. wide when the ship moves at 5·7 (or 2·85) knots over the ground. The record produced by the 'Transit Sonar' may be readily correlated with the corresponding Admiralty chart.

The Strata Recorder operates at 9·6 kHz, at which frequency the acoustic waves penetrate the sea bed. Good resolution is obtained on dry paper due to the use of short pulses and relatively narrow acoustic beams. A special circuit gives time-varied gain, triggered by the sea-bed echo, to employ the dynamic range of the paper in the optimum way.

The precision echo sounder records on dry paper 10 in. wide. The capacitor-discharge transmissions at 32 kHz are switched by a silicon controlled rectifier.

The position of the survey vessel at any instant may be obtained from a radio position-fixing system or by visual fixes.

Some examples of interesting sea-bed features are portrayed in the figures, each a combination of the simultaneous records correlated in time on the same scale, together with their geological and topographical interpretations. Sand waves of wavelength 13 ft (4 m) and as little as 9 in. (23 cm) high have been resolved quite readily.

1. Introduction

Recent advances in the designs of instruments for geological survey give improved results. In this paper three new instruments are considered:

(a) The sideways-looking sonar known as the 'Transit Sonar',

(b) the hydrographic echo sounder type MS. 36 and

(c) the Strata Recorder.

Each of these has advantages over existing types, but the use of these three instruments simultaneously on one vessel greatly enhances the value of the survey, since the three records can be correlated and interpretations made from one record are confirmed by the other records.

2. The Transit Sonar

This instrument has been fully described in the associated paper.[1]

In the figures given later in the present paper, the straight-line recorder has a maximum range-scale of 600 yd and an approximately true plan of the sea bed is portrayed when the vessel steams along a straight track at a constant speed of 5·7 knots over the ground. The record is made on dry paper 6 in. wide and the pulse length (1 ms) corresponds to a range resolution of 2·5 ft.

3. The Hydrographic Echo Sounder (Type MS. 36)

The new features in this echo sounder are the high stylus speed and the increased width of the record so that ranges up to 70 ft can be observed on an open scale without change of 'phase'.‡ Also the scale range and 'phase' in use are continuously indicated on the record, whilst the topography of the sea bed may be examined in detail using the high paper-speed.

This instrument which is made to a tropical specification, may be divided into three parts:

(i) electronic unit,

(ii) recorder and

(iii) transducers.

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‡ 'Phase' in this context refers to the advance of transmission with respect to the record to enable a partial range to be examined.
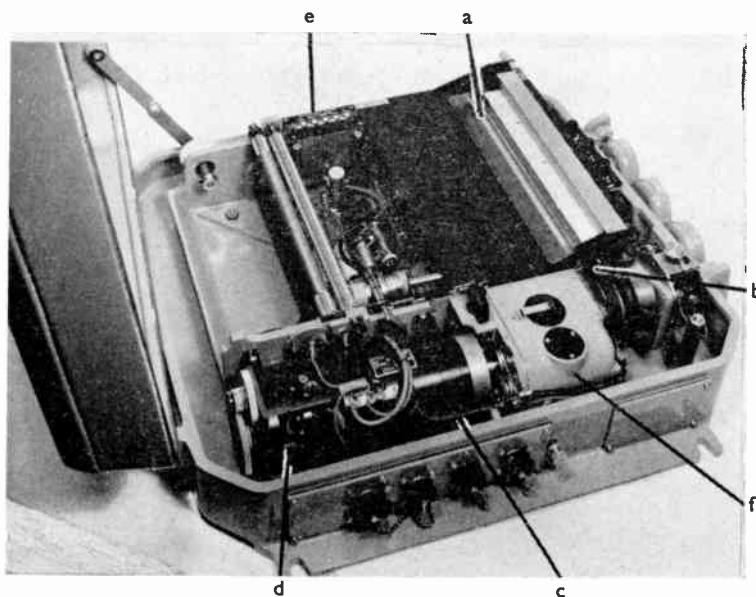
Fig. 1. The MS. 36 echo-sounder recorder (with lid raised and part of the platen removed).

a range scale	d governor
b belt	e tachometer
c driving motor	f gearbox

3.1. *Electronic Unit*

This comprises the transmitter, the receiver and the power unit.

The transmission is obtained from a capacitor-discharge circuit in which the capacitor is charged to 300 V and discharged through the transducer by a silicon controlled rectifier, triggered by the transmission contacts in the recorder. The pulse has a peak power of about 400 W. The transmission rate (on the 70-ft range-scale) is 9·8 pulses/second.

The receiver embodies three valves of which the first two operate in high-gain tuned amplifier stages and the third supplies the electrical power to the stylus for marking the record. This valve consists of a double triode operating in two distinct circuits. After rectification, the signals pass to one triode which acts as an output stage. The other triode gives d.c. coded signals for recording information regarding the scale-range and 'phase' in use, also the time-interval and fix marks. All these signals are applied to the stylus. The fix marks may be operated by a foot switch.

When the supply from the vessel is d.c., a solid-state inverter delivers a stabilized a.c. voltage to the echo sounder. The overall power required is 87 watts.

3.2. *The Recorder* (Fig. 1)

This straight-line recorder has an expanded scale (a) in that ranges up to 70 ft are recorded across dry paper 10 in. wide. The record may be read to within 2 in. on the scale. The general level of marking density may be adjusted to suit conditions.

The sprung stylus is mounted on a toothed rubber belt (b) driven by a d.c. motor (c) with an electrical centrifugal governor (d) adjustable against a built-in tachometer (e) to sound velocity calibrations between 4620 and 5160 ft/s. The gearbox (f) between the motor and the toothed pulley driving the belt, has an alternative ratio giving a scale-range of 140 ft.

Such a high-speed stylus requires special mechanical design to prevent bounce. The writing end of the stylus is constrained by a guide (whilst crossing the paper) to move along a straight path. The other end is in contact with a bus-bar which is connected to the output of the receiver. Since the stylus moves from top to bottom of the record, the profile of the sea bed is correctly orientated in relation to the water surface.

Other gear-trains rotate two printed-circuit contact disks and a commutator, as well as drive the paper (at a choice of three speeds: ½, 1 and 4 in/min). Two other contact-drums attached to the pulleys at either

end of the belt, are concerned with the transmission contact and give changes of 'phase' by $+50$ or $+100$ ft on the 70-ft scale[2] (or by $+100$ or $+200$ ft on the 140-ft scale), also a draught adjustment between 0 and 10 ft.

One of the printed-circuit disks causes a mark to be made at the top of the record every minute. The other disk allows two parallel bands to appear at the foot of the record, interrupted by a dot and dash coding to indicate the range-scale and 'phase' in use. These bands appear all the time. An alternating voltage generated by the commutator is applied to the tachometer which consists of ten resonant reeds tuned to the frequencies corresponding to the correct motor speeds for the sound-velocity calibrations engraved on the tachometer. In this way, the recording accuracy is $\pm 1\%$ in 340 ft.

The recorder case is watertight.

3.3. Transducers

The type of transducer depends on the installation, whether outboard or inboard.

In the outboard arrangement operating at 32 kHz, the two magnetostrictive transducers are mounted in a streamlined fibre-glass housing (taking the form of a 'shoe') on a support tube 8 ft long attached to the gunwale. Each transducer consists of a toroidal winding on a permanently-magnetized scroll of Permendur with a conical reflector of expanded neoprene giving a beam angle of 13 deg (between 3 dB points). Since the mechanical Q is about 5, a short pulse is obtained. The base of the housing consists of a fibre-glass acoustic window and the region between the reflector and the window is filled with water. The remainder of the housing is filled with polyurethane foam. This equipment is robust, light in weight and portable from ship to ship. (The records shown later in this paper were made with this outboard arrangement operating at 32 kHz.)

In the case of the inboard transducers mounted in separate cylindrical housings, the choice of frequency depends on the thickness of the hull of the vessel or of the diaphragm which is permitted. If the thickness of the hull plating does not exceed $\frac{1}{8}$ in. or if a diaphragm as thin as this is permissible, 32 kHz can be used and the reflectors are 8 in. in diameter. Alternatively, for sounding through thicker plate, 14 kHz is recommended with a reflector diameter of 14 in. The beam angles are approximately the same as in the outboard version.

4. The Strata Recorder

The new features of this equipment are the use of relatively narrow acoustic beams at a low frequency, the shortness of the pulse, the expansion of the record and the triggering of the receiver swept-gain by the sea-bed echo[3,4] which result in good acoustic resolution and the optimum use of the dynamic range of the recording paper. In this way, layers of slightly differing acoustic properties can be resolved close to the surface of the sea bed.

Two compromises had to be made in selecting the operating frequency.

First, the beam angle was chosen to give the highest angular resolution without needing to stabilize the transducer against rolling and pitching of vessel. A 3 dB beamwidth between 15 and 20 deg was decided upon.
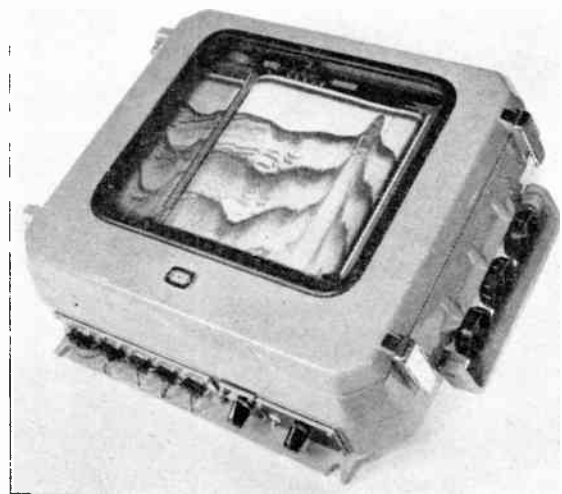
Fig. 2. The Strata Recorder—the case (21 in × 23 in) containing the mechanism and the triggered swept-gain circuit.

Secondly, for this beam angle, there was a compromise between frequency and size of transducer. Whilst a low frequency is required for penetration of the sea bed, this would need a large transducer which would be inconvenient for handling and transferring from one ship to another. Also for a given Q, the pulse length increases when the frequency is lowered so that resolution in range would be impaired. A frequency of about 10 kHz was selected (using a transducer of the lowest available Q), as experience indicated that this frequency would give adequate penetration.

4.1. The Recorder Mechanism (Fig. 2)

This is based on the MS. 36 recorder except that there is only one range scale (140 ft) with 50-ft phasing steps to a maximum of 340 ft (in water). The transmission contact on the recorder closes 30 ft before the zero on the range scale for two reasons: to allow for the delay in the transmitter and to obtain
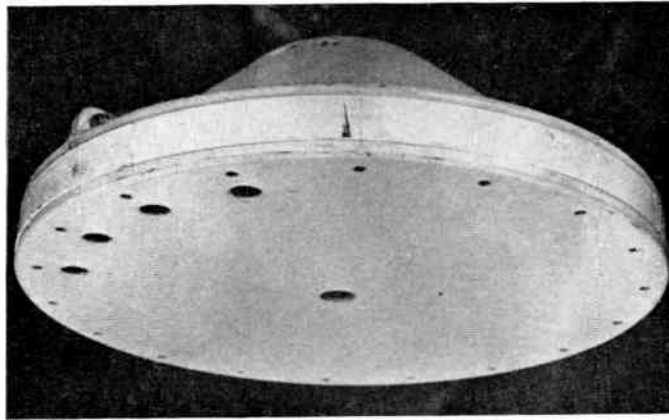
Fig. 3. One of the pair of transducers for the Strata Recorder, showing the conical reflector (23 in. diameter) and fibre-glass acoustic window.
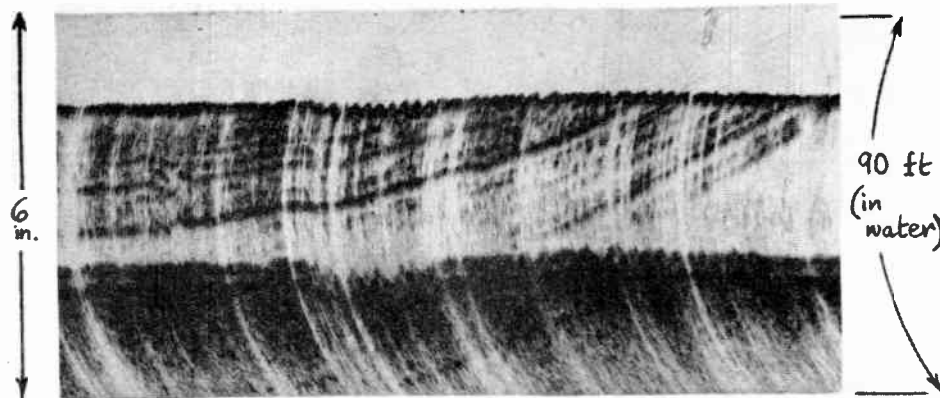
Fig. 4. A recording of strata rising to the surface of the sea bed off Yarmouth, Isle of Wight (taken on an experimental model).
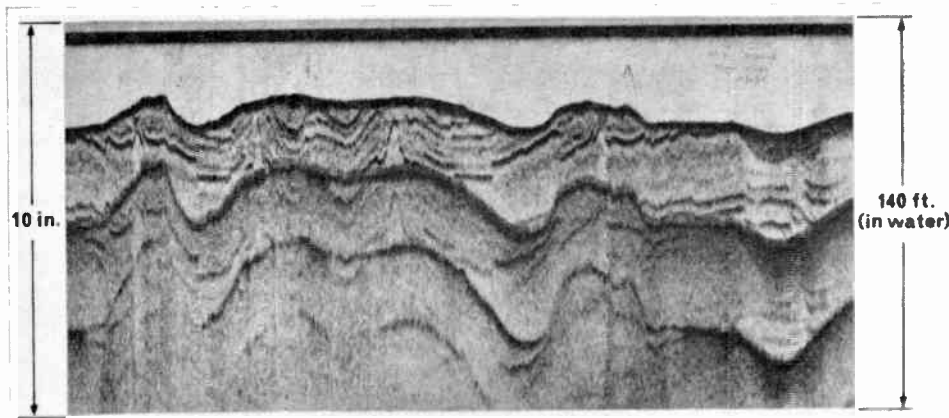
Fig. 5. A recording of strata in the bed of the river Crouch at Burnham showing faults, taken by the latest straight-line recorder. (Multiple echoes from sea bed and strata are seen.)

prior warning of the transmission so that it can be eliminated from the receiver.

There is a choice of paper speeds: $\frac{1}{4}$, 1 or 4 in/min. On the highest speed, detailed examination of the topography of the sea bed is possible within the limits set by the width of the acoustic beams and the motion of the surveying vessel.

4.2. Trigger and Delay Unit

Also in the same case as the recorder is the trigger and delay unit, using six transistors in three monostable multivibrator circuits.

The first circuit is switched on by the transmission contact (and recovers 70 ms later). The leading edge of the resulting pulse operates the transmission 'contactor' relay.

The second multivibrator is also switched on by the transmission contact but recovers 16 ms (or 40 ft) later, that is about 10 ft after transmission. The resulting negative going pulse (No. 2) closes the gate in the side-chain receiver (see Section 4.5 below) until the short-range reverberations have died away.

The third multivibrator is triggered at the end of gating pulse No. 2 (that is 10 ft after transmission) and recovers 136 ms later (equivalent to 340 ft). The resulting positive going pulse (No. 3) keeps the gate in the side-chain receiver open until just before the next transmission.

4.3. The Transmitter

This is in a separate case. A capacitor is charged to 1600 V and is discharged through the transducer by the 'contactor' relay. Due to unavoidable inertia in this relay, there is a delay of about 12 ms (equivalent to 30 ft) between the closing of the transmission contact in the recorder and the discharge of the capacitor. The peak power in the pulse is about 1 kW and the pulse length is 0·5 ms (equivalent to a range resolution of 1·25 ft in water).

4.4. Transducers (Fig. 3)

Two are used, one for transmission and the other for reception, each having a beam angle of 17·5°. Toroidally-wound Permendur scrolls were chosen for the transducer elements since these give a Q as low as 5. Each transducer is mounted in a conical reflector of expanded Onazote, embedded in a fibre-glass housing. The centre of the scroll is partly filled with expanded Onazote leaving a cavity resonator of depth $\lambda/2$ in water.[5] This has the effect of reducing the amplitude of the first side-lobe to 10% of that of the main lobe. At the base of the housing is an acoustic window, also made of fibre-glass, the region between the reflector and the window being free-flooding.

These transducers are hung over the side of the vessel on tubular poles mounted on outboard rigs. The operating frequency is 9·6 kHz.

4.5. The Receiver

This is contained in a separate case and, in fact, consists of two receiving amplifiers, each having the input connected across the receiving transducer

One amplifier is the main receiver using four valves which give a high gain. The first two are pentodes employed in tuned amplifier stages. The signal at 9·6 kHz is then rectified and applied to the output stage supplying the stylus in the usual way.

The second amplifier is a side-chain receiver used for sea-bed selection and gating.[4] In the first stage (a pentode), the received signals (consisting of transmission breakthrough, short-range reverberations and sea-bed echo) are amplified. These signals pass through a transistor gate, controlled by gating pulses Nos. 2 and 3 (as described in Section 4.2 above) which rejects the transmission and reverberations but allows later signals to pass. The first of these signals is the (first-time) sea-bed echo which passes through an amplitude-gate and triggers a thyratron. The resulting negative-going edge is amplified and inverted by a pentode stage to give a positive-going sea-bed trigger pulse.

Normally, the first stage in the main receiver is cut off by a standing negative bias. However, a diode gate is switched by the sea-bed trigger pulse and causes the bias to decay away at a rate controlled by an adjustable capacitor-resistor circuit. In this way, the gain of the main receiver rises at a rate which may be adjusted to suit the conditions and finally attains a value determined by the normal gain control.

The total power required by the Strata Recorder is 220 watts at 24 V d.c.

4.6. Strata Records

Two examples of typical records are seen in Figs. 4 and 5.

In Fig. 4 (taken on the experimental recorder having a rotating arm) the transmission is off the top of the record and the first sea-bed echo is seen with the second sea-bed echo below. (The second echo occurs because the returning pulse is reflected by the water surface and by the sea bed a second time.) Numerous strata are seen rising to the surface of the sea bed on the right-hand side of the figure. Some of the strata have a thickness equivalent to 1·2 ft in water.

