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*To promote the advancement of radio, electronics and kindred subjects by the exchange of information in these branches of engineering*

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# The Radio and Electronic Engineer

**The Journal of the Institution of Electronic and Radio Engineers**

# Ploughshares from Swords

## INDUSTRIAL SCRAP AS AN EDUCATIONAL RESOURCE

FOR nearly two years the Shell Thornton Research Centre and other Cheshire industrial organizations have salvaged surplus and unwanted materials and equipment and made it available to more than 50 schools in the county, in what is believed to be a unique scheme of co-operation between industry and education. Offcuts of metal, paper, wood and plastics, obsolete instruments and electrical equipment, surplus chemical glassware and laboratory equipment, outmoded electric typewriters and calculators, plus a great variety of other laboratory and office equipment are now finding their way out of industry's store rooms and into the classrooms and workshops of Cheshire schools.

The scheme, started by Shell, is simple: Employees at the industrial locations put on one side those items of scrap or surplus material which they feel might have some educational use. The Local Authority arranges collection several times a week and then—using labour financed by the Government's 'job creation' scheme—sorts it into common lots in several converted classrooms at an old primary school in Ellesmere Port. Teachers from individual schools then visit the resource store to select and remove the items they feel will be the most useful to their own teaching and project work. No charge is made—everything is free for the taking!

Proving most popular with the teachers are the offcuts of metals of wide variety—and oddments of electrical and electronic equipment, switchgear and instruments. The more complex items of redundant scientific equipment from Thornton are cannibalized by the storekeeper and his assistants for their individual components which are in great demand for project work.

We are not aware of any widespread adoption of a scheme such as this within the electronics industry. But there must surely be many benefits in investigating it further. For one thing, it would be an excellent way of strengthening and adding to the closer and more 'organized' links already being built up between industry and the schools—such links have often been more easily forged by the larger firms whereas a scrap re-use scheme could be coped with by quite small companies. The again, schools would be helped to make their laboratory and other practical work much more related to the realities of industry today, while improvisation, whose value is perhaps underestimated in this age of the packaged black box, can be fostered in the schools.

What would be the direct benefits to industry? The Thornton Research Centre has found that accumulations of obsolete equipment or scrap materials within the plant are eliminated which meets with the approval of the Health and Safety Executive and incidentally also reduces the commitment for insurance on unwanted stock.

The involvement of the Local Education Authority as an intermediary between firms and the schools as in the Cheshire area is clearly desirable. Most L.E.A.s have Science Advisors to whom in these days of smaller allocations for education an almost free resource could well prove a godsend. The electronics industry ought not to wait to be asked to look into the possibilities of recycling its scrap material and obsolete equipment as a contribution to practical teaching of science and technology.

F.W.S.

# ANNOUNCEMENTS

## Military Microwaves MM 80

The Military Microwaves 80 Convention is a combined Technical Conference and Exhibition to be held at the Cunard Hotel, West Kensington, London, from 22nd to 24th October 1980. Although the event is solely addressed to advanced military systems and microwave technology it is nevertheless unclassified. The nature and scope of the material is complementary to that usually presented at the European Microwave Conference, so that the two events are not competitive. The papers will all be presented by invitation only, and have been selected by the Technical Programme Committee, under the chairmanship of Dr J. Clarke of the Royal Signals and Radar Establishment, Malvern.

The Convention enjoys the support of the Institution of Electrical Engineers, the Institution of Electronic and Radio Engineers, the Institute of Electrical and Electronics Engineers Microwave Theory and Techniques Society, the Associazione Elettrotecnica Ed Elettronica Italiana and of EUREL.

The Military Microwaves 80 Convention provides a unique forum for discussion between microwave component and sub-system engineers, and defence system engineers. The theme of the conference centres around this interface, and will address the ability of the microwave industry to meet the needs and requirements of weapon system programmes for the 80's. Therefore the bulk of the material in the technical papers is concerned with realization of engineering concepts in practical military equipments, and will discuss the final stages of equipment development as well as experience in the early years of operational deployment. The successes and difficulties of achieving specified performance in environment and qualification tests is important in this regard and is to be fully discussed.

The titles of the 21 sessions are as follows:

Active Electronic Counter Measures
mm-wave Circuit Technology
Communications
Solid State Transmitters
Antenna Arrays
Radar Test Equipment
Transmitter–Receiver Devices for Radar
Guided Weapons
Instrumentation & Radar
Air Defence Radars
Electronic Warfare Support Measures
Radar
mm-wave Targets, Clutter & Propagation
Satellite Systems & Technology
Polarization Control
mm-wave Radiometers
Microwave Tubes
Radomes & Antennas
Receiver Technology
Special Antennas
Low Noise Receivers

Each session will last about 100 minutes and contain four papers on average. The technical presentations and the Proceedings will be in English. No simultaneous translation into any other language will be provided.