The interpretation of thickness naturally depends on the velocity of sound in the material of the sea bed which may change from one stratum to another. Also, the apparent thickness of the stratum increases at greater ranges, due to spreading of the beam.
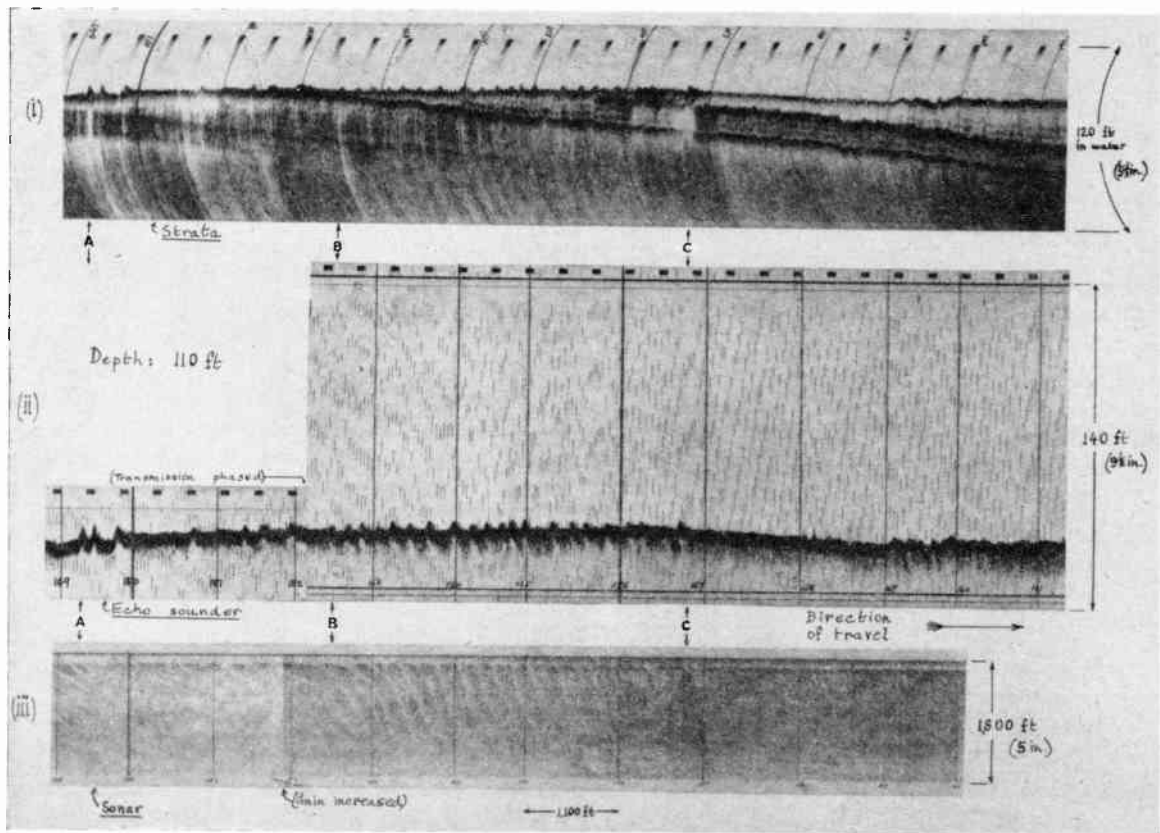
Fig. 6.

Figs. 6-7. Some typical records taken simultaneously on the
Strata Recorder, the hydrographic echo-sounder and the Transit
Sonar in the North Sea.

Hamilton *et al.*[6] have made measurements (at
100 kHz) on the acoustic properties of silt and sand
and found that the velocities of sound in these
materials (respectively) vary from −1·9% to +16%
as compared with sea water. (In harder materials,
the velocity of sound can rise as high as four times
that in water.[7]) This means that, in general, the
difference in range between two strata is likely to be
greater than that recorded on the range scale which
is calibrated in terms of water.

The record seen in Fig. 5 was taken on the latest
(straight-line) recorder and shows the transmission
with first, second, third and fourth sea-bed echoes
successively below. An interesting series of strata are
indicated, with a number of faults. (Multiple echoes
from the strata are also seen.)

5. Records Taken Simultaneously

A survey was made recently of parts of the North
Sea with three equipments running simultaneously
on one vessel, namely the Transit Sonar, the Strata
Recorder and the hydrographic echo-sounder.

Figures 6–8 show some of the results. (In Figs. 6
and 7, the upper record (i) being taken from the
experimental model of the Strata Recorder, has a
curved range co-ordinate on paper 6 in. wide, whereas
the current model gives a straight-line record on
paper 10 in. wide as in Fig. 5.) The position of the
vessel was obtained from the Decca Navigator and
visual fixes.

The same numbered position fix-marks were made
on each record but since the paper speeds were
different, photographic reduction in size is required
to make the horizontal time-scales coincide. The
scale ranges in both vertical and horizontal directions
are marked on the figures (as well as the actual sizes
of the records). The ship's speed was about 5·7 knots
and the weather was calm.

On careful examination of the three records at
corresponding points in time (that is on the same
vertical line), all the detailed structure is found to
correlate, bearing in mind the fact that the ship's
track is at the top of each sonar record (iii). Also
the information concerning sea-bed features seen in
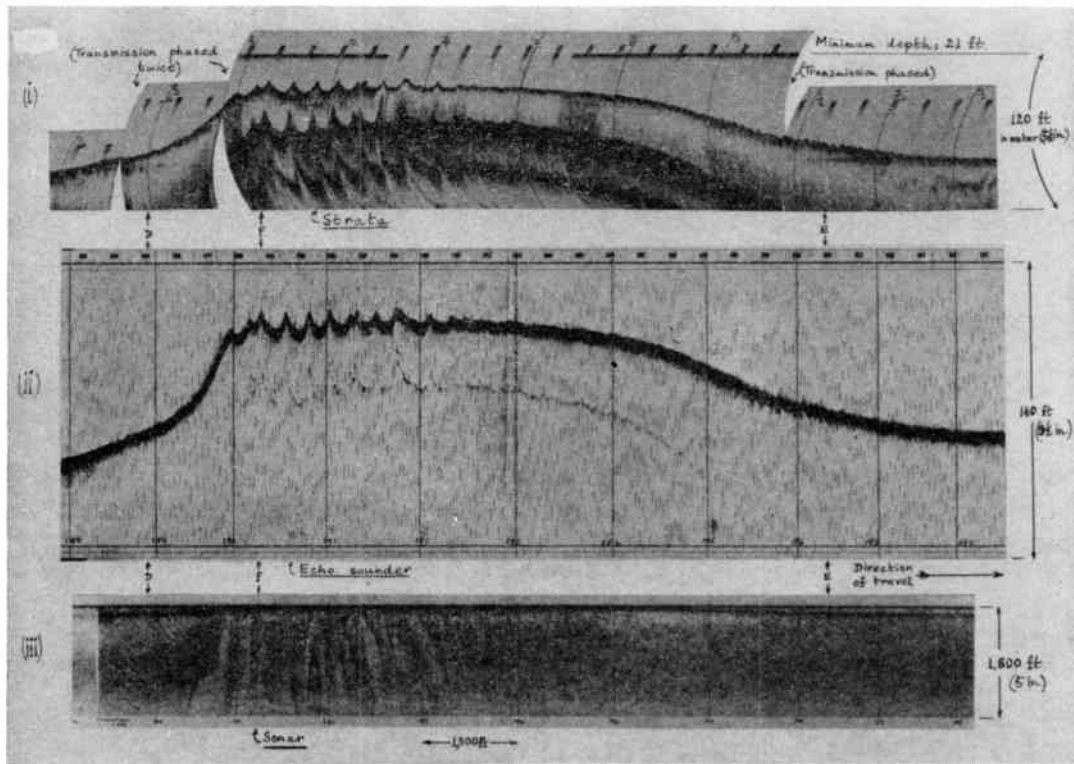
Fig. 7.

profile on the echo-sounder is augmented by the shape of the same feature in plan on the sonar record. In other cases, some of the crests on the sea bed observed on the echo-sounder are found to result from strata rising to the surface of the sea bed. From the records, the heights and wavelengths of the sand waves can be measured as well as their orientation to the ship's track.

Numerous interesting features are found. For example, in Fig. 6(i), acoustic shadows occur behind hillocks on the sea bed at points A, B and C. This may be because the convex surface of the sea bed may act rather like a diverging lens in the case of light. (The velocity in the sea bed is higher than in water, whereas the velocity of light in glass is less than in air.) The sound waves penetrate the sea bed to a depth of at least 50 ft (equivalent in water).

At points A and B in (ii) and (iii), short sand waves are seen superimposed on long ones. In the area of (iii) between B and C, the beam looks down the 'valleys' between the long sand waves whereas at point A, the angle between these waves and the track of the ship is more acute and the wave height larger, so that the acoustic shadows are more sharply defined. In general, the waves seen in (iii) appear to correlate with the successive strata breaking the surface of the sea bed in (i).

A large bank with sand waves on top is seen in Fig. 7, the crest rising to within 21 ft of the water surface. (Second and third sea-bed echoes also appear in (i). It may be noted that the level stratum at points D and E in (i) appears to run straight across under the bank. At point D in (iii), sand waves are recorded (on the left-hand slope of the bank) which, due to their inclination to the ship's track, are not easily seen on records (i) and (ii), whereas at F on top of the bank, the wavefronts are perpendicular to the ship's track and are clearly shown on all three records. From the shape of the bank, the predominating tidal flow is from right to left on the record. The sides of the bank D and E are relatively steep. (The echo-sounder records (ii) exaggerate the slopes about 35 times.)

Figure 8 (ii) shows detailed fine structure. At L are sand waves parallel to the track of the vessel, with ribbons of sand over shingle (M) and a sandbank at a range of 330 yd on the starboard side of the vessel. N is an example of a large ridge of sand with small sand-waves on top. This part of the record (ii) has been enlarged five times to show the high degree of detail given by the Transit Sonar. Some of the sand waves have a wavelength of only 23 ft. The ridges at O are possibly composed of chalk. Additional records were given in the original paper.[8]
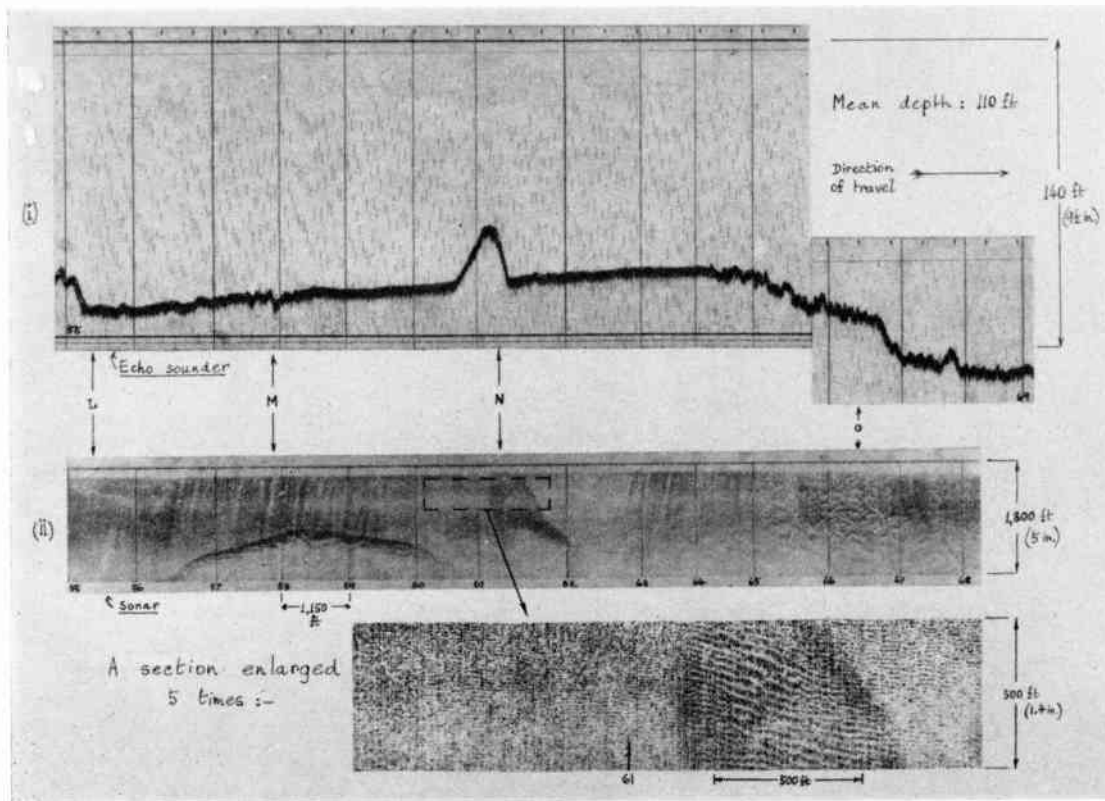
Fig. 8. Some typical records taken simultaneously on the hydro-
graphic echo sounder and the Transit Sonar in the North Sea.

The examples which have been given, show that the complex topography of the sea bed can be deduced from the sonar record with its immense detail and clearly defined shadows, corroborated by the echo-sounder record which gives the dimensions. The Strata Recorder provided additional information regarding the structure of the sea bed.

6. Acknowledgments

The authors are grateful to the Gas Council and N.A.M. (Gas Export) for permission to publish Figs. 6–8. A number of colleagues took part in the development of the equipment described in this paper which is published with the consent of the Directors of Kelvin Hughes, a Division of Smiths Industries Limited.

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Television Wired Broadcasting Systems

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Presented at a meeting of the Institution's Television Group held in London on 28th September 1966.

Summary: The paper outlines the growth of television wired broadcasting in the United Kingdom. After considering the fundamental problems involved in the use of video h.f. and v.h.f. frequencies, it describes the way these problems have been solved for h.f. and v.h.f. systems. Brief descriptions of practical h.f. and v.h.f. systems are given and this is followed by a comparison of the relative costs of these two systems. It is concluded that, if a large section of a town is to be wired, the h.f. system is the most economical overall. Blocks of flats in a town that has no h.f. distribution system can most economically be dealt with by a v.h.f. system. Various borderline cases in between the two extremes are dealt with.

The precautions necessary for distributing colour television are mentioned, and suggestions made for appropriate tolerances to which wired broadcasting operators should work. Finally, the author gives some indication of the way in which wired broadcasting may develop in the future.

1. Introduction

So as to set some bounds to this rather large subject, it is proposed to limit the present discussion to the distribution of television signals to private homes mainly for entertainment purposes and not to deal with point-to-point distribution (except insofar as this is necessary for wired broadcasting) or to deal with the field of educational or industrial television.

The growth of wired broadcasting of television in the U.K. is illustrated in Fig. 1 from which it can be seen that the total number of subscribers now receiving their television signals by wire is nearly one million.[1] Whilst there are many more v.h.f. than h.f. systems, the number of subscribers receiving their signals by h.f. is in fact more than twice as many as those receiving their signals by v.h.f.

It is important to make it clear why both these types of systems have been developed, and to consider the circumstances which lead to the choice of one type of system or the other.

Before this can be done, however, it will be necessary to explain why distribution at video frequencies is not regarded as a practical proposition over large areas and then describe briefly the main features of the two types of systems which are in use, i.e. the h.f. system on multi-pair cable, and the v.h.f. system on coaxial cable. The main factors which influence the choice of one or other of these two systems will then be discussed. Finally, colour television and some aspects bearing on the future of wired broadcasting will be mentioned.

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2. Wired Broadcasting Using Video Frequencies

No viable video system has ever been developed. It has the theoretical advantage of requiring the lowest bandwidth and lowest maximum frequency and thus the lowest attenuation along the network and would eliminate the need for a demodulator in the receiver. Nothing more can be said in favour of such a system. The following are its chief drawbacks:

(*a*) The ratio between the highest and lowest frequency transmitted is so great that correcting the frequency response becomes a problem,
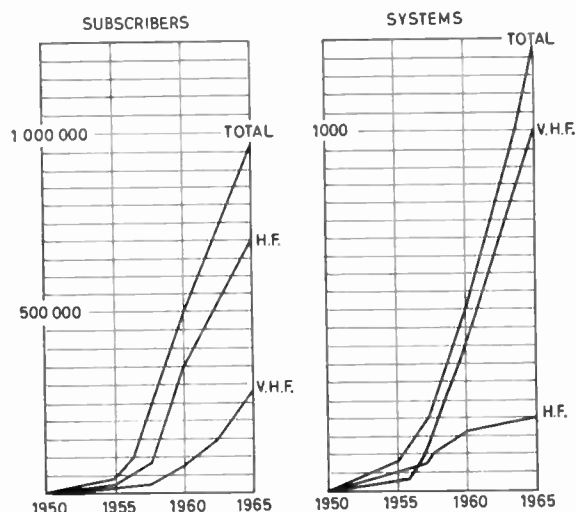
Fig. 1. Growth of television distribution in U.K.

(*b*) the velocity of propagation on cable varies much more at video frequencies than it does over the band (say) 5–10 MHz,

(*c*) the design of transformers for matching spurs into main stems would have been very difficult in 1950, and would not be particularly easy even today with modern materials,

(*d*) the network would have to be protected from the very much higher field strengths which obtain in the medium and long wave band, as compared with levels of field strength most commonly met with in the h.f. band,

(*e*) any pick-up of 50 Hz from the mains would produce hum on the picture,

(*f*) sound could not be distributed on the same pair as vision.

It has not therefore been worth struggling with these difficulties for the sake of somewhat minor advantages.

3. H.F. Systems

3.1. *Principles*

The main design philosophy was not how to design a method of connecting an aerial to a large number of standard aerial receivers in the cheapest possible way, but rather how to design a system as a whole, including special receivers, which would show the greatest overall economy. Since the receiver was likely to be the most expensive single item, the greatest simplification had to start at this point.

It is obvious that the most economical way of distributing sound programmes is to devote one pair of wires to each programme and to deliver sufficient power to each home to operate a loudspeaker direct without the need for an amplifier. This means that it is only necessary to provide each home with a programme selector switch, volume control, a step-down transformer and loudspeaker. This is a method of operation which has been well established for some thirty years. The question therefore in the wired broadcasting industry when television started again after World War II was 'could these same pairs of wires be used to carry the television signals as well, so that the same selector switch in the home is used to select television programmes complete with their associated sound?' This seemed very unlikely, and indeed proved to be so. The first reason was that some of the programme pairs consisted of open wires spaced about two inches apart. The attenuation of these was certainly low in the band 5–10 MHz but they were very vulnerable to interference and radiation. In 1950, however, a large proportion of open wire networks were being replaced by star-quad networks. These cables had considerably better performance as regards radiation and interference than spaced

wires and, using polythene as an insulant, had a comparatively low attenuation in the band 5–10 MHz when clean and dry, but unfortunately as these became dirty and wet the attenuation rose sometimes by as much as five or six times. This made their use quite out of the question for any length of more than a few yards. (Additional information on interference, radiation and the effect of dirt has been published by the author.[17])

The basic objective, however, seemed worth pursuing, and it was found that the chief drawback of quad cable, that is, its increase in attenuation when dirty, could be overcome by including a spiral of thin aluminium tape between the inner belt and the outer sheath. This aluminium tape had very little effect on the screening properties of the cable and where the voltage was more than 5 V it was found necessary to use a copper tape so as to limit the radiation to 100 microvolts per metre at 10 yards as laid down by the G.P.O. The crosstalk over the band 5–10 MHz (or 'crossview' as it came to be referred to) was remarkably good considering the cable had not been made specially for television purposes. By adopting the *tête-bêche* system of distribution it was found that protection of only about 28 dB was required between two 405-line programmes, and this was attainable between the two pairs of a screened quad cable for distances up to about a mile which was the maximum length of feeder normally in use.
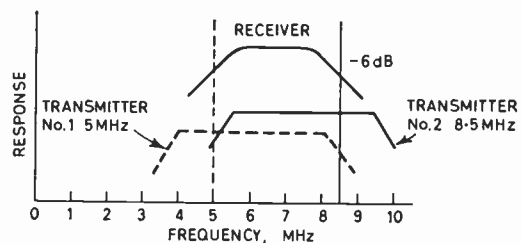
Fig. 2. *Tête-bêche* arrangement of the carriers and sidebands.

The *tête-bêche* arrangement of the carriers and sidebands is illustrated in Fig. 2, which also shows the frequency response of a wired receiver which can be seen to have flanks which are −6 dB at each of the two carrier points. Thus either programme can be equally well received on this receiver without it being necessary to carry out any switching within the receiver. The subscriber's installation thus takes the very simple form illustrated in Fig. 3.

An h.f. auto-transformer steps down the h.f. voltage on the feeder to a level suitable for the input to the receiver. The capacitor is included so as not to short-circuit the audio signal carried by the same pair. The resistors ensure that the source impedance

Fig. 3. Subscriber's installation.