Further information may be obtained from Microwave Exhibitions and Publishers Ltd., Temple House, 36 High Street, Sevenoaks, Kent TN13 1JG.

## New Training School for Microprocessor Appreciation Courses

One of the very necessary by-products of the 'microprocessor revolution' has been the burgeoning of courses, at all levels, intended for those who have to come to grips with this important electronic 'device'. Some of the courses are based on further education establishments, colleges, polytechnics or universities, some are run by specialist commercial training companies, and others, very appropriately, by electronic companies, whether semiconductor manufacturers or equipment manufacturers.

Among the last category are Cossor Electronics who have opened a new training school in Harlow town centre, near the Cossor factory, to cope with the demand for their 'Microprocessor Appreciation' and 'Introduction to Logic' courses.

The Company have been running these training courses for over a year, which, when first started, were seen as a natural extension to the customer training for their radar, telemetry and communications products. There is, however, such a demand from people in other industries who are entering electronics technology for the first time that the facilities have been extended and training staff now run the courses on a 'full-time' basis.

The 'Microprocessor Appreciation' course is designed as a general introduction to the subject and should interest managers, engineers and technicians who, either directly or indirectly, are now having to deal with microprocessor orientated products. It covers the broad parameters of how microprocessors work, what they can do, how they can be built into a product and how to select the right ones for a given application.

Similarly, the 'Introduction to Logic' course covers the broad parameters of small, medium and large-scale integrated logic devices and lays stress on their practical applications in electronic products.

Each course lasts for 5 days, but Cossor also offer abbreviated 2½-day versions aimed at managerial staff. Classes are kept small (up to eight people on any one course) and much of the time is spent in practical work, proving the theoretical matter on a range of teaching aids jointly produced with Feedback Instruments (based on Intel 8085).

Earlier this year the Institution's Editor was invited to experience an abridged version of the two abbreviated courses. The condensation of 10 days of lectures and practical work into 2 days was inevitably a severe trial of teacher and of pupils but the professional approach of these courses could be appreciated. Lectures and practice and the supporting handbook are obviously well conceived for the requirements of the engineers and others for whom the courses are intended.

## Corrections

In the paper 'Charge-coupled devices: Concepts, technology and limitations' in the May 1980 issue of *The Radio and Electronic Engineer*, one of the diagrams is in error. Page 203, Fig. 7(b): The 'profile' should be moved approximately 5 mm to the right so that the most shallow parts of the potential wells are aligned with the p-type implants in Fig. 7(a).

It is regretted that the photographs for Mr John Keen and Dr Daniel McCaughan on page 270 in the May issue were transposed. The Institution's apologies are extended to the authors and readers for any embarrassment which this error may have caused.

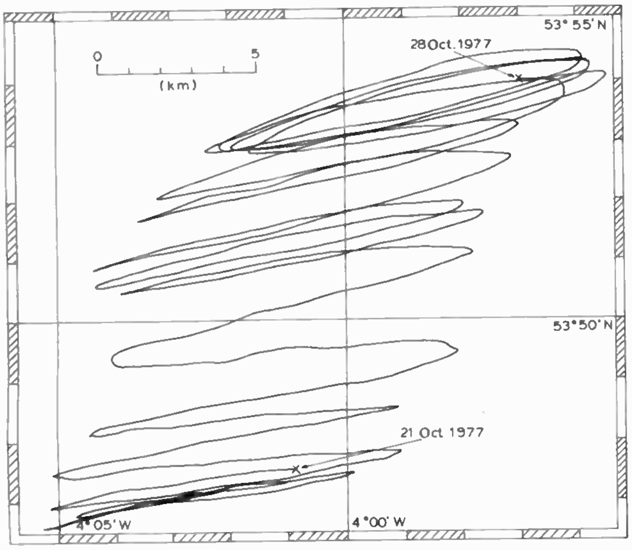


Fig. 7. Track of free-floating Tracer buoy in the Irish Sea. Test duration: 1 week. (After Booth[11].)

## 7 Conclusions

This novel version of the Tracer radio-navigation relay system permits the tracking of remote objects using the transmissions of the Decca Navigator system. The wide coverage and high reliability of these signals, and the low power consumption and small size of the converter, make the arrangement attractive for a wide range of monitoring and research applications. The accuracy and repeatibility of the results compare favourably with published Decca Navigator performance figures indicating that the Tracer equipment does not significantly degrade the quality of the position measurements. This observation allows the system's performance to be predicted in other situations.

## 8 Acknowledgments

The Tracer system was devised at the School of Electronic Engineering Science of the University College of North Wales. Engineering development was carried out by Industrial Development (Bangor) Ltd., in collaboration with Decca Survey Ltd., Leatherhead, Surrey.

The author acknowledges the agreement of the Directors of Decca Survey Ltd. to the publication of this paper, the valuable assistance of colleagues at Bangor and Leatherhead, and the cooperation of the Engineer-in-Chief, Trinity House, in mounting the trials on the Galloper Lightvessel. He especially notes the contribution of Dr. E. W. Roberts of ID(B) to the design and development of the system.

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*(Paper No. 1941/AMMS 102)*

# Contributors to this issue

**Geoffrey Bate** was born and educated in England. He obtained his B.Sc. (honours physics) and Ph.D. (physics) from the University of Sheffield. He served for four years as a Scientific Officer in the Royal Naval Scientific Service, working on infra-red photoconductors. He then went to the University of British Columbia, Vancouver, Canada as a Research Associate and later, Assistant Professor. He joined IBM in Poughkeepsie, New York, as a Staff Physicist to work on magnetic recording and later became Manager of Recording Physics, GPD Laboratory in Boulder, Colorado. In October 1978, Dr Bate joined Verbatim Corporation in Sunnyvale, California, as Vice President for Advanced Technology. He holds about fifteen patents and has published about thirty papers, mainly on the physics of magnetic recording. During 1980 Dr Bate was chosen as one of three Distinguished Lecturers of the Magnetics Society. He is also an affiliate faculty member of the physics department of Colorado State University.

**John Brush** is a lecturer in electrical engineering in the University of Dundee. A graduate of St. Andrews, he first worked as an assistant engineer at Smiths Aircraft Instruments, Cheltenham. For some time he was with General Electric, Syracuse, New York working on colour television developments and he returned to the UK in 1969 to take up his present appointment. He is concerned with the operation of the satellite receiving station at the University of Dundee which is funded mainly from grants obtained from the Natural Environment Research Council.

**Mike Underhill** (Graduate 1963) graduated in physics at Oxford in 1960 and joined the then Mullard Research Laboratories. He started work first on ultrasonics, then moved on to magnetic logic, control of focused electron beams for microcircuit manufacture, and now radio techniques and systems. In 1972 he obtained his Ph.D. as a collaborative student at the University of Surrey for the work on control of electron beams. Since 1968 Dr Underhill has been a visiting lecturer on the Systems Engineering M.Sc. course in the Department of Electronic and Electrical Engineering at the University of Surrey and he is currently lecturer on Sampled Data Systems.

**David Last** received the degree of B.Sc.(Eng.) in electrical engineering from the University of Bristol in 1961 and from 1961 to 1963 was a Graduate Apprentice with the British Broadcasting Corporation. He received the degree of Ph.D. in electronic and electrical engineering from the University of Sheffield in 1966 and was then appointed to a lectureship in the School of Electronic Engineering Science at the University College of North Wales, Bangor. He was made a Senior Lecturer in 1978. Dr Last has published a number of papers in the fields of non-linear circuit analysis, semiconductor and integrated circuit device optimization and radio navigation relay systems. He was joint recipient of the Brabazon Premium for 1977 for a paper on a receiver for a navigational relay system.

A biography and photograph of **Dr Nigel Mackintosh** appeared in the April issue of the Journal.

## Contributors to the May issue*

**David Lamb** is in the Microelectronics Group at Honeywell's Systems and Research Center, Minneapolis, and he is an Adjunct Professor in the Electrical Engineering Department of the University of Minnesota. He received his B.A. and M.A. in physics from the University of Oxford and his M.Sc and Ph.D. from London University. After working in industry from 1962–1968 for GEC and Associated Semiconductor Manufacturers he joined the Electronics Department of Southampton University. In 1978 he left Southampton to take up his present positions. Dr Lamb has been author or co-author of two books and more than seventy papers on semiconductor devices and technology.