Fig. 4. Factors affecting the overall economics of the distribution system.

feeding the drop-in (the vertical cable from the feeder to the programme selector switch) is correct. The programme required (vision and sound) is thus selected by means of a simple double-pole switch.

The next problem was to decide on the optimum input level to the wired receiver. Broadly speaking the choices open were:

(a) 1·5 V into 130 ohms which would have meant that no h.f. amplifying valves would have been required,

(b) 150 mV which would have necessitated one h.f. amplifying valve before the detector,

(c) 15 mV calling for two h.f. amplifying valves,

(d) 1·5 mV necessitating three h.f. amplifying valves.

Choices (a) and (b) would have required a large number of rather high-powered repeaters to maintain the level everywhere on the network, and the cost of these would have outweighed the saving in valves in the wired receiver. The main drawback to (d) is that the drop-in (the cable connecting the subscriber to the feeder cable) would have to be screened with copper tape to protect the wanted signal from interference from normal transmissions in the band, and this factor coupled with the extra valve required in the receiver was found to cost more than solution (c). In the case of (c) no screen is required if the drop-in is well balanced since the increase in attenuation due to dirt is negligible on a short length of some ten or twenty yards. A study of prevailing field strengths from sky waves in the band 5–10 MHz has shown that it is very rare for these to exceed some 20 or 30 mV/m,

and the protection of an unscreened drop-in when the wanted signal is 15 mV is adequate in this field strength without a screen.

A considerable amount of thought was also given to the optimum sending voltage from repeaters since this has a considerable influence on the overall economics of the system. The factors which are important are illustrated in Fig. 4. In Fig. 4(a) it can be seen that the cost per watt falls fairly rapidly at first as the power of the amplifier is increased and then the curve flattens out. Figure 4(b) illustrates the relationship between the area that can be covered and the power available from an amplifier, showing that the area covered rises rapidly at first with increase in power but ceases to increase so rapidly with a further increase in power. In Figs. 4(b) and (c), curves numbered (i), (ii) and (iii) represent the situation for either different street spacings or different signal levels applied to subscribers. It is now possible to combine results of curves shown at (a) with those at (b) in order to produce a third graph at (c), from which the lowest capital cost per unit area insofar as repeaters are concerned can be ascertained for any particular set of conditions. Actual figures have been deliberately omitted from these curves as, unless these are very fully qualified (for example with cable sizes), they might be misleading. Curves of this type were first prepared some fifteen years ago and have been continually revised as new valves and transistors have become available. It is interesting to note, however, that these curves have not changed very much, but there has been a tendency for the optimum sending power to fall. For example, as originally drawn for monochrome 405-line programmes where linearity

was not particularly important (providing, of course, there was no visible crushing) the optimum sending voltage was about 20 or 25 V into an 85 ohm feeder. More recently when considering 625-line programmes with colour the sending levels of 10 to 15 V into 110 ohms have been found to be practical and economical, bearing in mind the much lower level of distortion products permissible.

During the early 1950's a large amount of network made up of two screened quads for the vision programmes and quad cables of the 'sound-only' programmes was erected so as to be able eventually to distribute up to four television programmes plus some sound programmes. It was realized, however, that even this number of pairs might not be able to cater indefinitely for the probable increase in the number of television and 'sound-only' programmes. Research work was therefore directed towards trying to design a cable which would carry up to six television programmes plus a further three sound programmes. This required a nine-pair cable and an example is illustrated in Fig. 5(a). The six outer pairs are designed to carry vision programmes in the band 5–10 MHz together with their associated sound at audio frequencies, and the three inner pairs to carry 'sound-only' programmes.
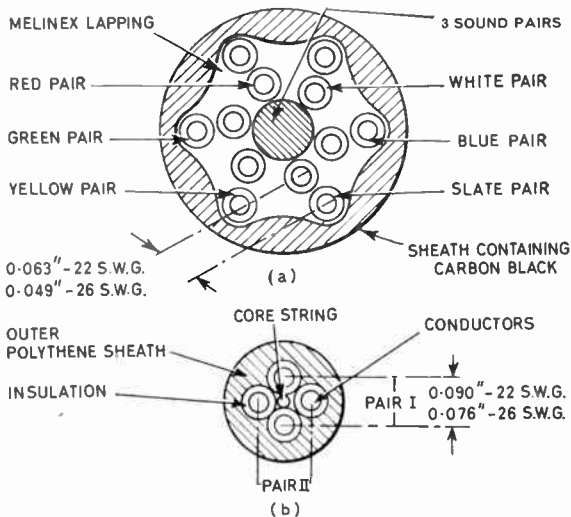
Fig. 5. A nine-pair cable.

A comparison between the spacing of the two wires of one pair (Fig. 5(a)) and that between a pair of screened quad cable (Fig. 5(b)), will show that the spacing is less on the twisted pair cable. This means that the pairs in twisted pair cable are less vulnerable to both radiation and interference as compared with

the quad cable, and also less vulnerable to the effect of dirt on the outer sheath. The presence of dirt can cause the attenuation of an unscreened cable to increase very greatly.[17] This is mainly due to the currents induced in the film of dirt. If the film resistivity were zero or infinity no power would be consumed. As a practical example the attenuation at 10 MHz of a cable with a film resistivity of 3 MΩ per yard is three times higher than it is with a film resistivity of 1000 MΩ or 100 kΩ per yard. By embodying a conductive compound in the sheath of the cable resulting in a sheath conductivity of approximately 20 kΩ per yard the effects of dirt on the outside of the cable can be completely eliminated, resulting in a considerable saving in cost. Thus a multi-pair cable could be used which was neater to erect and cheaper to manufacture than its screened-quad-plus-quad equivalent.

Table 1

Subjective tests carried out to establish crossview protection (all monochrome)

1. Synchronized carriers: same standards: carriers in phase	44 dB
2. Synchronized carriers: same standards: carriers in quadrature	32 dB
3. Synchronized carriers: one 405- and the other 625-lines	34–42 dB†
4. 405-lines: one-third line precision offset	40 dB‡
5. 625-lines: one-third line precision offset	38 dB‡
6. 405-lines: half line semi-precision offset	42 dB‡
7. 625-lines: half line semi-precision offset	40 dB‡
8. 405-lines: *tête-bêche* 3·5 MHz separation between carriers	28 dB
9. *Tête-bêche:* one 405-line programme interfering with one 625-line programme with a difference frequency between carriers set to an odd multiple of half line	34–40 dB*

† Dependent on phase of carriers and difference between frame frequencies.

‡ Flywheel synchronization required in receiver.

* Dependent on difference between frame frequencies.

While these cables were being developed it soon became clear that it would not be possible to obtain the same degree of crossview protection between adjacent pairs of wires as would be obtainable between alternate pairs or opposite pairs. A close study was therefore made of the various tricks that could be played with the arrangement of carriers so that these could be arranged to fit in with the crossview performance of the cable. A large number of subjective tests were carried out with panels of observers and some of these results are summarized in

ALTERNATES – BEST CROSSVIEW

{OPPOSITES AND / ADJACENTS} – SECOND BEST CROSSVIEW

6 PAIR
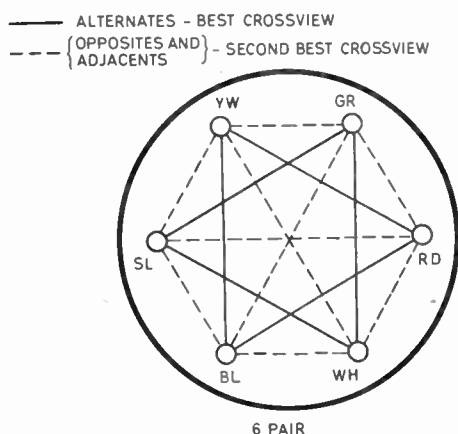
Fig. 6. Crossview protection obtainable between the pairs of the outer six of nine-pair cable.

Table 2

Practical application of data in Table 1 and Fig. 6

Colour of pair	Three 405-line + three 625-line programmes			Six 625-line programmes	
	Lines	Carrier frequencies MHz	Side bands	Carrier frequencies MHz	Side bands
Blue	405	8·3†	Lower	5·4	Upper
Slate	625	5·4	Upper	5·4 offset‡	Upper
Yellow	405	8·3	Lower	5·4	Upper
Green	625	5·4	Upper	5·4 offset	Upper
Red	405	8·3	Lower	5·4	Upper
White	625	5·4	Upper	5·4 offset	Upper
3 central pairs	Three sound programmes			Three sound programmes	

† Odd multiple of ½ line from 5·4 MHz.

‡ Precision ⅓ line from 5·4 MHz.

Table 1. From this it can be seen that the range of protection needed varies from the high figure of 44 dB for synchronized carriers down to the low figure of 28 dB for 405-line *tête-bêche*. The crossview protection obtainable between the pairs of the outer six of nine-pair cable is illustrated in Fig. 6. 'Best crossview' means about 45 dB at 8·5 MHz and 'second best' about 36 dB on a practical feeder.

The practical application of the information in Fig. 6 and Table 1 can best be illustrated by two examples from Table 2. One example illustrates the distribution of three 405 plus three 625-line plus three sound programmes, and the other six 625-line and three sound programmes. In the first example reliance is placed on the combination of *tête-bêche* half-line offsets, and in the second example reliance is placed on one-third-line precision offsets. Certain complications occur when colour is transmitted but it will be more convenient to deal with this as a separate subject later.

3.1.1. Coaxial cable

Crossview on cable varies with the length of the cable and the frequency used. If two identical coaxial cables run side by side the voltage induced in one from the other will be directly proportional to the length, i.e. doubling the length will increase the crossview by 6 dB. This is not usually important on the types of cable used for h.f. trunk systems since the crossview protection is of the order of 120 dB for a 250 yd length.

With a solid tube screen the protection will increase as the frequency is increased due to the skin effect of the screen. This relationship has been covered by Schelkunoff.[2] If braided cable is used the performance will depend upon the closeness of the mesh and the

contact resistance. Some information on this subject has been given by Hinchcliffe.[3]

3.1.2. Twisted pair cable

Doubling the length over which two twisted pairs run side by side will increase the unwanted signal by 3 dB. This is because the transfer of voltage is of a random nature due to a number of different causes, and thus follows a power law as distinct from the voltage law which applied to the coaxial cable where the voltage transferred is systematic and in phase.

Doubling the frequency in the range 5–10 MHz will reduce the crossview protection by 6 dB.

Care must be taken in using these simple relationships however, and further information will be found in a paper by Kinross and Russell.[4]

3.2. *Practical Application of H.F. System*

3.2.1. Aerial site

This is usually selected to be outside the main area of population which the system has to serve in order to ensure the best possible reception of signals. The type of aerials installed is dependent upon the strength of the signal that has to be received and any interference to which these signals may be subjected. High-gain v.h.f. aerials have been available for some years and Osborne has described high gain u.h.f. aerials for wired broadcasting.[5, 6] It is also well known how to arrange for a fixed null by the use of two aerials. It may, however, be worth describing the revival of an old idea to provide an adjustable null in any direction. This has proved useful recently on Bands I and III, when unexpected interference has tended to crop up from various directions. The
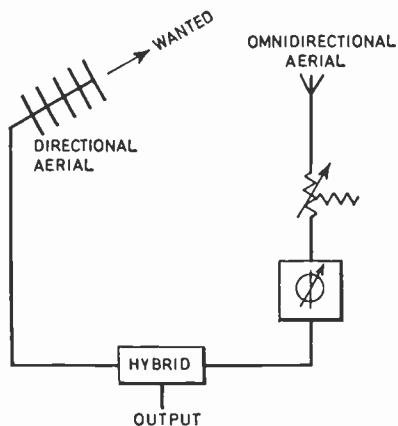
Fig. 7. Aerial with adjustable null.

method is illustrated in Fig. 7. The omni-directional aerial picks up the wanted and unwanted signal. The ratio of these to each other is different from their ratio in the main directional aerial and it is therefore possible to cancel out the unwanted signal by means of the adjustable delay line and attenuator in the omni-directional aerial feed. In cases of particular difficulty the omni-directional aerial may be replaced by a directional aerial pointing at the interference but this tends to reduce the flexibility of the system.

When the carrier frequencies of two television programmes had to be synchronized it used to be done by demodulating each signal and using the video frequency to modulate a single master oscillator. The apparatus required to do this was quite bulky and is shown in the top half of Fig. 8. Since then methods of pulling two carriers into complete synchronism have been developed and the resultant saving in equipment is shown in the lower half of Fig. 8. The method used is similar to that employed for reconstituting the suppressed colour sub-carrier in a colour receiver from the colour burst.

3.2.2. Trunk system

The object of a trunk system is to carry the signals generated at the aerial site to the furthermost parts of the network with the maximum reliability and the minimum amount of distortion. For that reason no subscribers are ever connected to a trunk system. Since it is desirable to introduce the minimum amount of crossview into a trunk system separate coaxial cables are used for each television programme and the accompanying sound is distributed on separate audio pairs. The sound used to be transmitted at several hundred volts representing several kilowatts of power. This voltage was stepped down by means of transformers at the distribution points to a level of 65 V, which is the normal level used on feeders.

More recently, with the advent of transistors, it has been found more economical to distribute the audio voltage at a level of a few volts which feed the input of transistor amplifiers installed at distribution points. Up to nine of these amplifiers can be fed from one d.c. power supply and the latter can be quite small since, as subscribers switch from one programme to another, so some amplifiers take more d.c. and others take less. The amplifiers have an output of 150 W which is sufficient to feed 600 subscribers. (The loudspeakers fitted to wired receivers are very much more efficient than those fitted to aerial receivers because the extra cost of providing this efficiency is less than the amount saved in amplifier costs.)
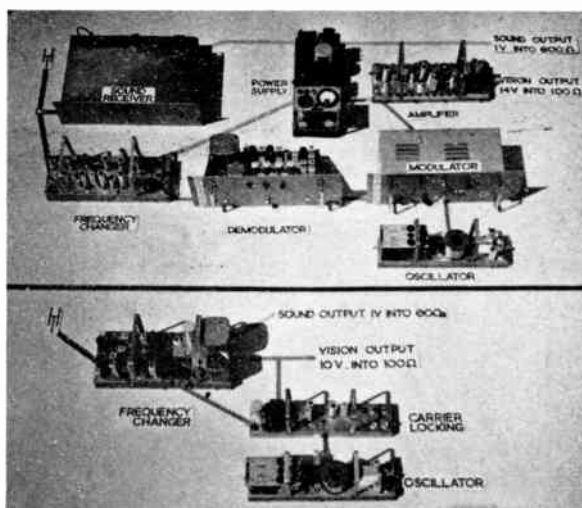
Fig. 8. Apparatus used to synchronize carrier frequencies of two television programmes.
Upper half: old type, Lower half: new type.

3.2.3. Distribution points

These are the points which obtain their vision and sound inputs from the trunk system, and from which emerge the cables carrying the signals to the dwellings of subscribers. These distribution points contain high-level link step-down transformers or transistor audio amplifiers, combined trunk and feeder vision repeaters and group delay and frequency correcting devices. The distribution point is normally capable of feeding some 16 000 route yards of feeder network, or between 1000 and 1200 dwellings. It is usual to place the distribution point in the centre of the housing area which it has to serve and the usual number of feeders emerging is three or four. The diameter of the area served is usually about $1\frac{1}{2}$ miles.

3.2.4. Feeder network

This consists of the nine-pair cable described above. Sound programmes are transmitted at 65 V (r.m.s.

on peaks) with a regulation not worse than 3 dB on full load. Vision programmes have a sending voltage of about 14 V, falling to 70 mV at the ends of the feeders. Two types of cables are used in approximately the following proportions:

  15% of 20/24 copper screened cable with an outside diameter of 0·55 in. This is used wherever the level of the vision signal is above 5 V.

  85% of 26/27 s.w.g. unscreened cable with an outside diameter of 0·38 in. This is used between levels of 5 V and 70 mV, the lowest level to which the signals on the feeder are allowed to fall.

The heavier of the above cables is used on the main stem while the lighter gauges are used on spurs. They are connected to each other by means of junction boxes which measure approximately 8 in by 4½ in by 1¼ in. These junction boxes contain h.f. transformers which step down the h.f. voltage in the main stem to that required in the spur without causing any visible reflections on the main stem.

3.2.5. Subscriber installation

The main features of the subscribers' installations have already been mentioned and illustrated in Fig. 3. The junction box contains the 'inserts' for connecting the drop-in cable to the feeder cable. There is one 'insert' for each television programme and each 'insert' contains a step-down h.f. transformer, audio blocking capacitor and a pair of isolating resistors. These inserts are chosen to attenuate the television signal by any amount from 9 dB up to 28 dB according to the level of signal on the feeder, and are of varying types capable of feeding one, two or four subscribers. They are designed in such a way that when connecting a junction box to the feeder it is not necessary to cut or even to strip the insulation from any of the feeder wires since contact is made by means of serrated phosphor-bronze washers which bite through the insulation and make contact when the terminals on the inserts are screwed up. The drop-in cable is connected to the outlets of the inserts by means of pins and eyelets. Each wire is first inserted in the pin and the eyelet is pushed over. The action of doing this not only strips off the insulation but makes contact with the conductor. Thus, though there are quite a large number of connections to be made in a junction box these can, in fact, be made quickly by relatively unskilled labour. The drop-in cable itself consists of three pairs of 27-gauge cable at the centre, surrounded by six pairs of 26 s.w.g. The programme selector switch can be a completely standard switch wafer, but a printed circuit type of switch has also been designed since this avoids the need for any connecting wires from the switch to the drop-in

terminals and eliminates any chance of crossview due to careless wiring. If an aerial receiver is installed an inverter is required to change the incoming frequencies from the h.f. band to the v.h.f. band. Its circuit is very simple, consisting of a transistor oscillator and a diode mixer, and since its power consumption is only two watts it can be left switched on permanently.

4. V.H.F. Systems

4.1. *Principles*

When only one television programme was broadcast it was only necessary to introduce a single channel amplifier between the aerial and a coaxial cable which fed subscribers through simple resistive pads. Even in those days, however, one had to be careful not to try to obtain too much output from the amplifier, or cross modulation occurred between sound and vision in the output valve. When the I.T.A. started in 1955 it was possible to add another aerial and amplifier and combine the outputs from the two amplifiers into a coaxial cable. This was the simplest form of head-end equipment but it meant distributing the I.T.A. programme on Band III and the considerably higher attenuation of the cable at those frequencies made it difficult if not impossible to feed the blocks of flats previously comfortably energized on Band I. Thus most systems (unless they were extremely small) had to use a frequency changer to convert the incoming Band III signal into a vacant channel on Band I. A number of problems began to come to light, particularly as v.h.f. systems were extended from blocks of flats to small towns.

(i) It was sometimes no longer possible to use the same carrier frequency on the cable as that of the transmissions received by people with aerials of their own in the vicinity. If the transmitter was fairly close its signals induced voltages in the coaxial network or, if the cable was well enough screened to prevent this, induced voltages directly into subscribers' receivers, with the result that ghost images appeared. If the transmitter was some distance away radiation from the coaxial cable could produce ghost pictures on the receivers of those people trying to receive the programme with their own aerial. Thus a frequency changer at the head-end was often required.

(ii) If broad band repeaters were used, i.e. covering a band from about 40 MHz to about 230 MHz and amplifying several programmes, care had to be taken to avoid cross-modulation and intermodulation between programmes.

(iii) Use of adjacent channels on the network must be avoided since the selectivity of receivers is not usually good enough although modern receivers are improving in this respect.