**Norman Foss** is currently a Senior Principal Record Scientist at Honeywell's Systems and Research Center in Minneapolis, USA, where he is primarily responsible for the application and development of charge-transfer devices as related to the area of infra-red imaging and signal processing. Prior to joining Honeywell in 1975, he worked at CBS Laboratories, Stamford, Connecticut for 10 years in the area of electro-optics research including optical processing i.r. and u.v. sensor development and solid-state video displays. Mr Foss received a B.Sc. in 1964 and an M.Sc. in 1966 from the University of Minnesota and is a member of the Research Society of America and the American Physical Society. He is the author of over 20 publications in the field of electro-optics, infra-red imaging and signal processing.

* Information received too late for inclusion.

# Applicants for Election and Transfer

THE MEMBERSHIP COMMITTEE at a meeting on 13th May 1980 recommended to the Council the election and transfer of the following candidates. In accordance with Bye-law 23, the Council has directed that the names of the following candidates shall be published under the grade of membership to which election or transfer is proposed by the Council. Any communication from Corporate Members concerning the proposed elections must be addressed by letter to the Secretary within twenty-eight days after publication of these details.

**May Meeting** *(Membership Approval List No. 273)*

## GREAT BRITAIN AND IRELAND

### CORPORATE MEMBERS

**Transfer from Member to Fellow**

NETHER, Ronald Egmont. *Cobham, Surrey.*

**Transfer from Graduate to Member**

NOBLE, Ronald Harvey. *Maidstone Kent.*

**Direct Election to Member**

BURNHAM, Frank. *Camberley, Surrey.*
DAVEY, Robin Livingstone. *Reading, Berkshire.*
MIRZA, Zahid-Uddin. *Cambridge.*
RUSSELL, Richard Thomas. *Gravesend, Kent.*
TITTERING, Robert Geoffrey. *Leicester.*

### NON-CORPORATE MEMBERS

**Direct Election to Graduate**

WILSON, Graham John. *London.*

**Transfer from Student to Associate Member**

WOODFIN, Harold. *Manchester.*

**Direct Election to Associate**

CAUNTER, Denis Reginald. *Chelmsford, Essex.*

**Direct Election to Student**

CHRISTODOULIDES, George Demetrius. *Enfield, Middlesex.*
SLADE, Michael John. *Taunton, Somerset.*

## OVERSEAS

### CORPORATE MEMBERS

**Transfer from Graduate to Member**

RUBAROE, Gramini Tissa. *Colombo, Sri Lanka.*
UCHE, Onyemaechi Alexander. *Port Harcourt, Nigeria.*
ZAVAHIR, Mohamed Wazeer. *Galle, Sri Lanka.*

**Direct Election to Member**

CHAN, Ka Keung. *Hong Kong.*
LEE, Ping Woon Edward. *Hong Kong.*

### NON-CORPORATE MEMBERS

**Transfer from Student to Graduate**

CHIN, Tsun Kat. *Tai Hang, Hong Kong.*
LAU, Wai-Keung Silas. *Kowloon, Hong Kong.*
LI, Chi Chiu. *Hong Kong.*
LI, Wai Kwong. *Hong Kong.*
TSUI, Lap Fung. *Hong Kong.*

**Direct Election to Student**

CHAN, Hon Kong. *Hong Kong.*
CHAN, Kin Man. *Hong Kong.*
CHAN, Sin Fai. *Hong Kong.*
CHAN, Sai Kwong. *Hong Kong.*
CHAN, Tak Ming. *Hong Kong.*
CHAN, Wing Wai. *Hong Kong.*
CHEUNG, Sina Lam. *Hong Kong.*
CHIA, Song Leng. *Singapore.*
CHONG, Nyong Kok. *Hong Kong.*
CHUNG, Wai Ming. *Hong Kong.*
CHUNG, Yau Wing. *Hong Kong.*
FUNG, Shu Yan. *Hong Kong.*
HON, Hou Kwan. *Hong Kong.*
HUI, Ching Man. *Hong Kong.*
IP, Sing Ka. *Hong Kong.*
LAU, Hin-Ho. *Hong Kong.*
LING, Yu Ming. *Hong Kong.*
SHAM, Chun Ming. *Hong Kong.*
TAM, Wing Lap. *Hong Kong.*
TSANG, Kam Tim. *Hong Kong.*
TSANG, Ting Kwong. *Hong Kong.*
TSE, Kin Ming. *Hong Kong.*
WONG, Chun Hung. *Hong Kong.*
WONG, Yiu Sang. *Hong Kong.*
YUEN, Fu-Tat Valentine. *Hong Kong.*

Lists of applicants for election and transfer will in future be published in *The Electronics Engineer*. This will enable an appreciable reduction to be made in the present delay which inevitably occurs between the Membership Committee's meeting and the confirmation of membership. The names of those whose elections or transfers have been confirmed will be recorded later in the Journal.

# Standard Frequency Transmissions

*(Communication from the National Physical Laboratory)*

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

| APRIL 1980 | MSF 60 kHz | GBR 16 kHZ | Droitwich 200 kHz | APRIL 1980 | MSF 60 kHz | GBR 16 kHz | Droitwich 200 kHz |
|---|---|---|---|---|---|---|---|
| 1 | −1.2 | 18.9 | 20.9 | 16 | 0.0 | 19.2 | 19.5 |
| 2 | −0.9 | 18.9 | 20.9 | 17 | 0.3 | 19.4 | 19.4 |
| 3 | −0.8 | * | 20.7 | 18 | 0.3 | 19.1 | 19.3 |
| 4 | −0.8 | * | 20.6 | 19 | 0.4 | 19.2 | 19.1 |
| 5 | −0.7 | * | 20.5 | 20 | 0.5 | 19.5 | 19.0 |
| 6 | −0.6 | 19.0 | 20.4 | 21 | 0.4 | 19.5 | 19.1 |
| 7 | −0.6 | 18.8 | 20.3 | 22 | 0.6 | * | 19.1 |
| 8 | −0.4 | 18.9 | 20.1 | 23 | 0.8 | 19.5 | 19.0 |
| 9 | −0.4 | 19.3 | 20.0 | 24 | 0.8 | 19.8 | 18.9 |
| 10 | −0.4 | 19.0 | 19.9 | 25 | 1.0 | 20.0 | 18.8 |
| 11 | −0.4 | 19.0 | 19.8 | 26 | 1.0 | 19.7 | 18.7 |
| 12 | −0.2 | 19.2 | 19.7 | 27 | 1.3 | 20.2 | 18.6 |
| 13 | −0.1 | 19.4 | 19.7 | 28 | 1.4 | 20.5 | 18.5 |
| 14 | 0.0 | 19.2 | 19.7 | 29 | 1.4 | 20.4 | 18.4 |
| 15 | 0.0 | 19.1 | 19.6 | 30 | 1.4 | 19.9 | 18.4 |

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

# Members' Appointments

## CORPORATE MEMBERS

**Col J. D. Parker, M.B.E., B.Sc.** (Fellow, Member 1944) has been appointed to the Board of Avel-Lindberg as a Non-executive Director. This follows the acquisition of the former Danish-controlled company by a British holding company, Agron Investments, owned by the directors and members of the staff of Avel-Lindberg. Colonel Parker who is Secretary General of the Comité International Radio-Maritime has been a member of several IERE conference organizing committees.

**W. E. Willison** (Fellow 1973, Member 1960) has been appointed Technical Manager of Chubb Electronics, based at its Technical Development Centre near St Albans. He was Engineering Manager with Chubb Integrated Systems and he was earlier with GEC General Signal and with Elliott Automation. Mr Willison has served on conference organizing committees and as a member of the Computer Group Committee. At the 1963 Convention on Electronics and Productivity he presented a paper on 'Data logging in power generating stations', published in the Journal in June 1963.

**Major F. P. Barnett, B.Sc., REME** (Member 1976, Graduate 1973) who has been Officer Commanding 12th Air Defence Regiment RA Workshop REME at Dortmund for the past two years, has returned to the United Kingdom and taken up an appointment in the Procurement Executive at the Ministry of Defence.

**C. R. Carter** (Member 1968, Graduate 1964), formerly a Senior Scientific Officer with the SRC Appleton Laboratory, has formed a partnership Carter Associates, based in Langley, Buckinghamshire.

**L. Choy, B.Eng., P.Eng.** (Member 1970, Graduate 1964) has joined the British Columbia Telephone Company at Burnaby, British Columbia, as an Engineer. He previously held a similar appointment at Roxboro, P.Q., with Northern Telecomm.

**K. E. Elsheikh** (Member 1979) has taken up an appointment as Video Engineer with U.A.E. Television in Abu Dhabi following five years as Head of the Video Maintenance Section of the Sudan Television Service in Khartoum. He has spent the past two years studying for the Plymouth Polytechnic Diploma in electrical and electronic engineering.

**C. P. Hyslop** (Member 1968) has been appointed Product Manager for the Micro Switch Division of Honeywell Control Systems in Lanarkshire.

**R. N. Jones** (Member 1970) has left Scitec Corporation of Sydney NSW and is now Company Director with ACE Transformers Pty, Southport, Queensland.