(iv) With the growth in the number of programmes, and the increase in the number of obstacles in the way of using various channels in Band I described above, it has for some time been essential to use channels in Band III which means that many more repeaters have to be installed to overcome the increased attenuation of the cable.

(v) Another difficulty on v.h.f. systems is that dual standard receivers are generally manufactured to operate on v.h.f. on 405-lines and on u.h.f. on 625-lines. Since it is not practical (except on extremely small systems) to distribute u.h.f. frequencies due to the high attenuation, these receivers have to be modified to operate on 625-lines on v.h.f. This is now generally recognized by manufacturers and the change is no longer so difficult to make on modern receivers.

The two most important points which govern the level at which broad band v.h.f. repeater amplifiers operate are the fact that if the output is too high cross-modulation sets in, and if the input level is too low noise becomes a limiting factor.

A brief summary of the empirical rules that have been found to govern these factors may be usefully stated, as follows:

Noise: Doubling the number of repeaters in tandem will increase this by 3 dB.

Cross-modulation: This takes the form of the un-wanted picture or its synchronizing pulse appearing as a background to the wanted picture. Doubling the number of repeaters in tandem worsens the cross-modulation by 6 dB.

Inter-modulation products: These take the form of random patterns built up from the sum and difference frequencies of the various signals in the system and their harmonics. Doubling the number of repeaters in tandem increases this effect by 3 dB.

Increasing the output of all the programmes by 1 dB increases the cross-modulation on any of these programmes by 2 dB.

Doubling the number of interfering programmes makes the cross-modulation worse by 3 dB.

The application of these rules can be best considered by an example. It is generally considered that the worst level of cross modulation that is acceptable is −46 dB and a tolerable signal/noise ratio in terms of ratio of video-to-r.m.s. noise is usually taken as 42 dB. In the case of a repeater whose signal/noise ratio with an input of 5 mV is 46 dB and has −60 dB of cross-modulation with two television programmes each at 100 mV, the operating levels if it is desired to operate sixteen in tandem, carrying five programmes are:
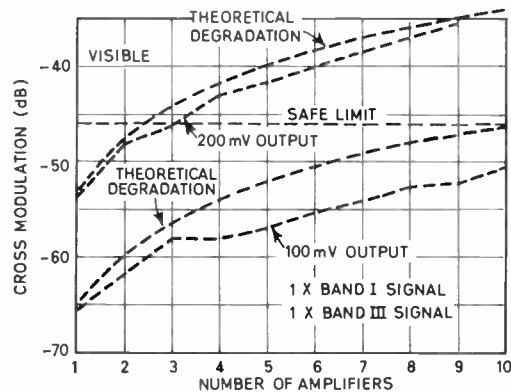
Fig. 9. Dependence of cross-modulation on repeater output.

minimum input:	12·5 mV
maximum output:	40 mV
gain:	10 dB

The 6 dB cross-modulation law is not always closely followed in practice, and it appears to depend upon the absolute level of output of the repeater. An example of this is illustrated in Fig. 9, where it will be seen that repeaters with outputs of 200 mV follow the theoretical law much more closely than those of only 100 mV.

4.2. Brief Description of V.H.F. Systems

As in the case of the h.f. system, no subscribers are connected to the trunk cable which usually consists of a coaxial cable about 0·6 inch in diameter with an attenuation of 1·3 dB per 100 ft at 230 MHz. Trunk repeaters are usually required every 400 to 600 yards. The amplifiers which energize the feeders are usually called distribution amplifiers and their most important function is to provide the largest possible output without cross-modulation so that the longest length of feeder can be energized from one point. It may be mentioned here that it makes for the most economical overall system if about half the permissible cross-modulation is produced in the distribution amplifiers and the rest of it is taken up in the tandem repeaters. If the permissible cross-modulation was evenly divided throughout all the repeaters this would seriously limit the output from a distribution amplifier, and limit its radius of action.

One usually reckons that a distribution amplifier can energize approximately 450 yards of cable or the equivalent of about 30 dwellings. Subscribers are connected to the feeders by means of 'tap-off units'. These are often designed in such a way that it is only necessary to make a small hole in the outer sheath and screen of the feeder in order to enable a probe from the tap-off unit to make contact with the inner conductor. The range of loss of tap-off units varies

from 10 dB to 40 dB (depending on the signal level on the feeder): simple resistors are usually used for the higher values but it is usually worth paying extra for transformers for the lower values due to the greater efficiency of these which enable a greater length of feeder to be energized from a given distribution amplifier. Careful isolation is required between the drop-in and the subscriber's receiver to minimize the risk of mains voltage finding its way into the network. G.P.O. regulations call for a pair of good quality capacitors between the coaxial drop-in and the coaxial socket of the receiver.

The earliest v.h.f. systems used cable with a single copper braid screen. Later it was found necessary to use two such screens. More recently cables have been developed using a screened longitudinal copper tape which has got a welded seam, or alternatively is held in position by means of a loosely-woven copper braid. Both of these types of cable have got better screening properties than the early copper braid type. A more recent development, which has been described by Hinchcliffe,[7] has been the use of flat strips instead of braid.

5. Comparison Between H.F. and V.H.F. Systems

The growth of the two systems has already been illustrated in Fig. 1. This shows that there are more v.h.f. than h.f. systems, but more h.f. than v.h.f. subscribers. The correct figure for v.h.f. subscribers is, however, higher than shown because these graphs take no account of small installations working in individual blocks of flats where no special wired broadcasting licence is required if cables do not run from one property to another. The accurate figure for v.h.f. subscribers is therefore not known and the only point which is certain is that h.f. systems are, on average, appreciably larger.

The simplest way to consider the relative costs of the two systems is to eliminate all those items which cost about the same for each system. These are: aerial arrays, low-noise amplifiers, way-leaves, cable-erection labour, underground work.

The feeder cable also costs about the same, and this is best illustrated in Table 3, where the attenuation of coaxial cable at 230 MHz with that of twisted pairs at 5 MHz is compared. On this basis, the twisted pair cable is less expensive than the coaxial. It is interesting to note that for a given attenuation the overall diameter is also smaller on the multi-pair cable which is an important point when considering the appearance of the network on private houses.

As regards the trunk system the cost of this per 100 yards of trunk route is usually greater for an h.f. system than for a v.h.f. system because it embodies

one coaxial cable per television programme and individual pairs for each sound channel. On the other hand, the individual coaxial cables can be very much smaller and cheaper than their v.h.f. counterpart because the attenuation at 5 MHz is so much less than it is at 230 MHz. There is another and almost more important counterbalancing factor, namely that h.f. distribution points are usually over one mile apart, whereas v.h.f. ones are much closer. This means that more trunk route is needed for v.h.f. systems and since most trunk systems have to be buried underground, nowadays the additional cost of trenching and making good for this extra length can cost as much if not more than the extra cost of the cable used in the h.f. system. It is, of course, true that the v.h.f. trunk system only requires one coaxial cable and it may therefore not be necessary to put quite such a high proportion of it underground.

Taking all these factors into account, the cost comparison on trunk systems can best be summarized by saying that the h.f. trunk system costs more per yard of trunk than a v.h.f. system but there is less of it, and these two factors approximately cancel out. As regards trunk repeaters, the spacing between these on v.h.f systems is usually about $\frac{1}{4}$ mile and on h.f. systems between one and two miles. However, with the recent great improvement in performance of v.h.f. broad band repeaters only one is required to cover all the programmes on v.h.f. systems, whereas a separate repeater is of course required for each of the h.f. programmes since they are on separate pairs. Thus here again these costs approximately cancel out.

Table 3
Comparison of different types of feeder cables

Cable type	Conductor size s.w.g.	Cost 1000 yd	Attenuation dB/100 ft		Outside diameter inches
			220 MHz	5 MHz*	
Coaxial		£200	·98	—	0·6
Coaxial	12	£168	1·3	—	0·57
Coaxial	17	£76	2·35	—	0·35
9-pair screened	20/24	£200	—	0·63	0·55
9-pair unscreened	26/27	£92	—	0·98	0·38

* On each of six pairs.

The main item which must cost more for v.h.f. systems than for h.f. systems is of course the distribution or feeder repeaters. A typical coverage of a v.h.f. feeder repeater is some thirty or forty dwellings, whereas the coverage in a Rediffusion h.f. system is usually about 1000 dwellings. Burke[8] has mentioned

that in a British Relay Wireless (B.R.W.) system 56 000 homes are fed from eleven distribution points, i.e. about 5000 homes per distribution point. This is because B.R.W. use higher powered repeaters and heavier cable than Rediffusion. Similarly, a champion of v.h.f. systems could claim to cover 100 dwellings but the cable for this would cost about £170 per thousand yards instead of £92 for nine pairs unscreened. Comparing the number of h.f. and v.h.f. repeaters for cables of *similar cost* we need about thirty times more v.h.f. repeater points or seven times as many repeaters to distribute four programmes.

We now come to the items which cost more for the h.f. system and these are:

(a) Subscriber's drop-in: the junction box is more complicated and a programme selector switch is required; an inverter is also required if an aerial set is installed.

(b) Head-end equipment: synchronizing and offsetting carriers to a high degree of accuracy is more expensive than frequency changing on v.h.f. systems; also audio amplifiers are needed for h.f. systems.

In large towns the head-end equipment forms a very small proportion of the total cost of the system and we find that the additional cost of v.h.f. amplifiers is just about counter-balanced by the extra cost of the items on the h.f. system just mentioned.

We are thus left with the most important single item when comparing the two systems, and that is the cost of the subscriber's receiver. The simple wired receiver required for h.f. systems retails at some £20 less than an aerial receiver, but this difference might be reduced by £6 if the aerial receiver can be bought without a u.h.f. turret since this is not needed on a v.h.f. system. However, if Band II f.m. sound is included in a television receiver to enable it to receive a number of sound programmes such as are normally distributed as a matter of course in h.f. systems, several pounds must be added to its cost.

Thus the question of cost can be summarized quite briefly as follows: The cost of networks complete with amplifiers is approximately the same for both h.f. and v.h.f. The cost of the head-end equipment is slightly more for an h.f. system, but this cost tends to disappear in a large town. The cost of the wired receiver for an h.f. system is appreciably less than an aerial set on a v.h.f. system.

The following deductions can be made from the foregoing:

(i) In a town the most economical system overall is the h.f. system.

(ii) If there is a small block of flats in a town that has no h.f. distribution system or is in an area

so small that the number of v.h.f. repeaters is of little consequence, the choice will almost certainly be for a v.h.f. system.

(iii) In a hotel, however, where the owner (as distinct from the owner of a block of flats) would be interested in owning or renting the cheaper and more reliable wired receiver, the choice would probably be an h.f. system if the number of h.f. receivers installed is large enough to support the additional cost of an h.f. head-end equipment.

(iv) If a block of flats is in a town which is already wired for an h.f. or a v.h.f. system, the argument is more evenly balanced. The owner will certainly want to have some sort of wired system to prevent the erection of private aerials all over the roof and he will probably be influenced in his choice by the type of system that is already in use in the town to avoid having to install his own private head-end equipment.

It might be mentioned in passing that Rediffusion have developed a system that provides the advantages of both techniques, but unfortunately this system must at present be limited to large blocks of flats in towns already wired with an h.f. system. In a few blocks of flats in Portsmouth a normal nine-pair twisted pair cable has been installed, but on one of the outer pairs three television programmes are distributed in the band 40–220 MHz. The attenuation of these pairs is about 8 dB per 100 ft at 220 MHz so that it is not economical to cover more than an individual block of flats. Such an arrangement means that both h.f. and v.h.f. inputs are available in each flat, and if a tenant has a normal aerial receiver he can connect it to the network without the need for the individual inverter which is usually required. Each v.h.f. signal is derived from the h.f. network once only in each block by means of a standard inverter as already described. This means that only three inverters are required in each block of flats instead of possibly fifty or sixty. These blocks at Portsmouth are almost ideal for this particular type of system since they consist of eighteen floors with eight flats per floor.

A few years ago the saving in maintenance cost of a wired receiver as compared with an aerial set was about 30%. Recent fault rate figures for the period February to July 1966 for receivers which are alike in all respects except the fact that one type is wired and the other type operates from an aerial show that the fault rate for wired receivers is slightly less than half that of aerial receivers. One might conclude that this must be due to the introduction of u.h.f. tuners, but in fact this only accounted for some 3% of all the faults in the aerial sets and the rest of the

faults were fairly evenly distributed among such items as v.h.f. tuners and audio amplifiers which do not exist on wired sets. An interesting point was that the faults which required the attention of a serviceman but were in fact only due to faulty adjustment of a receiver were twice as many for aerial receivers as for wired receivers. The same staff maintained both types of receiver. More experience in the manufacture of dual-standard aerial receivers will no doubt reduce this differential, but wired receivers will always cost less to maintain.

One small point in favour of h.f. systems worth mentioning is the ease and simplicity of providing for the selection of programmes by remote control.

Comparisons of h.f. and v.h.f. systems as seen by housing authorities have been published.[9-11]

Table 4
Relative costs for h.f. and v.h.f. systems

	Charges per week	
	h.f. multi-pair	v.h.f. coaxial
Service rental	1s. 6d.–2s. 3d.	1s. 3d.–2s.
Sound receiver	—	6d.
Television receiver	7s. 2d.	8s. 8d.
Saving in power	− 3d.	—
Total	8s. 5d.–9s. 2d.	10s. 5d.–11s. 2d.

Finally, perhaps the most important single argument is the relative cost of h.f. and v.h.f. to the general public. Table 4 shows a comparison based on a 'community service', i.e. where the operating company has a bulk contract with a housing authority

to wire every house in the district. The charges shown are typical of those being made today.[12]

6. Colour Television

Designers of wired broadcasting systems have had difficulty in planning for colour television during the last few years. First it was thought that they would have to distribute 405-line programmes in colour using a scaled-down N.T.S.C. system. Next it was to be 625-lines, using a scaled-up N.T.S.C. system. Later it was decided that the system should be PAL on 625-lines. More recently there was discussion of colour on 405-lines as well. As a result of the Postmaster-General's announcement in February 1967, there will be a mixture of 405-lines in monochrome and 625-lines in colour for some years, as far as wired-broadcasting is concerned, and the following factors must therefore be borne in mind.

The first point to consider is how low we must keep the various forms of distortion which can occur on a wired broadcasting network. As a starting point for this, in Table 5 are shown some figures extracted from the report of the E.B.U. *ad hoc* Committee which were published in the *E.B.U. Review*.[13] The only figures which have been added are those in the last column which are suggested as a reasonable target for wired broadcasting systems. The figures in this last column are generally low compared with the figures in the other columns, the only exception being that concerning the ratio of amplitude of chrominance to luminance. Since colour receivers always include a control for adjusting the amplitude of the chrominance information, it is felt that if steps are taken at the aerial site to make the ratio of chrominance to luminance as nearly correct as possible (if necessary

Table 5
Types of distortion on wired broadcasting network for colour television

Type of distortion	Grade 2A 'Fairly good to good' PAL(D)	Grade 2A 'Fairly good to good' PAL(S)	Arithmetic sum of all items except receiver	Receiver only	Suggested for wired broadcast network
Group delay error between luminance and chrominance carriers	+200 ns −300 ns	+200 ns −300 ns	± 75 ns	?	±40 ns
Level dependent phase of the chrominance signal	± 40°	± 12°	± 22°	± 5°	± 3°
Ratio of amplitude of chrominance to luminance	± 2·5 dB	± 2·5 dB	> ± 3 dB	—	± 2 dB (after correcting received signal)
Level dependent amplitude	25%	25%	43%	20%	5%
IQ crosstalk	—	—	—	—	<10%

by means of a separate chroma a.g.c.) and this is then maintained to within ± 2 dB over the rest of the network, this should be quite adequate. Changes on broadcasting networks are due to changes in programme sources and there is room for some improvement here.

The author has seen no figures covering these forms of distortion for v.h.f. equipment except in the U.S.A.[14] and it is important that British manufacturers of this type of equipment should now publish some figures as soon as possible. It is probably true to say that there is very little risk of serious group delay distortion occurring over an individual television channel on an amplifier covering the band 40–220 MHz: nevertheless there may be odd and unexpected effects which should be looked for and measured. V.h.f. aerial site equipment will have to be selective and this may cause group delay distortion towards the band edges similar to that encountered in aerial receivers and these two effects will add together. The fact that the group delay distortion at the transmitter is in the opposite direction will not be sufficient to compensate for both these effects.

As regards h.f. equipment developed for the Rediffusion system, it has been necessary to design 625-line repeaters to be extremely free of distortion in any case since the second harmonic of the 5·4 MHz luminance carrier falls among its own upper sidebands. There has therefore been no problem in meeting the limits suggested for level dependent phase or level dependent amplitude with up to six repeaters in tandem. This number of repeaters will cover a radius of some 9 miles or an area of 250 square miles, which is ample for most wired broadcasting systems.

As regards group delay distortion the amount of this in one repeater is shown in Fig. 10(a). The addition of a single group delay correction element will reduce this error to that shown at (b) and with six repeaters in tandem the overall result is shown at (c). We have found this quite adequate for monochrome pictures, and it is probably adequate for colour. However, the addition of two more elements of group delay correction reduces the error in one repeater to that shown at (d) or to six repeaters in tandem to that shown at (e) and this amount of distortion will be completely invisible.

H.f. wired receivers require so little amplification and selectivity that the group delay error from the h.f. input to the detector stage is only some 50 ns as compared with the 200 or 300 ns usually encountered on aerial receivers. Thus the overall group delay through the repeaters and the h.f. receiver of an h.f. system can actually be less than on a single aerial receiver.

Further information on distortion measurements on h.f. systems has been given by Gargini.[15]

A piece of equipment which has been found very useful for appraising the suitability of networks for colour is based on proposals made by Tanner[16] whereby a $10T_c$ or $20T_c$ pulse is transmitted as double sideband modulation on a chroma subcarrier along with the same pulse as modulation on the luminance carrier. (T_c is the half-period of 1 MHz, which for testing purposes is regarded as the nominal bandwidth of one sideband of the chrominance signal.) This provides a simple indication of the difference in delay as between the luminance and chroma carrier. It also gives some indication of group delay distortion over the chroma band.
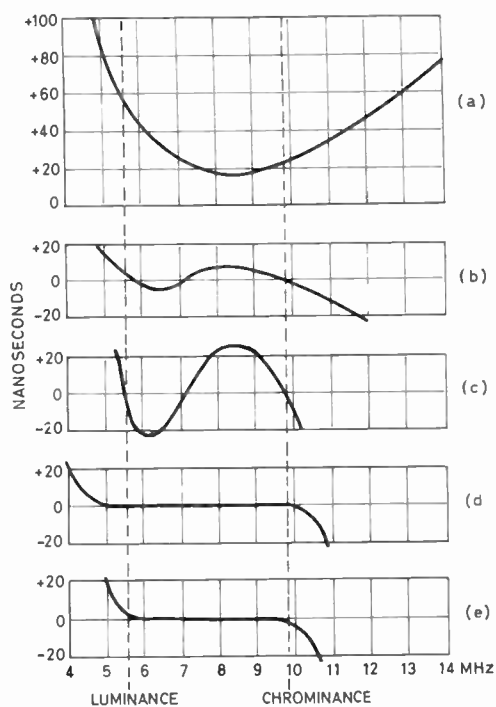
Fig. 10. Group delay distortion in repeater systems with and without correction elements.

(a) 1 repeater without group delay correction.

(b) 1 repeater with 1 group delay correction element.

(c) 6 repeaters with 1 group delay correction element in each.

(d) 1 repeater with 3 group delay correction elements.

(e) 6 repeaters with 3 group delay correction elements in each.