**J. Lowrie, B.Sc.** (Member 1970) has been appointed to the post of Senior Transmitter Manager at the BBC's Kirk O'Shotts Transmitting Station; he has been Transmitter Manager at the Holme Moss Transmitting Station for the past five years.

**J. N. Price** (Member 1973) has taken up an appointment with GEC Telecommunications (Nigeria) as Project Manager. He was previously with the General Electric Company in Ibadan where he held the post of Field Contract Controller.

**R. E. C. B. Smith** (Member 1973, Associate 1966) who has been with Cable and Wireless since 1977, has been appointed Chief Engineer, Development and Production at Smale House, London. He was formerly Assistant Chief Engineer, Voice Systems, in the Engineering Department at the company's Head Office.

**P. D. Starling, C.G.I.A.** (Member 1973) has been promoted to Principal Scientific Officer and is moving from the RAF Guided Weapons Range at West Freugh to join the Admiralty Scientific Service as Project Manager at the British Underwater Test and Evaluation Centre, Kyle of Lochalsh.

**R. Taylor, P.Eng.** (Member 1973) who went to Bell Northern Research in 1977, following seven years with Standard Telephones and Cables, Foots Cray, has now joined Bell Canada at their headquarters for technology development in Ottawa as a Supervisor Engineer, Transmission Standards.

**P. W Walters, B.Sc., M.Sc.** (Member 1978, Graduate 1974) has been with GEC Computers for 2½ years as a Senior Engineer, and has now left to take up an appointment as a Senior Design Engineer with the Data Information Systems Division of Siemens in Munich.

## NON-CORPORATE MEMBERS

**Lam Lai-Yin, M.Sc.** (Graduate 1976) has been promoted to Signalling Engineer II with the Hong Kong Mass Transit Railway Corporation.

**S. F. Monk** (Graduate 1971) formerly Third Engineer with the C.E.G.B. Leicester District, has now been appointed Second Engineer, C.E.G.B. S.E. District, Midlands Region.

**R. K. Partap, B.Sc.** (Graduate 1980, Student 1977) has joined Marconi Radar Systems as an i.f. receiver designer (microwave). He was previously an electronic graphic designer (p.c.b.) with Rank-Toshiba, Plymouth.

**J. Wilson** (Graduate 1978), formerly a Technical Officer with the Post Office Telecommunications in Glasgow, has been promoted to Assistant Executive Engineer and is now with Network Control in London.

**B. T. Hill** (Associate Member 1978) has taken up an appointment as Principal Engineer in the System Design and Validation Group, Project Ptarmigan with the Plessey Company.

**A. W. Hollingsworth-Palfrey** (Associate Member 1974) is now Assistant Field Controller, Zone 6 Nigeria with Telephone Cables of Dagenham, Essex. He was formerly Project Manager in the Posts and Telecommunications Project Group.

**Lt Cdr C. F. Mitchell, RN (Ret.)** (Associate Member 1973, Associate 1970) is joining Watkins-Johnson of Windsor as Senior Applications Engineer. His final service posting was as Senior Engineer Officer, Special Communications Unit, HMS *Mercury*.

**T. Muritu** (Associate Member 1980) who was with Standard Telephones and Cables, New Southgate as a Systems Test Engineer II, has returned to Nairobi to take up an appointment as Technical and Marketing Manager with Zurobi, an electrical and electronic engineering company.

# Obituary

The Institution has learned with regret of the deaths of the following members.

**Louis Guy Eon** (Fellow 1944, Member 1943) died on 6th October 1979 while on holiday at Weybridge, Surrey. Mr Eon who was born and educated in Canada, worked for the Northern Electrical Company for a number of years before the war as a sound engineer. In 1940, he joined the Canadian Army as a Signals Officer, and for some years, while in London, he was attached to Anti-Aircraft Command. After the end of the war, he remained with the Army, holding appointments with the Defence Research Board, and from 1955 to 1958 was Assistant Defence Research Attache at the Canadian Embassy in Washington. In 1960, he returned to the Defence Research Board where he remained until his retirement because of ill health in 1972.

**Captain Lucien Hix, RN (Ret.)** (Fellow 1955) died in January aged 73 years. Lucien Hix obtained a degree in physics following study at King's College, University of London in 1928 and up to the war taught mathematics and physics at schools in Belfast and near London. He joined the Royal Navy in 1941, serving as a Radar Officer, and up to his retirement from the Navy in 1962 he had held appointments at the Admiralty Signals and Radar Establishment and in the Electrical Engineering Department of the Admiralty in Bath. He subsequently served until 1970 at the headquarters of the British Railways Board. Captain Hix took a leading part in the activities of two Institution local sections, serving on the Committee of the Southern Section, and he was first Chairman of the South Western Section. His Inaugural Address entitled 'The Drift of Electronics' was published in the December 1959 Journal and discussed how man's faculties could be imitated by electronic devices and also considered the importance of switching techniques and interconnection using solid state devices.

**Lt-Col C. F. Newton-Wade** (Fellow 1926) died recently in Brisbane, Australia, aged 85. Colonel Newton-Wade had a long and distinguished career in radio engineering dating back to when he joined the Siemens Company in 1913 and was concerned with the installation of radio equipment in ships during the first world war. In the inter-war years, he held appointments with telecommunications organizations in the Far East and in the second world war he served with the Australian Army, belonging to a unit preparing for the re-occupation of Borneo and Hong Kong; during this period he was for some time at the War Office in London. At the cessation of hostilities he was involved in the reconstruction of electricity supplies and communications in Borneo and subsequently, after demobilization, joined the Civilian Government as Postmaster General and Controller of Telecommunications which included duties of Chief Electrical Inspector.

When he considered that his work in this direction had reached a satisfactory conclusion, he resigned from government service and returned to Australia where he worked in the engineering department of the first television transmitting station in Queensland. His retirement from active work in 1966 allowed him to become involved in amateur radio activities with the Wireless Institute of Australia and also to travel widely to many parts of the world.

## EMERITUS PROFESSOR E. E. ZEPLER—AN APPRECIATION

**Emeritus Professor Eric Ernest Zepler** (Fellow 1944), who died on 13th May at the age of 82 after a long illness, made an outstanding and pioneering contribution to radio receiver development as well as to the teaching of electronics. The son of a country doctor in Westphalia, he received a broad, cultured, early education. Among other talents he was a competent chess player at the age of 8 and by 14 he was composing chess problems, an interest which was to continue unabated in later years.

After studying physics at the Universities of Berlin and Bonn he took a D.Phil. at the University of Würzburg. He then continued his research at the Physical Institute, Würzburg, until joining the staff of Telefunken, Berlin, in 1925. His advancement was rapid and he became head of the radio receiver laboratories in the firm which was, and still is, the leading telecommunications company in Germany. In 1935 he was forced to flee the country with his family, leaving all possessions behind, and came to England as a refugee. He obtained a post with Marconi's Wireless Telegraph Company and continued his radio work.

At that time the design of radio circuits was something of a 'black art' requiring, it was thought, a good deal of experience together with much trial and error. However, with his clear analytical mind he made a detailed and logical study, as a result of which the apparent complexities of radio receiver operation were unravelled and placed on a clear scientific basis. It thus became possible, for the first time, to predict accurately, and in advance of construction, the characteristics including the sensitivity and selectivity, of a radio communications receiver. The results of his work were described authoritatively and with enviable clarity in his first book 'The Technique of Radio Design', first published in 1943 and running to three printings. This book was a classic and remained the standard reference for nearly twenty years—a remarkable feat in a subject which was developing so very rapidly.

His name is associated with many famous radio receivers and transmitters, for example the R.1155 and T.1154 used by Bomber Command during the war. In fact equipment of his design was used by both the Royal Air Force and the Luftwaffe.

From 1941 to 1943 Eric Zepler was a lecturer in the (then) University College, Southampton before moving to the Cavendish Laboratory at the University of Cambridge. Three years later he returned to Southampton and in 1947 separated, with his group, from the Department of Physics to form an independent Department of Electronics. In 1949 a Chair of Electronics was created for him.

This Chair and the new Department were the first in Electronics in this country, and probably in the world. The Department he led, in his modest but very effective manner, initially included technician courses subsequently transferred to the College of Technology. An enviable reputation was soon established and the postgraduate Diploma in Electronics became renowned as the outstanding qualification for professional electronics engineers in the United Kingdom. His pioneering work in the teaching of Electronics, as well as in research, attracted staff of the highest calibre from which two Vice-Chancellors and a steady continuing stream of Professors have emerged.

During this period Eric Zepler and S. W. Punnett led a team from the Department producing two further very successful books on Electronic Devices and Networks which were translated into several other languages. More recently he co-authored a further specialized text with Professor K. G. Nichols on 'Transients in Electronic Engineering'.

Eric Zepler took a leading role in establishing electronics as a separate and 'respectable' discipline from traditional electrical engineering. To this end he was an enthusiastic member of this Institution, helping to formulate the Institution's educational policies, and became its President in 1959/60 after holding a number of senior positions including those of Vice-President and Chairman of the Education and Examinations Committee.