A problem which exists on h.f. systems and not on v.h.f. systems is crossview. The only definite and immediate problem that we know we must deal with is the addition of colour to one 625-line programme distributed alongside a number of 405-line monochrome programmes. The method of doing this is

illustrated in Table 2 and only very small changes are now necessary to enable colour to be added to the 625-line programmes.

Consider the case of the mixed standards shown on the left-hand side of Table 2. The carriers that matter most are shown in Fig. 11 assuming that the 625-line programme is operating on the PAL system. For simplicity it is assumed that the chroma information is entirely $(B-Y)$ and we can for the moment ignore the phase alternation of the $(R-Y)$ component.

The difference between the unwanted 8·3 MHz 405-line carrier and the wanted 5·4 MHz 625-line luminance carrier is an odd multiple of half line and thus favourably offset. The other difference frequencies are odd multiples of $\frac{1}{4}$ or $\frac{1}{3}$ line and thus also reasonably favourable.[18]

If we now consider the $(R-Y)$ component the effect of reversing this in phase on alternate lines has been shown by Bruch[19] to be equivalent of a $\frac{1}{2}$ line change in frequency. This turns the $\frac{1}{4}$ line relationships into $\frac{3}{4}$ line which are similar and turns the $\frac{1}{3}$ line relationship into $\frac{5}{6}$ line which is not quite as good as $\frac{1}{3}$ line but the best compromise that can be achieved.

We would have preferred the half-line relationship that existed between chroma and luminance on the N.T.S.C. system but tests have shown that the crossview protection on multi-pair network will be adequate for the system shown in Fig. 11. Consider now the
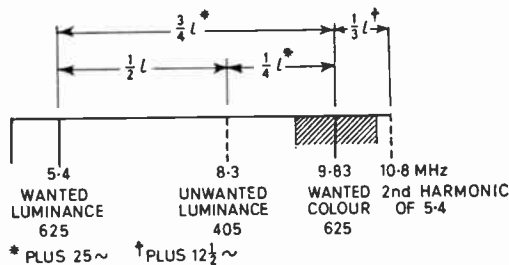
Fig. 11. Multi-pair network system with adequate crossview protection.

synchronized luminance carriers on 5·4 MHz. The chroma sub-carrier will only be approximately synchronized due to the permissible tolerance on the colour sub-carrier frequency. B.B.C. tolerances in this respect have always been very good and limited to $\pm$ 1 Hz. This is acceptable. Some European countries suggested recently that this tolerance should be increased to $\pm$ 5 Hz, which could result in a 10 Hz beat between two chroma sub-carriers. This was found to require some 9 dB more protection than a 2 Hz beat, and would not be acceptable on the system described. This increased tolerance is also

undesirable in respect of precision offset luminance carriers, and thus will affect the interference between broadcast transmitters. Protests have been lodged and it is hoped that European Broadcasting Authorities will agree to a tolerance of 2 Hz, since this is in fact not a difficult accuracy to achieve with modern equipment. Thus, there should be no difficulty in distributing three 625-line colour programmes alongside three 405-line monochrome programmes.

7. The Future

7.1. *Stereophonic Broadcasting*

Regular broadcasts have recently started on the f.m. transmitters of the Third Programme of the B.B.C. and there should be no difficulty in re-broadcasting these in Band II on a v.h.f. coaxial system. Similarly, there would be no difficulty in doing so using audio frequencies on the standard nine-pair cable of an h.f. system. This might, however, be a rather extravagant use of a pair which, though spare today, might be needed later on for an additional programme. A more rational way of dealing with stereophonic sound would be to distribute $(L+R)$ as at present by audio on one of the normal pairs, and $(L-R)$ by f.m. sound at a carrier frequency of about 11 MHz. By operating at a somewhat lower level than the television signal on the same pair, the f.m. sound could be amplified by the same repeater that is used for the television channel. Subscribers wanting to receive stereophonic sound would be provided with an f.m. demodulator and audio matrixing unit. Such a demodulator would, incidentally, be considerably simpler than the type of Band II stereo f.m. receiver required to receive stereophonic signals from a coaxial v.h.f. network.

7.2. *Bandwidth Compression*

These techniques, when perfected, are likely to apply more to point-to-point transmission over long distances, rather than to feeding large numbers of receivers over comparatively short distances. If applied at all to wired broadcasting therefore the use is likely to be limited to trunk systems. In some circumstances bandwidth expansion may be more desirable than compression as illustrated in the following example which relates to an out-of-band colour sub-carrier system, as suggested by Gargini.[15] In the Rediffusion h.f. system already described with a luminance carrier at 5·4 MHz and chroma sub-carrier at 9·83 MHz a pilot carrier is also transmitted at $9\cdot83 \div 3 = 3\cdot28$ MHz. This is multiplied by three in the wired receiver and used to demodulate the colour information. This is a cheaper way of reconstituting the suppressed chroma carrier and saves a quartz crystal and eight active devices as compared with a conventional synchronous detector

on N.T.S.C. Something similar will no doubt be achieved on PAL.

7.3. Rural Districts

A vital economic factor in wired broadcasting is the amount of cable required for each dwelling. Wired broadcasting at present, therefore, tends to be a viable proposition only in built-up areas. However, there has recently been so much difficulty in working out plans for obtaining complete coverage of Great Britain at u.h.f. that wired broadcasting may not be out of the question even in rural districts. A recent estimate of the cost of providing three television programmes to an area consisting of a village of some 400 dwellings, 4 hamlets with 85 dwellings, and 10 isolated farms spread out over an area of seven square miles showed that wired broadcasting would be cheaper than the installation of radio transmitters. It was found that conventional methods of h.f. network wiring could be applied to the houses in the village and hamlets but special arrangements would be necessary for the farms. One farm, for example, would have required a 'drop-in' of nine-pair cable of about one mile in length. A cheaper solution in this case would be to provide a three-wire cable. Two of the wires would carry the selected sound and vision signals while the combination of the three wires would be used to select the programme required by that particular subscriber from the nearest section of conventional network.

7.4. Systems Conversion

While mixed 405- and 625-line standards continue it may be worth while installing system conversion units in large towns, thus enabling all the subscribers on wired broadcasting networks to use nothing but single standard receivers.

7.5. Wired Broadcasting Networks as Communication Channels

Wired broadcasting networks could provide a channel for all sorts of messages, statistics and control systems. For example, the Electricity Council is very interested in the consumption of electricity at any one moment for such different purposes as space heating, cooking and water heating. A simple set of transducers to supply this information could be installed in a few homes on a sampling basis and the information fed back over the wired broadcasting network to a central point where results could be analysed. At present the system of clocks and controls installed in private homes for controlling off-peak electric power at cheap rates is complicated and expensive. A cheaper form of control could be provided over wired broadcasting networks. A practical way of combining a wired broadcasting network with

a telephone system at present seems remote as the requirements are so different. However, if one is prepared to look far enough ahead to the day when as many homes have telephones as today have television, even this possibility should certainly not be ruled out altogether. The main co-operation that can be seen in the immediate future between wired broadcasting and telephone services is over such matters as the sharing of trench facilities and pole routes.

7.6. Pay Television

In a system of this type subscribers have to pay each time they watch a programme. A pay television system must be able to perform the following functions:

(i) Inform a subscriber how much he will have to pay to watch a particular programme. There are technical and non-technical solutions to this.

(ii) If he is a cash subscriber, ensure that he has put enough money in the coin box to cover the cost of the programme he wishes to watch, or has just watched.

(iii) If he is a credit subscriber ensure that the amount he owes is shown either on the subscriber's equipment or in the central control.

A number of solutions to these problems have been described.[20-23]

The Government have sanctioned the experimental operation of pay television by wire and one system is now operating in the Westminster and Southwark area in London.[23] Another will be starting shortly in Sheffield. These experiments are expected to provide useful information on the public reaction to this type of service, and should also show whether pay television can find sufficient new programme material to justify itself. No advertisements are being allowed, and it may be two or three years before a decision is made as regards introducing this type of system into other towns.[24]

Both these experiments are being carried out on h.f. systems. The availability of multi-pair cable reduces the cost and complexity of the equipment needed to control the coin box or credit meter in the subscriber's home as compared with what would be required on a v.h.f. system.

7.7. Optical Methods

To be able to produce a picture in every home without the need for a cathode-ray tube or a time-base unit must surely be a worthwhile objective. To achieve this by established optical methods would be very difficult but do not lasers and masers and fibre optics open up all sorts of exciting new possibilities?

Is it not possible to imagine a bundle of fibres carrying the component parts of a picture by means of individual laser or maser beams? There is already some evidence to show that this can be done. If these beams were to fall on a sensitive phosphor surface in each home, might this not create a bright enough picture to enable this to be projected through a lens on to a screen?

7.8. Conclusion

Finally, we come to a complete paradox. There have been examples of a limited form of wired broadcasting in most of the capital cities of Europe right back to the beginning of this century. It required conventional broadcasting to popularize the idea of entertainment and education in every home in the country. This has now become such a well-established need that the author believes the rational way to carry on with it in future will be by the growing use of wired networks.

8. Acknowledgments

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An Optical Communication Link for Laboratory Demonstration using Microwave Modulation of the Light Beam

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Summary: A simple system designed to demonstrate the possibility of transmission of information over an optical path, using a microwave sub-carrier, is described. It is shown that a satisfactory signal/noise ratio can easily be obtained under laboratory conditions, and the practical limitations of such a system are briefly outlined.

1. Introduction

In the last decade a considerable amount of research has been concentrated on methods of modulating a laser beam as a step towards a practical optical communications link.[1] A major problem in the design of such a system concerns the manner in which full use can be made of the very wide bandwidth which is potentially available, and a possible approach may be to modulate the optical carrier with a number of sub-carriers in the microwave region.[2] Each sub-carrier would then be modulated with information at a rate corresponding to frequencies in the video range.

In order to show how such a link might operate, an embryonic version of a complete system has been constructed in which a low-power laser is modulated by a single X-band microwave sub-carrier, and audio frequency modulation has been transmitted over a short distance by pulse-duration modulation of the microwave source. In a simple demonstration of this type it is not necessary to recover the microwave frequency during the demodulation process, and a photomultiplier capable of responding to changes in the light signal up to a rate of a few hundred kilocycles is adequate. However, if more than one microwave sub-carrier were present it would be necessary to use some form of microwave photodetector, so that the information channels could be separated before the final demodulation process.

2. The Transmitter

The optical carrier, which is derived from a helium-neon gas laser operating at 0·63 µm, is passed through a polarizing prism and then enters an electro-optic modulator. The modulator, shown in Fig. 1, consists of a crystal of potassium dihydrogen phosphate

(KDP) mounted in a microwave cavity designed to resonate in a TMO13 mode at 9·3 GHz.[3,4] The cavity, shown in Fig. 2, is loaded with polystyrene, and the dimensions can then be chosen such that the phase velocity of the microwave field matches that of the light. Constructive interaction between the microwave and optical signals will then take place along the whole length of the crystal. The microwave signal is itself a sub-carrier, which is derived from a low power (10 watt) magnetron and, for our demonstration, pulse width modulated with an audio signal, the pulses having a mark/space ratio of 1 : 4 and a repetition frequency of 10 kHz. This frequency is sufficiently high for thermally induced modulation to be negligible.

A diagram of the modulator circuit is shown in Fig. 3. The transistors TR1 and TR2 form a multivibrator producing square waves at a frequency of 10 kHz; these pulses are passed to TR3, which is an emitter follower, and then into an R-C network which delivers a sawtooth waveform to the base of TR5. A long-tailed pair comparator circuit is formed by TR5 and TR6, and the audio frequency information signal is introduced at the base of TR6. The mean width of the pulse can be varied by the potentiometer RV2, and an emitter follower (TR9) is included so that this section of the circuit can be mounted away from the magnetron. The valve circuitry is mounted adjacent to the magnetron, and the pulse from TR9 is amplified and fed into a standard modulator power amplifier.[5] The magnetron output is coupled to the cavity, as shown in Fig. 1, and a small fraction of the reflected power is detected and used for monitoring the cavity resonance.

3. The Receiver

The optical detection system consists of an analysing prism set at right angles to the plane of polarization of the laser beam, followed by a photomultiplier with a red sensitive photocathode. If the KDP crystal is set with its optic axis along the direction of light

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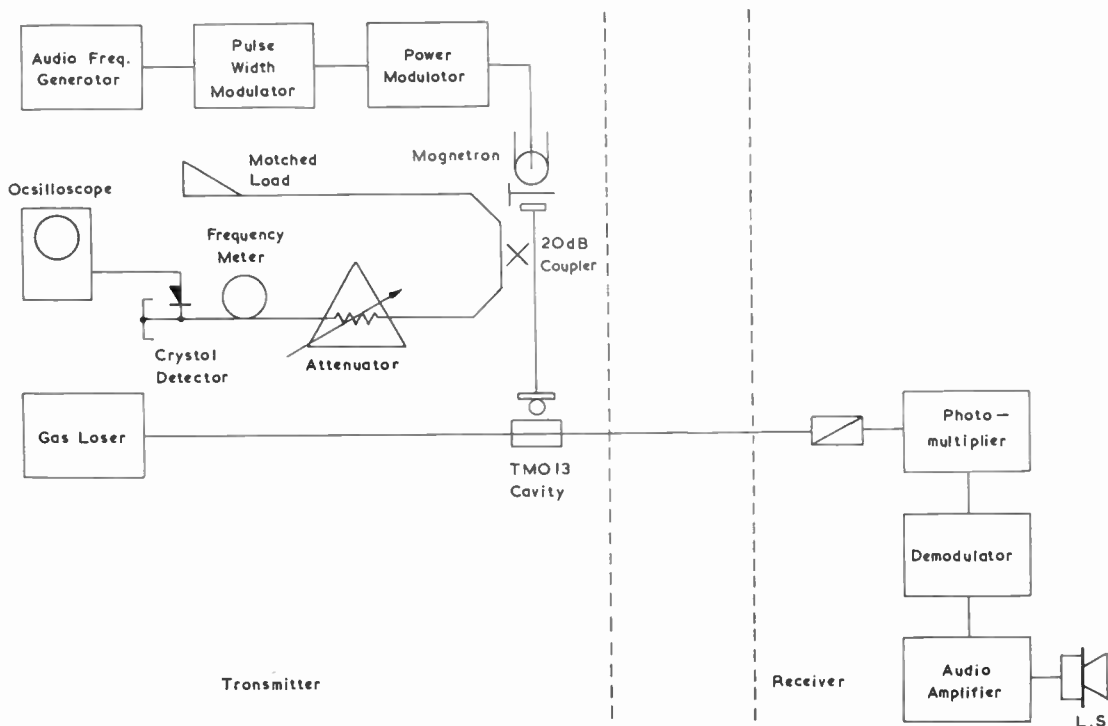
Fig. 1. Block diagram of the transmitter and the receiver.
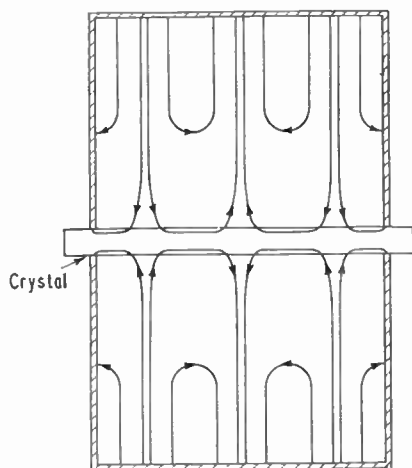
Fig. 2. Microwave cavity.

propagation, and its X- or Y-axis in the plane of the electric vector of the light, it may be shown[6, 7] that the signal emerging from the analysing prism contains only the odd sidebands of the modulation, the carrier and even sidebands being suppressed. (This corresponds to the analysis given by Mézec *et al.*[7] when the dephasing $\Delta\phi_0$ due to the compensator is zero.) The

photomultiplier output is then applied to the circuit shown in Fig. 4; the diodes D1 and D2 limit the amplitude of the signal applied to the base of TR10, and demodulation is achieved by the R-C network (39 kΩ/0·002 μF). A low-pass filter removes the pulse repetition frequency and harmonics, and the remaining information signal is amplified and fed to a loudspeaker.

4. Conclusions

With a modulation index of 0·05 and a path-length of a few metres the overall signal/noise ratio is better than 20 dB. If the beam were to be propagated over a useful distance through the atmosphere this signal/noise ratio would be considerably reduced by turbulence, absorption, etc. For a practical communications system it would, therefore, be desirable to use a light guide or a pipe containing a controlled atmosphere as a more efficient transmission medium. It would also be an advantage if a larger modulation index could be achieved, and further investigations into methods of increasing this are in hand.

5. Acknowledgment
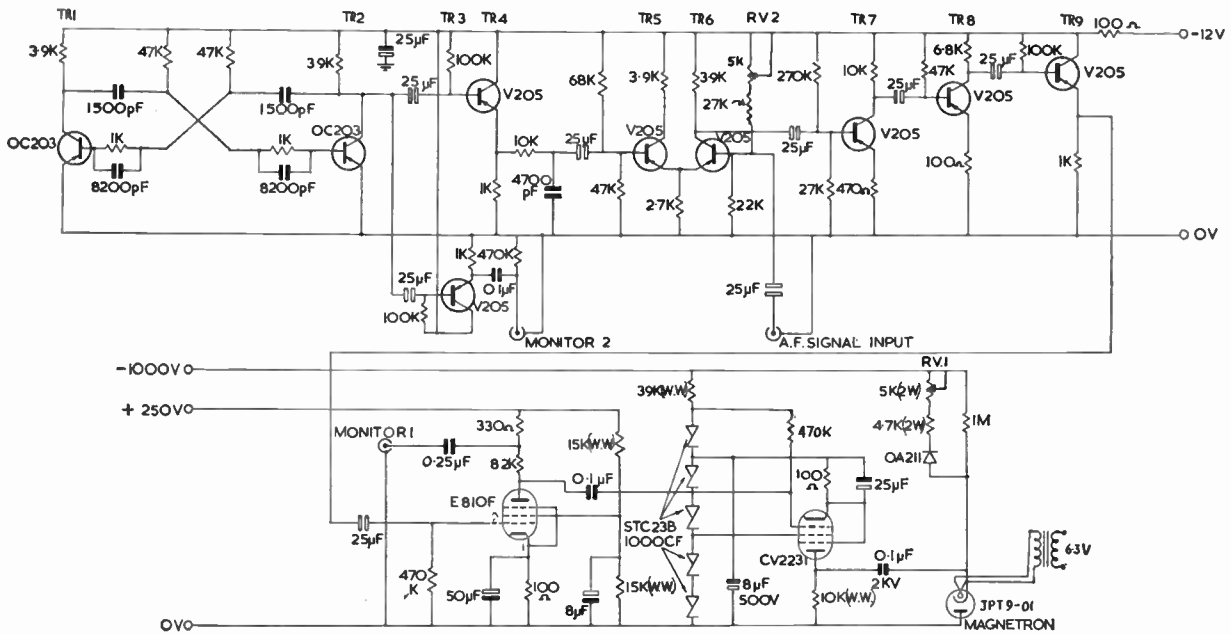
Fig. 3. Circuit diagram of the pulse width modulator.

(*Note:* The resistor between the base of TR6 and the lower end of RV2 should read 27kΩ.)
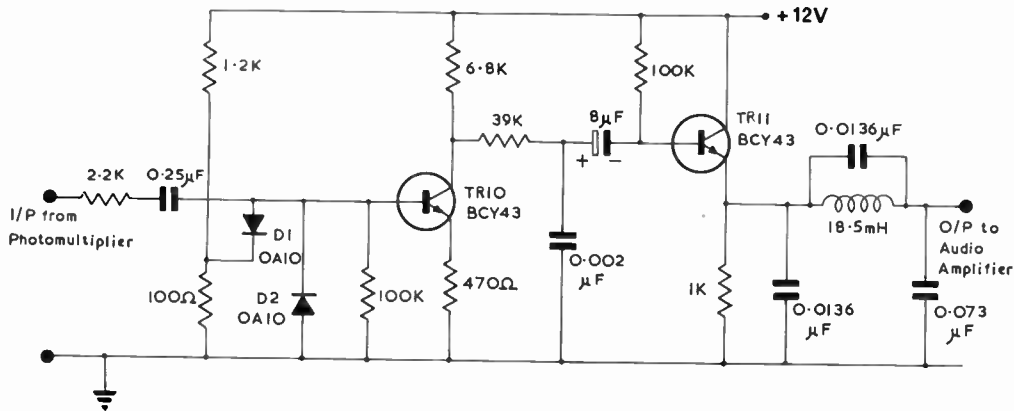
Fig. 4. Circuit diagram of the pulse width demodulator.

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A New Sound Velocity Meter

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Summary: The design of a new instrument for measuring the velocity of sound in sea water with improved accuracy is described. A particular feature of the instrument is that it may be towed through the water without causing appreciable errors.

1. Introduction

The standard technique for measuring the velocity of sound in a liquid is to make use of the 'sing-around' principle (see Fig. 1) in which a short pulse of sound is transmitted over a precisely known path. When the pulse is received it immediately re-triggers the transmitter so that the pulse repetition frequency (p.r.f.) of the system is given by:

$$f = \frac{1}{T} = \frac{C}{L} \qquad \text{......(1)}$$

where T is the travel time of the pulse, L is the path length and C is the velocity of sound.

If the liquid is moving past the instrument with a component of velocity v parallel with the instrument as shown in Fig. 1(a) there is, of course, an error in the measurement since eqn. (1) is modified to:

$$f = \frac{1}{T} = \frac{C+v}{L} = \frac{C}{L}\left(1 + \frac{v}{C}\right) \qquad \text{......(2)}$$

In several instruments which are commercially available[1] this error is reduced by folding the path as shown in Fig. 1(b) so that, if the flow is smooth, and if the angle ϕ is small, eqn. (2) becomes:

$$f = \frac{1}{T} = \frac{1}{\dfrac{L}{2(C+v)} + \dfrac{L}{2(C-v)}} \simeq \frac{C}{L}\left(1 - \frac{v^2}{C^2}\right) \ \text{......(3)}$$

However, in practical cases, the flow is turbulent, v varies along the track of the pulse and the errors can be considerably higher than eqn. (3) would indicate.

This new instrument uses the arrangement shown in Fig. 1(c) where the go and return paths are coincident so that first-order errors are cancelled, whether the flow is turbulent or not, and the maximum error which can occur is indicated by eqn. (3) with v now representing the average velocity along the path.

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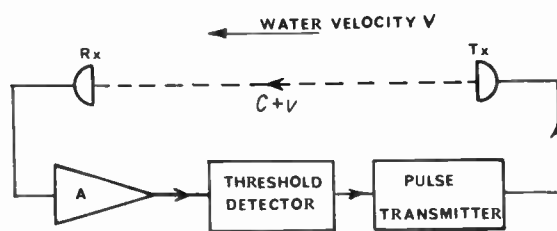
(a) With a straight path: sensitive to flow.
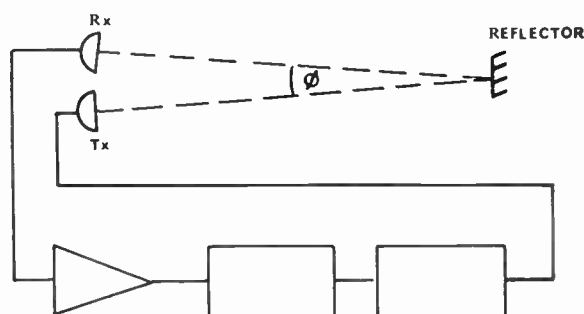
(b) With a folded path: insensitive to smooth flow.
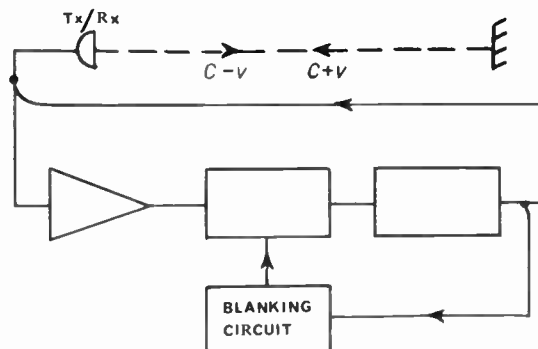
(c) With a single transmit/receive transducer: insensitive to turbulent flow.

Fig. 1. Sing-around velocity meter systems.
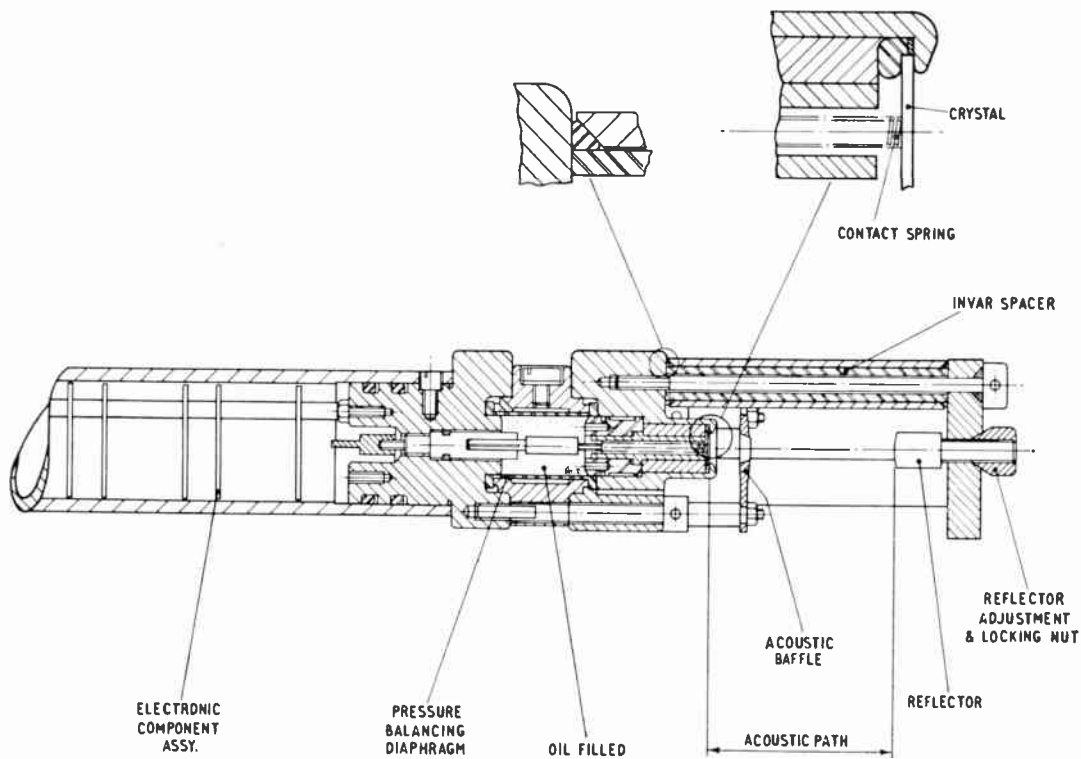
Fig. 2. The measuring head.

For $v = 25$ ft/s (7·62 m/s, 15 knots) and $C = 5000$ ft/s (1524 m/s) this maximum error is about 1 in 40 000 so the design aim has been to limit all other sources of error to comparable values.

2. Mechanical Design

The layout of the instrument head which consists of a housing for the transmit/receive transducer, a reflector, and three spacing bars to define their separation is shown in Fig. 2. The electronic circuits are housed in a stainless steel tube coaxial with the instrument head so that the whole instrument is 23 in (58·4 cm) long and 2 inches (5·08 cm) in diameter.

The operating temperature range is $-4°C$ to $+30°C$ and the pressure range 0 to 5000 lb/in² (350 kg/cm²) (although the head itself has been tested to 10 000 lb/in² or 700 kg/cm²) so that the problems of mechanical design are threefold:

(a) To ensure that the path length is independent of temperature.

(b) To ensure that it is independent of pressure.

(c) To arrange that the only significant acoustic path is the direct one of transducer—reflector—transducer and that interfering secondary signals are kept to a minimum. (Since the secondary signals combine with wanted signal in a

very uncertain phase relationship their effect is to make the system hypersensitive to small changes in signal amplitude or in sound velocity.)

The spacers are constructed from invar which has a temperature coefficient of expansion of only 4 parts in 10⁶ per deg C and both the reflector and the transmit receive transducer are mounted on stainless steel stalks whose coefficient is 20 parts in 10⁶ per deg C. Since, as in a 'grid-iron' pendulum, the expansion is in opposite directions and since the lengths of the spacers and the stalks are arranged to be in a 5 : 1 ratio the overall temperature coefficient of the path length is nominally zero and, in practice, can be relied on to be less than 1 part in 10⁶ per deg C which is not significant. The invar spacers are fitted inside oil-filled stainless steel tubes with 'O' ring seals in order to minimize corrosion.

Changes of static pressure will distort the position of the transducer relative to the reflector unless a pressure balancing mounting technique is used and in this instrument this has been done by filling the rear of the transducer assembly with oil which is at ambient pressure. The transducer itself is a PZT4 ceramic disk, 10 mm in diameter and 0·4 mm thick, which is platinum plated on the surface exposed to the seawater and silver plated on the other, electrical

WITHOUT BAFFLE

WITH BAFFLE
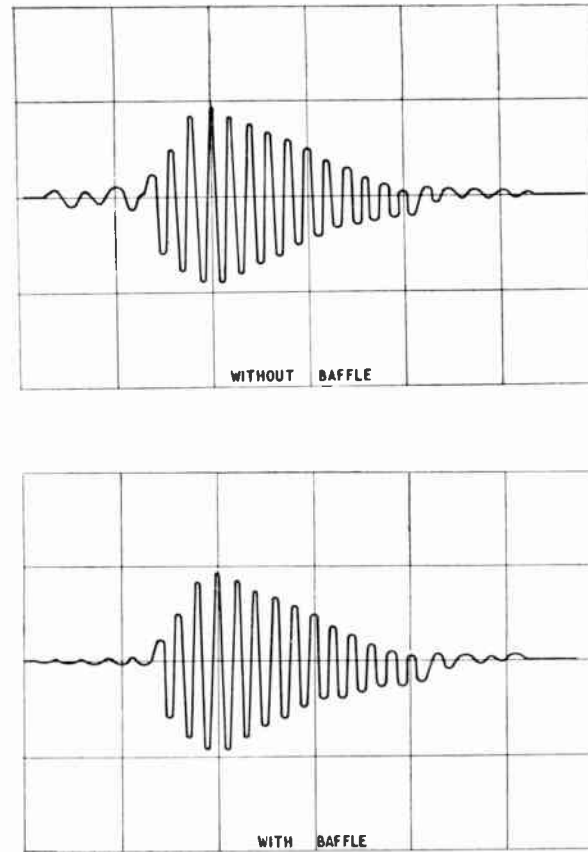
Fig. 3. Acoustic baffle effect.

connections being made via the carefully machined support rim (which is part of the main body of the instrument and is at ground potential) and via a coil spring which makes a pressure contact on the silver-plated surface. The disk is kept in position against its support rim by compressing an O-ring whose mean diameter is the same as the diameter of the disk and the O-ring (which acts both as a seal and as a locating mechanism) is compressed by a screwed tube which is fitted with an inner insulating sleeve. The pressure balance oil reservoir consists of a flanged neoprene rubber tube which has sea-water at ambient pressure on its outer surface. The sea-water has access via holes drilled in the front face of the trans-ducer assembly so that both static pressures at depth and dynamic pressures which could arise when the instrument is towed through the water are balanced. The effect of the whole arrangement is that changes in neither pressure nor temperature can exert any force tending to distort the transducer.

The reflector is mounted on an adjusting screw so that fine adjustments to the path length can be made and it has an accurately ground spherical surface with a radius of 1 inch so that difficulties of alignment are

minimized. The curvature also helps to reduce the amplitude of secondary signal which has been reflected twice from the reflector and once from the crystal. A further source of unwanted signal is from the front face of the support rim which emits energy slightly in advance of the main beam because the velocity of sound in the stainless steel of the support rim is higher than it is in sea-water. The amplitude of this inter-fering signal is reduced by the acoustic baffle which has an aperture smaller than the effective diameter of the disk. The baffle is so constructed that it has no surface parallel to the plane of the disk so that the unwanted part of the signal is reflected out of the system. The effect which the baffle has is illustrated in Fig. 3. (This picture is, of course, that obtained under open-loop conditions. Under normal conditions with the loop closed all but the first fraction of the received signal is smothered by the next transmitter pulse.)

3. Electronic Design

Equations (1) to (3) are not exact since there is a delay t between the receipt of a signal and the next transmission. The more exact equation for still water is therefore:

$$f = \frac{1}{T+t} \simeq \frac{C}{L}\left(1 - \frac{Ct}{L}\right) \qquad \ldots\ldots(4)$$

The practical implications of this are twofold. Firstly the output frequency is not exactly proportional to sound velocity so that if no correction for this is made a systematic error will result. It is consequently important to make t as small as possible. Secondly, variations in t due for instance to changes in tem-perature or in signal amplitude will give rise to errors so it is also important to make t as constant as possible.

In this instrument T is about 100 μs and t is of the order of 100 ns so that systematic error is about ± 1 part in 16 000 over the velocity range 4700 ft/s to 5000 ft/s (or 1432–1524 m/s). (The instrument is initially set up near the middle of this range.) The error can be reduced by a factor of ten or more if a nominal correction is made and, of course, can be eliminated almost entirely by carrying out two cali-brations at widely differing velocities and constructing a calibration curve.

The total delay of about 100 ns is made up of several components which with only slight oversimplification may be listed as follows:

(i) A delay of about 10 ns occurs from the time at which the received signal exceeds the trigger level to the instant where the maximum trans-mitting voltage is applied to the transducer. The delay depends on the switching times of the semiconductors used and is consequently tem-perature dependent by about $\pm 10\%$ over the
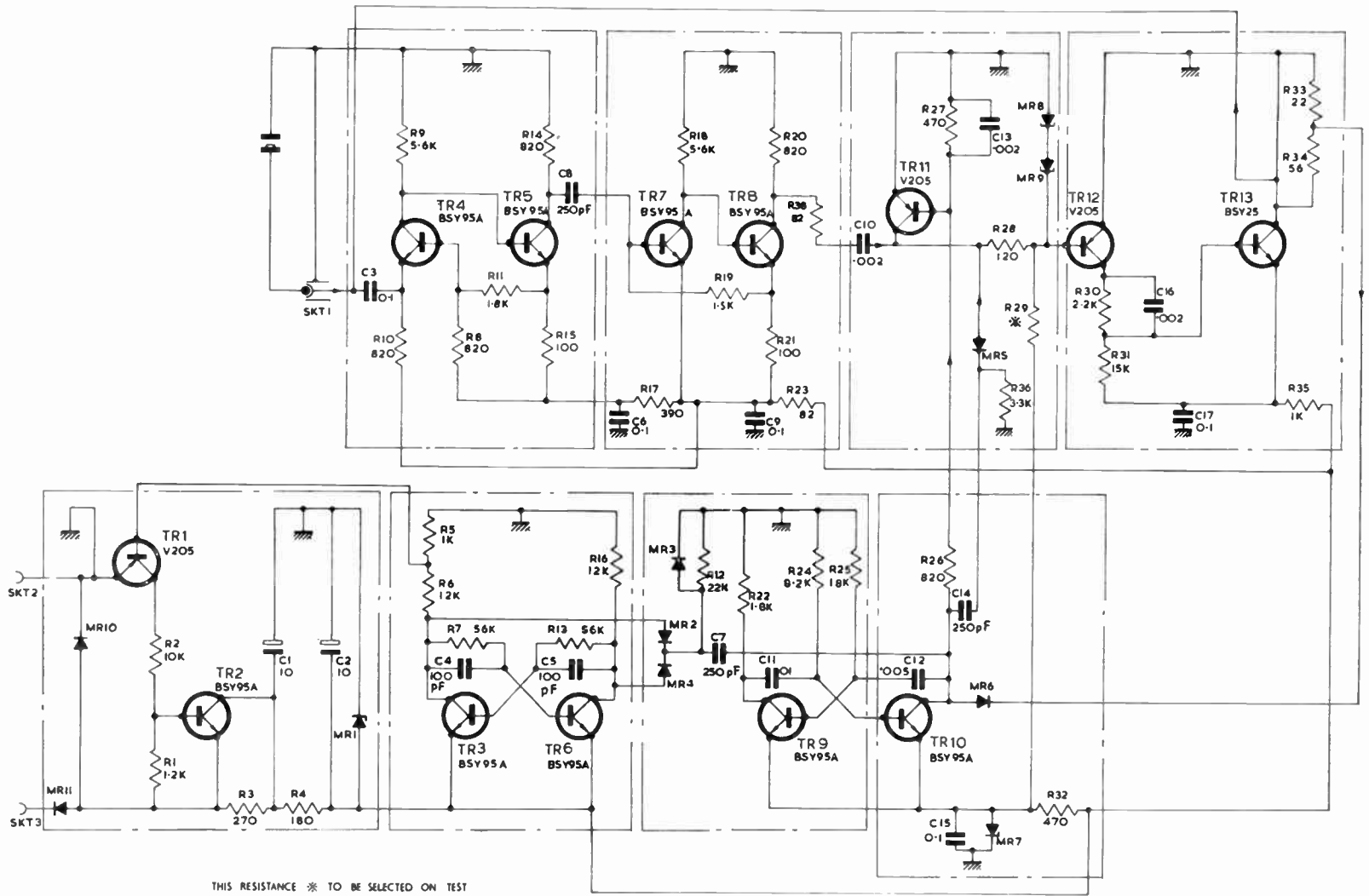
Fig. 4. Circuit diagram of the sound velocity meter type MO30.

THIS RESISTANCE ✳ TO BE SELECTED ON TEST

working range so the resulting error in measured sound velocity is only ± 1 part in 10^5.

(ii) A delay of about 50 ns occurs between the maximum voltage being applied to the transducer and the maximum strain appearing at its surface. The delay is relatively constant in value since it depends only on the thickness of the transducer disk and the velocity of sound in it. A similar delay occurs on reception.

(iii) There is a delay of about 23 ns in the receiver.

(iv) The above delays are reduced by about 33 ns since the threshold of the triggering circuit is set at about half the peak of the first cycle of the received signal.

The last delay can obviously be altered by a change in the triggering level, in the transmitter power, or in receiver gain, the effect being small if the carrier frequency of the pulse is high. At the chosen carrier frequency of about 5 MHz (which is about as high as can conveniently be obtained with a robust electrostrictive crystal) a change in triggering level of 3% is equivalent to a change in circuit delay of approximately 1 ns. To achieve this order of stability the receiver amplifiers have heavy negative feedback, metal oxide resistors are used to define circuit gain, supply rails are stabilized with Zener diodes and the triggering threshold itself is set by the peak current of the tunnel diodes which varies by less than 1% over the operating temperature range.

The delay can, of course, also be altered by changes in the amplitude of the received signal which could arise from a dependence on pressure or temperature of transducer sensitivity, reflector alignment, or attenuation in the water path. The stress-free method of mounting the transducer and the use of a spherical reflector have, in this instrument, eliminated the first two effects and any variations in attenuation due to pressure changes are too small to be measurable. However, changes of attenuation with temperature are probably just significant since at a frequency of 5 MHz the attenuation at 20°C is about 5 dB per yard which implies a loss over the path length of the instrument of less than 1 dB. Little information on the temperature dependence of this loss is available but if, as seems reasonable, it can vary by no more than $\pm\frac{1}{2}$ dB over the temperature range then the maximum error introduced by this mechanism is less than ± 1 part in 50 000.

4. Circuit Details

The circuit diagram of the instrument is shown in Fig. 4.

The receiver amplifier consists of two feedback pair amplifiers (TR4 and 5 and TR7 and 8) which are identical except for the fact that in the first pair the signal is injected on the emitter of TR4 instead of the base in order to obtain a polarity inversion. Each pair has an open-loop current gain of about 60 dB and a closed-loop gain of 26 dB.

The transmitter which consists of TR13 driven by TR12 is capable of delivering a peak current of about 1 amp. A very rapid rise-time is achieved by supplying a large excess base drive of almost 5 mA to TR12. This drive is switched on when the combined signal current (via C10 and R28) and preset bias current (via R29) exceeds the triggering current of the tunnel diodes MR8 and MR9. This current is nominally 5 mA and their switching time is about 2 ns. R29 which is the only adjustable component in the circuit is selected on test so that the triggering point is about half the peak signal amplitude. Components R30, R31 and C16 are introduced to ensure that the mean current under non-typical conditions (when, for example, R29 is being adjusted) cannot rise to destructive levels.

The astable multivibrator (TR9 and TR10) which is designed to free-run at a little less than the lowest possible sing-around frequency has three functions:

(a) To trigger the system into operation when first switched on. This it does by routing a negative going spike to the tunnel diodes via the differentiating circuit C14, R36 and MR5.

(b) To prevent the reverberations which occur immediately after the transmitter pulse from falsely re-triggering the transmitter. This it does by clamping the signal line to earth for about the first two-thirds of the reception period via R26 and TR11.

(c) To provide an output signal via C7 to the divide by 2 bistable (TR3 and TR6).

The multivibrator is synchronized to the sing-around frequency by a fraction of the transmitter pulse which is injected via MR6 so that function (a) occurs only when no sing-around signal is present.

The instrument is designed to operate with any length of single-core cable between 0 and 2000 m. The surface unit (which uses no novel techniques and is therefore not described in detail in this article) operates from either 90 V to 250 V a.c., 50 to 400 Hz or from 11 to 30 V d.c. and delivers a constant current supply stabilized at 45 mA. Within the instrument, the supply is smoothed by C1, C2 and R4 and is stabilized to 15 V (positive earth) with the Zener diode MR1. The artifice of using a constant current supply enables MR1 to work at a constant power level which does not depend on the length or resistance of the cable. Diodes MR10 and MR11 prevent damage which might otherwise be caused by attempting to use incorrect power supplies.

The output signal is, because of the 'divide by two' action of the bistable consisting of TR3 and TR6, an exactly symmetrical square wave whose frequency is one half the sing-around frequency. This system has the merits that the output signal can be made exactly equal to the speed of sound in feet per second with a convenient crystal-reflector separation of 3 in and that, being symmetrical, the output signal has no even harmonic content and a maximum fundamental component. This output signal is used to square-wave-modulate the terminal voltage of the supply cable by intermittently shorting out R3 with TR2. The resulting signal is extracted at the surface unit by means of a current transformer, amplified and fed to any general purpose frequency counter.

5. Calibration and Testing

It would ideally be useful to carry out calibrations to an accuracy 1 in 100 000 which implies that the water temperature must be held steady to 0·004°C.[2] Attempts to obtain stability of this order from a thermostatic control were not successful and the method finally adopted was to use a large 2000-gallon (~9000 litres) tank situated in a temperature-controlled room. A small 5-gallon (22·5 litre) tank containing triple-distilled water in which the instrument was immersed was placed within the larger tank and the whole assembly allowed to stabilize at room temperature. This arrangement gave sets of readings which were stable over a period of hours to better than 1 in 100 000 and allowed checks to be made that variations in power supplies had no significant effect.

The calibration technique now adopted with production instruments uses the same tank. The temperature is read to an accuracy of ±0·02°C and the reflector is adjusted with its micrometer screw until the reading obtained agrees with that indicated by Wilson's tables for the measured temperature. The reflector is then locked into position with its special tapered lock nut and the instrument rechecked. It is considered that this provides an initial setting repeatability of 1 in 20 000 although the overall stability of the instrument may well be better than this.

In May 1965, a prototype instrument was taken aboard R.R.S. *Discovery* and, with the help of the National Institute of Oceanography, simultaneous readings were made of salinity (by a Nansen bottle and subsequent laboratory measurement), depth and temperature (with reversing thermometers) and sound velocity (with the prototype instrument). The measured values of sound velocity were then compared with ones obtained by calculation from Wilson's tables.[3] The results obtained are shown in Table 1. The average difference is about 1 in 10 000 which is considered to be within the experimental error since the measurement accuracies of temperature, depth and

salinity were estimated at ±0·02°C, ±1% and ±0·03% respectively, and these lead to uncertainties in the calculated velocity of about ±0·2 ft/s, ±0·6 ft/s and ±0·1 ft/s (0·05, 0·15 and 0·025 m/s respectively).

Table 1
Sea trials in R.R.S. *Discovery*

Depth in metres	Velocity meter feet/s	Calculated velocity feet/s	Discrepency feet/s
50	4910·6	4910·7	−0·1
200	4912·9	4913·7	−0·8
500	4926·7	4928·1	−1·4
1000	4936·4	4936·0	+0·1
1500	4916·4	4916·3	+0·1
			Average −0·4 ft/s

An instrument has also been mounted on the rotating arm of the circular tank of the Admiralty Research Laboratories, Teddington. The results obtained are shown in Fig. 5. (The version of the instrument used was one which indicates directly in metres per second, the separation of crystal and reflector being 5 cm and the 'divide by two' bistable omitted). Reliable readings with a scatter of less than 0·1 m/s were obtained at towing speeds of up to 36 ft/s or 10 m/s (22 knots) but above this speed the readings begin to fall suddenly. What causes this is as yet problematical but it is considered likely that it is due to the presence of small air bubbles induced by cavitation. If this is so, higher towing speeds will probably be satisfactory at greater depths.
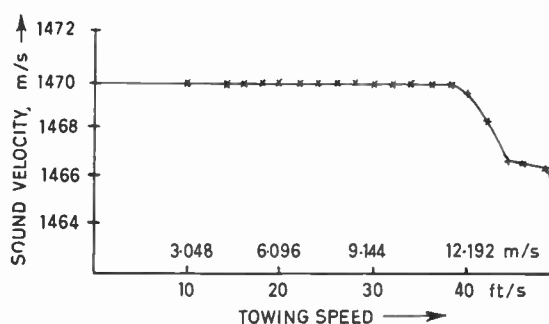
Fig. 5. Graph of sound velocity versus towing speed.

From the overall results of the test programme it has been concluded that the instruments have a long-term overall accuracy of better than 1 in 10 000 within the temperature range of −4°C to +30°C and for depths of up to 2000 m with towing speeds of up to 22 knots. Attempts to devise a testing technique

which is capable of unambiguously indicating any spurious short-term fluctuations have, so far, proved to be unsuccessful. In still water at room temperature any such fluctuations are less than 1 part in 10^5.

6. Derivatives

Two versions of the instrument have been developed: the one described here which indicates velocity in feet per second and another which indicates metres per second. A third version which incorporates a pressure transducer and a fourth which will have full ocean depth capability are under development.

A subsidiary application of the instrument has been discovered to be the detection of the presence of very small air bubbles which have a marked effect on sound velocity.

7. Acknowledgments

The authors wish to thank the Directors of The Plessey Company Limited for permission to publish this paper and to thank the staff of the National Institute of Oceanography and the Admiralty Research Laboratories for the advice and help which they have willingly given in testing the instrument. They also wish to acknowledge the important contribution made by Mr. R. C. Chesters, now with Rank-Pullin Ltd., during the early stages of development.

8. References

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2. V. A. Del Grosso, 'Problems in the absolute measurement of sound speed', *U.S. Navy J. Underwater Acoustics*, October 1966.
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STANDARD FREQUENCY TRANSMISSIONS

(Communication from the National Physical Laboratory)

Deviations, in parts in 10^{10}, from nominal frequency for May 1967

May 1967	24-hour mean centred on 0300 U.T.			May 1967	24-hour mean centred on 0300 U.T.		
	GBR 16 kHz	MSF 60 kHz	Droitwich 200 kHz		GBR 16 kHz	MSF 60 kHz	Droitwich 200 kHz
1	− 299·8	+ 0·4	− 0·4	17	− 300·1	− 0·1	− 0·7
2	—	0	− 0·4	18	− 300·0	+ 0·2	− 0·8
3	− 300·1	− 0·1	+ 0·5	19	− 300·1	− 0·2	− 0·5
4	− 300·0	− 0·1	0	20	− 299·8	+ 0·2	− 1·0
5	− 300·0	0	− 0·3	21	− 299·9	+ 0·2	− 0·9
6	− 299·9	0	− 0·8	22	− 299·9	+ 0·1	− 1·0
7	− 300·0	+ 0·1	− 1·1	23	− 299·9	+ 0·2	− 1·2
8	− 299·9	0	− 1·0	24	− 300·0	0	− 1·3
9	− 299·9	+ 0·1	− 1·0	25	− 299·6	+ 0·3	− 1·5
10	− 300·0	+ 0·1	− 0·9	26	− 299·9	+ 0·1	− 1·2
11	− 300·0	− 0·2	− 0·7	27	− 299·8	+ 0·1	− 0·7
12	− 299·9	0	− 0·9	28	− 299·8	+ 0·1	− 0·2
13	− 300·0	0	− 0·4	29	− 299·8	+ 0·3	− 0·3
14	—	—	—	30	− 299·7	+ 0·3	+ 0·1
15	—	—	—	31	− 299·6	+ 0·3	+ 0·1
16	− 300·0	− 0·1	− 0·2				

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

Conversion of a PAL Colour-Television Signal into a Symmetrical Dot-sequential Signal

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Summary: An N.T.S.C.-type colour-television signal can be converted into a driving signal for a one-gun display tube without full demodulation of the sub-carrier signal. With the adoption of the PAL system for colour-television broadcasts in many countries of Europe the question arises whether similar methods can be devised for transcoding a PAL signal into a driving signal for dot-sequential display systems. The analysis of this problem leads to the conclusion that a slight modification of the PAL switching axis or a slight modification of the composition of the luminance signal is desirable. Practical conversion methods are discussed and some experiments are described. Employment of the standard PAL signal instead of the modified signal introduces small errors which are hardly observable. It is concluded that conversion of the PAL chrominance signal into a driving signal for one-gun display systems is well within the possibilities and does not require complex circuitry.

1. Introduction

In present-day colour-television receivers usually three-gun display tubes are employed, which have to be driven either by colour difference signals and the luminance signal or by primary colour signals. The methods for deriving these signals from the composite colour-television signal are well known and need not be repeated here. Thanks to the presence of three electron beams the red, green and blue colour phosphors can be excited simultaneously.

With three-gun tubes, in particular with the three-gun shadow-mask tube, very good picture reproduction can be obtained. However, there are some inconveniences. The most important among these is the necessity for geometrical coincidence of the three primary colour pictures, which can only be obtained at the cost of considerable circuit complexity.

From the early beginnings of colour television, efforts have been made to construct display systems employing only one electron beam, which strikes the primary colour phosphors sequentially. Among the methods which have been put forward, those that employ colour switching in dot-sequential rhythm have received particular attention. In spite of these efforts, the shadow-mask tube is still the only display tube that is in widespread practical use; however, there is also renewed interest in sequential displays, in particular for applications in small colour television sets.

2. Analysis of PAL-C.C.S. Conversion

2.1. *Formulation of the Problem*

Loughlin[1] has shown that two modes of dot-sequential operation can be distinguished, commonly

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denoted by r.c.s. (reverse colour sequence) and c.c.s. (continuous colour sequence). In r.c.s. operation the colour switching sequence alternates, e.g. according to the scheme RGB—BGR—RGB, etc. In c.c.s. operation there is a fixed colour sequence, thus RGB—RGB, etc. Loughlin has also shown that conversion of the N.T.S.C. signal into the c.c.s. driving signal is relatively easy, thanks to the analogy between the c.c.s. driving signal and the standard N.T.S.C. signal. This conversion does not require full demodulation of the composite colour signal.

Recently, most countries in Western Europe have adopted the PAL colour-television system as the transmission system to be employed for colour broadcasts. In view of this decision it seems interesting to investigate the possibilities of using similar techniques for forming a c.c.s. signal from the standard PAL signal.

In the PAL system[2,3] the modulation of the colour sub-carrier is different in successive scanning lines. During successive lines the signal can be written respectively as

$$E_c = Y + \frac{1}{1\cdot14}(R - Y)\cos\omega t + \frac{1}{2\cdot03}(B - Y)\sin\omega t \tag{1}$$

and

$$E_c = Y - \frac{1}{1\cdot14}(R - Y)\cos\omega t + \frac{1}{2\cdot03}(B - Y)\sin\omega t \tag{2}$$

where E_c denotes the composite colour-television signal, Y denotes the standard luminance signal, R and B denote gamma-corrected primary colour signals and ω is the angular frequency of the sub-carrier.

From eqns. (1) and (2) one sees that the PAL signal

during one half of the scanning lines is identical to the N.T.S.C. signal; during the other half the phase of the $(R-Y)$ component is reversed. For convenience we shall name the former lines odd lines, the latter ones even lines.

Figure 1 shows a phasor diagram of the PAL chrominance signal in successive lines. We shall call the axis of phase reversal the switching-axis of the signal. Originally a PAL system was proposed by Bruch in which the switching-axis was not the $(R-Y)$ axis but the I axis of the N.T.S.C. signal.[4] If we want to formulate the PAL signal in more general terms, not specifying in advance the switching-axis,

we can rewrite eqns. (1) and (2) as

$$E_c = Y + \frac{1}{1 \cdot 14}(R-Y)\cos\omega t + \frac{1}{2 \cdot 03}(B-Y)\sin\omega t \quad \ldots\ldots(3)$$

during odd lines and

$$E_o = Y + \frac{1}{1 \cdot 14}(R-Y)\cos(\omega t - 2\theta) +$$
$$+ \frac{1}{2 \cdot 03}(B-Y)\sin(\omega t - 2\theta - 180°) \ldots\ldots(4)$$

during even lines.
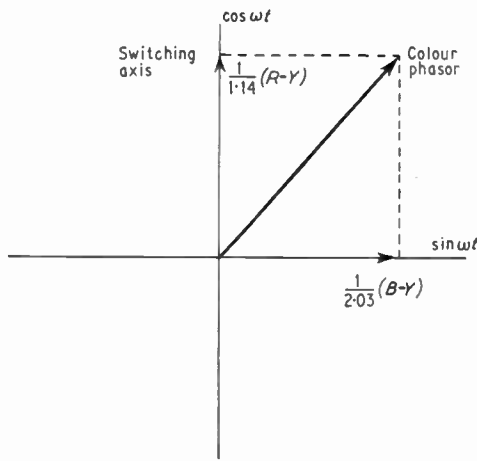
(a) Odd lines.
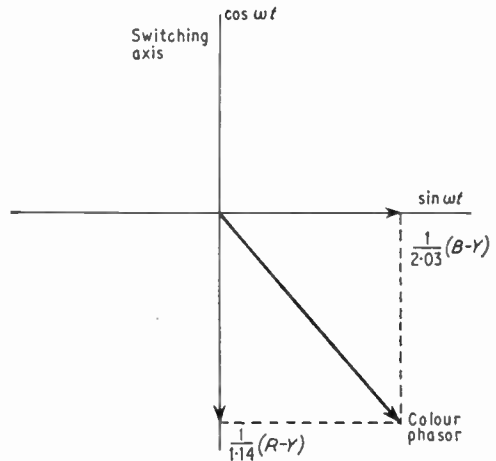
(b) Even lines.

Fig. 1. Phasor representation of the PAL chrominance signal.
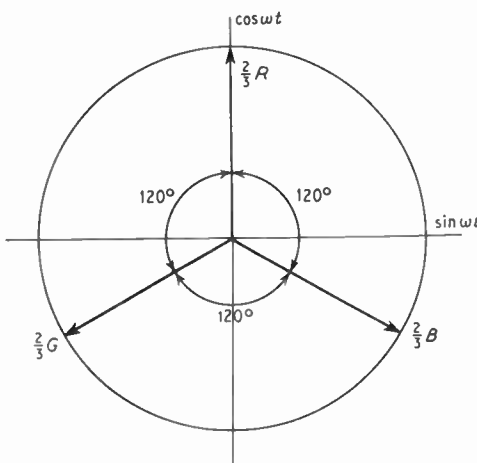
(a) Symmetrical representation.
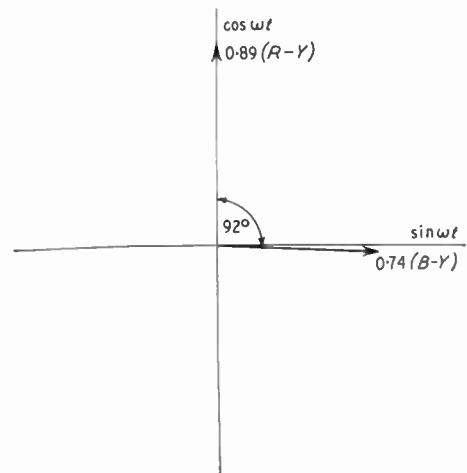
(b) Representation showing colour-difference signals.

Fig. 2. C.C.S. dot-sequential chrominance signal.

According to this notation, in the original PAL system θ was 57°, while the PAL system that has now been adopted employs $\theta = 90°$. For our analysis we shall maintain θ as a free system parameter.

The symmetrical c.c.s. signal can be written as

$$E_w = \tfrac{1}{3}(R+B+G) + \tfrac{2}{3}[R\cos\omega t + B\cos(\omega t - 2\pi/3) + \\ + G\cos(\omega t - 4\pi/3)] \quad\ldots\ldots(5)$$

where ω is the colour writing-frequency. For convenience we shall assume that the writing frequency equals the sub-carrier frequency. Putting $\tfrac{1}{3}(R+B+G) = M$ and applying simple trigonometric transformations we can rewrite eqn. (5) as

$$E_w = M + 0.89(R-Y)\cos\omega t + 0.74(B-Y)\sin(\omega t - 2°) \quad\ldots\ldots(6)$$

A static phase shift of 19°, which does not alter the signal essentially, is omitted in the formula. Figure 2 shows phasor diagrams of the c.c.s. chrominance signal according to eqns. (5) and (6).

2.2. Mathematical Analysis of the Signal Conversion

Loughlin has shown that the N.T.S.C. signal can be transformed into the c.c.s. signal by applying two operations. First the luminance signal Y has to be transformed into the 'monochrome' signal M by means of a 'Y to M converter'.

This is done by adding to Y a correction signal $M - Y$, which can be derived from the chrominance signal by synchronous detection. The second operation consists in transforming the N.T.S.C. chrominance signal into the c.c.s. chrominance signal. This can be done by employing a so-called 'elliptical-gain amplifier'. This is essentially a mixer circuit whose operation can mathematically be described as multiplication of the N.T.S.C. input signal by

$$k[1 + 2m\cos(2\omega t + \xi)] \quad\ldots\ldots(7)$$

where k, m and ξ are constants which have to be appropriately chosen. For the sake of completeness it must be mentioned that the term with frequency 3ω, which originates also in the above-mentioned multiplicative operation, is assumed to be eliminated by means of a filter.

It can easily be seen that multiplying the chrominance signal by eqn. (7) can only produce the c.c.s. chrominance signal if a certain static phase shift is accepted. Hence, we write the c.c.s. chrominance signal more generally as

$$E_{cc} = 0.89(R-Y)\cos(\omega t + \delta) + \\ + 0.74(B-Y)\sin(\omega t + \delta - 2°) \quad\ldots\ldots(8)$$

In the case of the PAL system the input chrominance signal is different during odd and even lines. Hence, the conversion parameters k, m and ξ have also to be different in alternating lines. In addition, the

parameter δ will change from line to line. It hardly needs to be said that this change of δ is very inconvenient, since it requires line-sequential phase shifting of the output signal of the converter, which is a modulated signal.

To avoid this difficulty, we shall now look for a solution, giving equal values of δ in odd and even lines. It is here that we need the freedom in choice of θ.

To simplify the notation we shall use the subscript 'o' for odd lines and 'e' for even lines. In addition, we define A, B, C and D as follows:

$$\frac{1}{1\cdot14}(R-Y) = A, \quad\ldots\ldots(9)$$

$$\frac{1}{2\cdot03}(B-Y) = B, \quad\ldots\ldots(10)$$

$$0\cdot89(R-Y) = C \quad\ldots\ldots(11)$$

and $\quad 0\cdot74(B-Y) = D. \quad\ldots\ldots(12)$

By doing so we obtain the following identities:

(a) for odd lines

$$k_o(A\cos\omega t + B\sin\omega t)[1 + 2m_o\cos(2\omega t + \xi_o)] \\ \equiv C\cos(\omega t + \delta) + D\sin(\omega t + \delta - 2°) \quad\ldots\ldots(13)$$

(b) for even lines

$$k_e[(A\cos2\theta + B\sin2\theta)\cos\omega t + \\ + (A\sin2\theta - B\cos2\theta)\sin\omega t] \times \\ \times [1 + 2m_e\cos(2\omega t + \xi_e)] \\ \equiv C\cos(\omega t + \delta) + D\sin(\omega t + \delta - 2°) \quad\ldots\ldots(14)$$

These identities lead to two essentially different sets of conditions, which are characterized by the values of the parameters presented below.

First set	Second set
$\delta = 1\cdot2°$	$\delta = 1\cdot2°$
$\xi_o = \xi_e = 175\cdot2°$	$\xi_o = \xi_e = 175\cdot2°$
$\theta = 92\cdot4°$	$\theta = 2\cdot4°$
$m_o = 0\cdot199$	$m_o = 0\cdot199$
$m_e = 5\cdot02$	$m_e = 5\cdot02$
$k_o = 1\cdot26$	$k_o = 1\cdot26$
$k_e = 0\cdot25$	$k_e = -0\cdot25$

Comparison of the two sets of conditions shows that in the first set k_o and k_e have equal signs, while in the second one k_o and k_e have opposite signs. For this reason the first set is the more practical one and should be preferred. In addition it should be noted that the signal composition according to the first set is very near to the adopted PAL signal, in which

(a) Odd lines.   (b) Even lines.
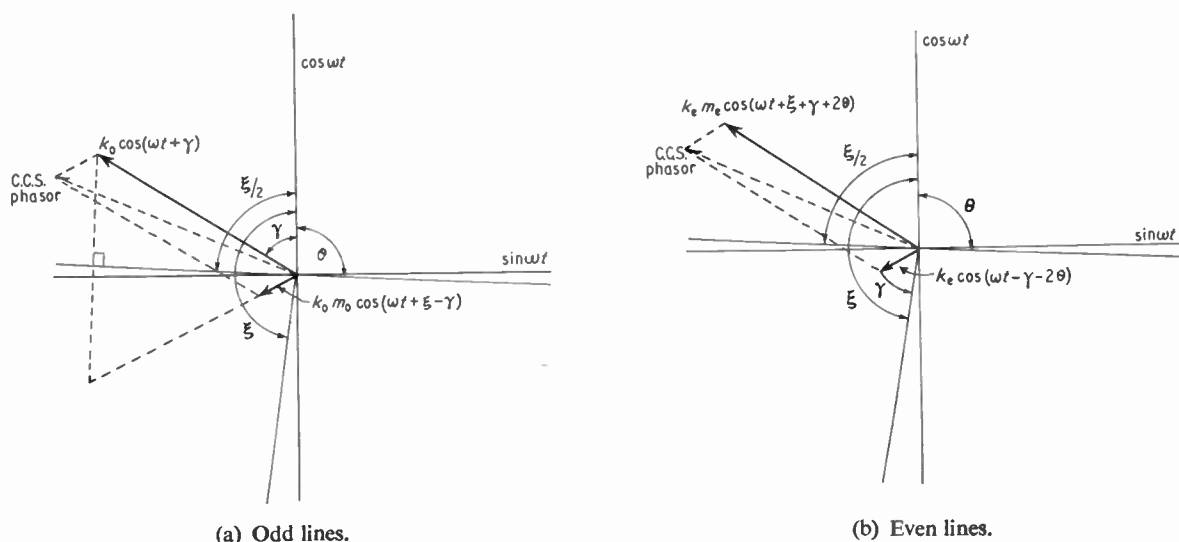
Fig. 3. Construction of c.c.s. phasor.

$\theta = 90°$. The difference is so slight that it seems reasonable to assume that even without modifying the PAL standard, so that $\theta = 92·4°$, the conversion method would work well.

2.3. *A Simple Phasor Representation of the Conversion Process*

The foregoing analysis offered formal proof that accurate transformation of the PAL signal into a c.c.s. signal is possible if a slight shift of the 'switching-axis' is accepted. To acquire a better insight into the background of this transformation, we shall consider it now in a different way, lending itself better to simple visualization.

Multiplication of the chrominance part of eqns. (3) and (4) by expression (7) yields two terms. One of these is the unchanged original signal, except for the amplitude factor k. In a phasor diagram of the chrominance signal the other can be represented by the original signal, mirrored by the axis corresponding to $\cos(\omega t + \xi/2)$. Its amplitude factor is km. For odd lines this is illustrated in Fig. 3(a). In this figure $\cos(\omega t + \gamma)$ represents the original chrominance phasor, while $km\cos(\omega t + \xi - \gamma)$ represents the reflected phasor. The addition of both yields the c.c.s. phasor.

Obviously the value of ξ is fixed by the requirements for obtaining the correct c.c.s. phasor in odd lines. In even lines the original signal is the mirror image of the odd-line signal with the axis $\cos(\omega t - \theta)$ as the mirror axis. We can now arrange things so that this signal is equivalent to the converted signal component in odd lines, while at the same time the converted signal in even lines is equivalent to the original signal in odd lines. Obviously this requires $\theta = -\xi/2 + 180°$ while the amplitude factors of the

original and converted signal components have to be interchanged. Hence $k_o = k_e m_e$ and $k_e = k_o m_o$. This is illustrated in Fig. 3(b).

3. Alternative Method of Modifying the PAL Signal

It has been shown that for exact transformation into c.c.s. to be possible, the PAL signal has to undergo a minor modification, namely, the switching-axis has to be shifted by $2·4°$. Modifying the PAL standard accordingly would slightly complicate decoding techniques for three-gun tubes. In such decoding it is profitable when the switching-axis coincides with the $(R - Y)$ axis. Hence, a choice has to be made between optimum adaptation to c.c.s. conversion or optimum adaptation to three-gun decoding.

However, it must be said that there is so little difference between the signal wanted for c.c.s. conversion and the standard signal that distinguishing between them is hardly within the tolerances of normal signal-processing circuitry.

Nevertheless it is interesting to investigate whether it is possible to design a PAL signal which avoids this dilemma. Obviously this requires that exact c.c.s. conversion be possible with $\theta = 90°$. Essentially this means that in the c.c.s. chrominance signal $(R - Y)$ and $(B - Y)$ should be in quadrature. This can be obtained by a slight modification of the luminance signal.

The problem of obtaining a c.c.s. signal having $(R - Y)$ and $(B - Y)$ in quadrature has been investigated by Wirth,[5] who found that modification of the luminance signal to $Y = 0·30R + 0·57G + 0·13B$ leads to the desired composition of the chrominance signal. In this case, the value of θ for optimum adaptation to c.c.s. conversion is 90°, although a slight aberration

from constant-luminance operation has to be accepted. However, in view of usual practice in gamma correction, this does not seem to be of much consequence. The proposed change might even lead to better monochrome reproduction of a colour television signal, since the large luminance error in saturated blue, introduced by the usual method of gamma correction, is somewhat reduced.

4. Comparison of Practical PAL-C.C.S. Conversion Circuits

Since the signal processing for odd and even lines is different, the employment of line-sequential switch-

ing cannot be avoided in the PAL-c.c.s. transcoder. The position of the switch is very important, since it determines the complexity of the other circuitry to a large extent. The most obvious method is depicted in Fig. 4. The main objection to this method is the necessity of two elliptical-gain amplifiers, one of which with parameters $m_e = 5 \cdot 02$ and $k_e = 0 \cdot 25$, presents many difficulties.

Other solutions are depicted in Figs. 5 and 6. These methods both make use of the formerly described property of the mirrored signal to resemble the original signal in the other line. A choice has to be
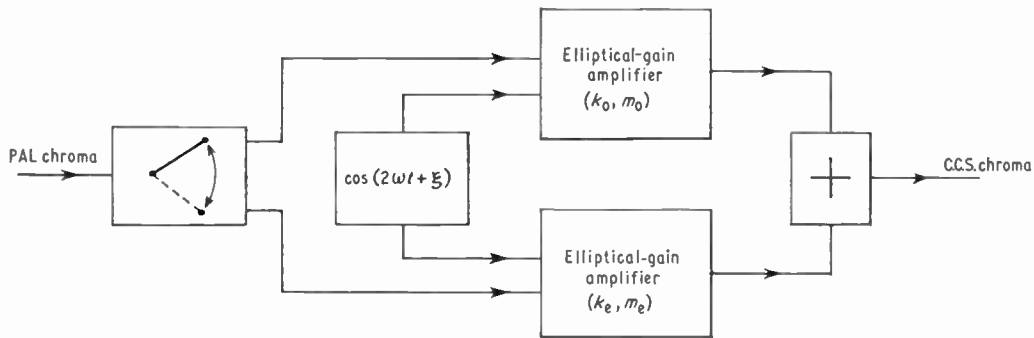
Fig. 4. PAL-c.c.s. conversion with two elliptical-gain amplifiers.
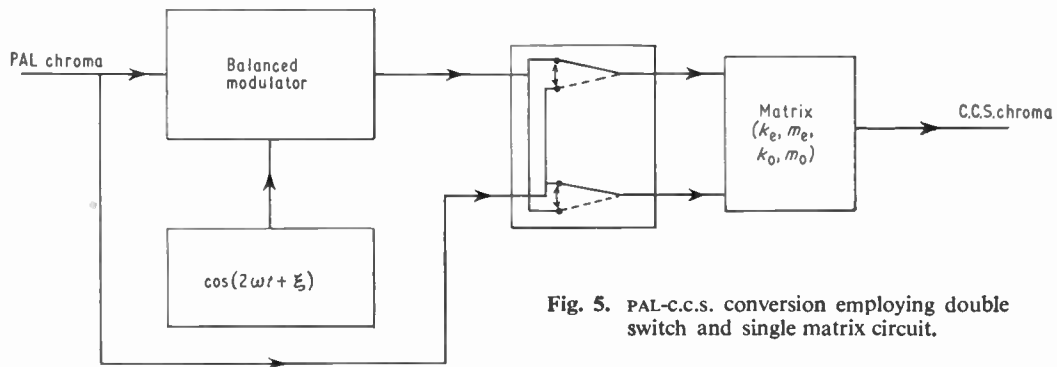
Fig. 5. PAL-c.c.s. conversion employing double switch and single matrix circuit.
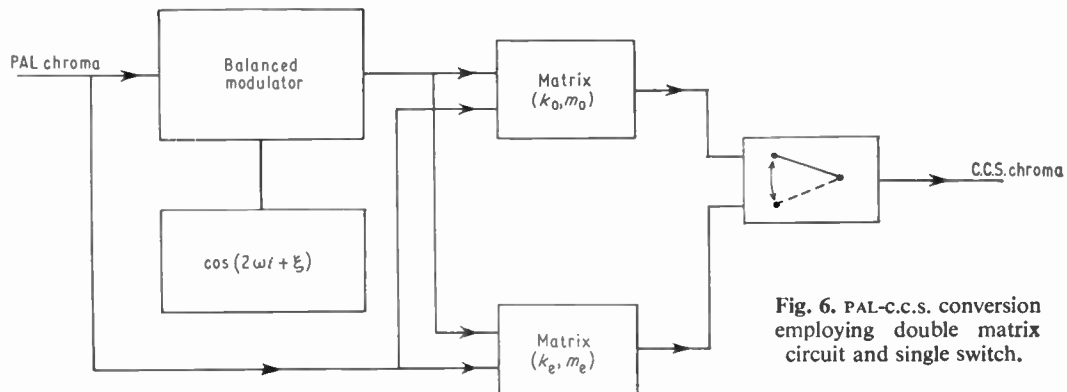
Fig. 6. PAL-c.c.s. conversion employing double matrix circuit and single switch.

made between:

(i) a complicated matrixing arrangement and a simple switch;

(ii) a simple matrix and a complicated switch.

In our view, the last method should be preferred, since a properly designed switch always works well and does not require alignment, while the alignment of a simple matrix network is simple and contains fewer components which can change with age.

5. Other Methods for Converting PAL into C.C.S.

The method just described for conversion of a PAL signal into a c.c.s. signal does not depend on the application of a 64-μs delay-line as has to be used in so-called 'PAL-delay' decoding techniques. This feature is particularly desirable if decoding according to the principles of 'simple PAL' is aimed at. In simple PAL, compensation of differential phase errors is effected by visual integration. As is well known, in this way only small errors can be compensated; however, circuit complexity is considerably reduced since the expensive delay-line and its associated circuitry is not necessary. In particular, this method seems attractive for small colour-television sets where one-gun display tubes may be preferred above small-screen shadow-mask tubes. Of course, the proposed conversion method does not exclude the possibility of applying a 64-μs delay-line for reducing differential phase errors. An obvious method for effecting delay-line decoding consists of simple averaging of the converted c.c.s. signals of two subsequent scanning lines. Figure 7 illustrates this principle.
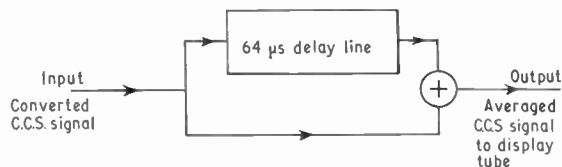
Fig. 7. Reducing differential phase errors by the addition of the signals of two successive lines.

If delay-line decoding is aimed at, other methods for PAL-c.c.s. conversion are also available. We shall discuss here briefly a method proposed by Bruch,[6] based on a PAL system in which the switching-axis is the I axis. With the aid of a 'comb filter' employing a 64-μs glass delay-line in conventional manner the modulated colour-difference signals are obtained. After line-sequential switching of the I signal and 90° phase shifting of the Q signal one obtains the signals

$$I \cos (\omega t + 33°) \qquad ......(15)$$

and

$$Q \cos (\omega t + 33°) \qquad ......(16)$$

These signals are then matrixed to yield

$$(R - Y) \cos (\omega t + 33°) \qquad ......(17)$$

and

$$(B - Y) \cos (\omega t + 33°) \qquad ......(18)$$

Subsequently the signal described by expression (18) undergoes a phase shift of 92° by means of an all-pass network. By proper matrixing of this signal and the signal described by expression (17) the c.c.s. chrominance signal is obtained. An illustration of this method is shown in Fig. 8.
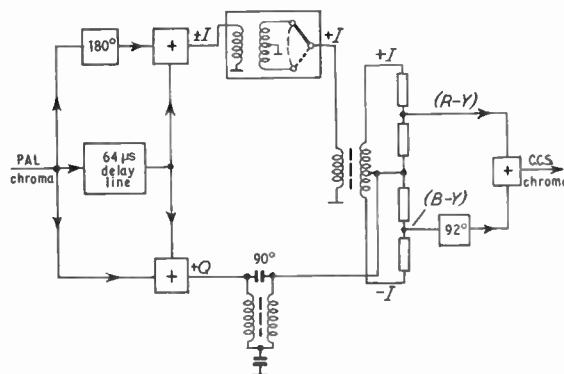
Fig. 8. PAL-c.c.s. conversion with delay-line decoder and phase-shifting networks.

An obvious disadvantage of such methods is the need for a 64-μs delay-line and phase shift networks, which have to provide constant attenuation and phase shift over a broad frequency spectrum of about 3–5 MHz.

6. Experiments with a Practical Converter Circuit

During the course of the investigations, a PAL-c.c.s. transcoder was built following the outlined ideas. Since no single-gun display-unit was available, the c.c.s. signal was decoded according to well-established methods known for N.T.S.C. decoding, so that an RGB monitor-display system could be employed. The signal source consisted of an N.T.S.C. encoder, followed by an N.T.S.C.-PAL transcoder. This procedure offered the possibility of varying the switching-axis of the PAL signal at will.

With the nominal value of $\theta = 92·4°$, no difference was observed between the displayed c.c.s. picture and the N.T.S.C. picture displayed at the same time, apart from a slightly enhanced visibility of the colour carrier in red and green. However, this must be attributed to the employment of the usual N.T.S.C. sub-carrier (odd multiple of half the line-scanning frequency), instead of the PAL sub-carrier frequency, which in fact should be employed. With the PAL sub-carrier, better dot interlace will be achieved for

all colours. With $\theta = 90°$, as in standard PAL, so-called 'venetian blinds' were barely visible. It may be recalled that this phenomenon occurs when phase and/or amplitude differences are present in odd and even lines. The aberrations for the primary and complementary colours, as observed on a vectorscope were found to be within 2% in amplitude and $3°$ in phase. No investigations were undertaken with a different luminance signal as outlined in Section 3.

7. Conclusions

The analysis and the experiments show that conversion of a PAL signal into a c.c.s. signal without full demodulation of the transmitted signal is well within the possibilities. Ideally the switching-axis of the PAL signal or the composition of the luminance signal should be modified slightly but the standard PAL signal is so near to the signal theoretically needed that employment of the former introduces errors which can hardly be observed. Of course the conversion circuitry is more complicated than in the case of an N.T.S.C. signal since a line-sequential switch is needed.

The method described avoids the employment of a delay-line and constant phase shift networks. The circuitry needed consists of simple and stable circuits; their alignment does not present serious difficulties. If wanted a delay-line can be added for compensation of differential phase errors.

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SUBJECT INDEX

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

INDEX TO ILLUSTRATIONS

INDEX OF PERSONS

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

Authors of papers which are given in abstract form are denoted by A.

Biographical references are denoted by B.

Contributors to discussion are indicated by D.

INDEX OF ABSTRACTS

This index classifies under subject headings the abstracts published throughout the volume in 'Radio Engineering Overseas', summaries of papers presented at the Symposium on 'Electronic Instrumentation' held in India in 1966, and abstracts of papers scheduled for presentation at the Conference on 'The Integration of Design and Production in the Electronics Industry' held at Nottingham in 1967.

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