In 1973 the Institution recognized his contributions to electronic engineering by founding the Eric Zepler Premium which is awarded for outstanding papers published in the Journal on the education and training of electronic and radio engineers.

On his first retirement in 1973 he began a completely new career in the University's Institute of Sound and Vibration Research. Freed from the heavy administrative burden of leading an active, rapidly-expanding, department he was able to concentrate on problems of hearing. To his own surprise, but not to that of his colleagues, he made many fundamental contributions to our understanding of the way in which the ear responds to impulsive sounds.

Among his wide and varied interests, which included art, music, bridge and literature, chess remained his principal enthusiasm. He published a number of books and, at various times, played for the Essex and Hampshire county teams. He was granted the title of International Master of Chess Composition, an honour which pleased him immensely.

An honorary degree of Doctor of Science was conferred on him by the University of Southampton in 1977.

He is survived by his wife, son and daughter.

W.A.G.

# Letters to the Editor

## The Technical Author

*From:* **F. P. Thomson**, O.B.E., C.Eng., M.I.E.R.E.
**B. H. Henson**

The letters by Mr Powell* and Mr Stone† are more than timely reminders that the training and deployment of technical communicators—whether they be styled 'writer', 'author' 'documentation engineer', etc.—is in need of drastic and urgent reappraisal and change.

Some eight years ago the Institution nominated me for membership of the City and Guilds of London Institute's advisory committee charged with the training of technical authors; I have been invited to only one meeting. On this occasion what appeared to be a pre-arranged short list of its members was agreed as a 'moderating committee' and those of us excluded from it apparently were regarded as being of no further importance, for I was given no means of learning what the moderating committee has done since, or how it continued to function despite the death of at least one of its distinguished members. As the C & G courses for technical authors have been widely accepted by colleges of further education, it is obvious there must be change if the status of the profession is not to be impaired further.

For some years several of us have been convinced of the need for a polytechnic-level national and international centre for communications studies which, most suitably, should be sited in the industrial heartland of England. The centre should offer diploma, degree and postgraduate diploma courses in several discrete fields of communication, for it is increasingly difficult to be a specialist in every aspect. The centre should aim to be both a national and international institution of learning for, by attracting students from all parts of the world, it should enhance the use of the English language as the major media for scientific and technical communication. The centre should, ideally, be an international clearing house for information and help to relate peripheral research projects in various universities, e.g. research into readability, graphics and recognition efficiency, international standards, print economics, techniques of presentation, the selection and training of suitable personnel.

Mr Powell has emphasized urgency. Mr Stone has emphasized the essentiality of a far more-broadly-based training curriculum, which would include legal studies. The most rapid and efficient way to move in the right direction is to find a suitable polytechnic which, for a modest amount of money raised from industry and commerce—perhaps internationally—could start setting up the centre within say two years.

Some five hundred years ago a certain William Shakespeare, of the County of Warwick, was instrumental in helping to establish English as the international language of literature. The Lanchester Polytechnic in the heart of Coventry (less than 20 miles from Shakespeare's Stratford) is also sited in the industrial heartland of England and easily accessible both nationally and internationally. Moreover, an honours degree course in communication studies is on offer, thus proving this Polytechnic has the basic ingredients for easy expansion into the centre of communication studies I have sketched; Mr Stone's requirement should be satisfied by facilities available from the excellent Department of Legal Studies. Donations of between £5,000 and £10,000 from 100 to 150 of Britain's major companies could bring this project into reality in short time and ensure a steady supply of highly competent technical communicators.

39 Church Road, F. P. THOMSON
Watford WD1 3PY
*19th March 1980*

The CGLI advisory committee for the communication of technical information was established in 1972 to take responsibility for three schemes—technical graphics, technical communication techniques, and technical authorship—which had previously been the concern of separate committees. At its first meeting, in November 1972, the committee established two special sub-committees to review the industrial needs, the examination structures and the aims of the three schemes: these were additional to the small moderating committees composed of experienced teachers and industrialists which consider draft question papers annually.

The two sub-committees reported to the second meeting of the advisory committee in May 1974. By this time, the Technician Education Council had been formed and, as the three schemes fell broadly within the TEC area of responsibility, it was decided to await action by that body rather than to embark on major syllabus revisions. No further meetings of the advisory committee have been held, although all members were invited to suggest any minor amendments to the syllabuses in January 1976 when the scheme pamphlets were reprinted. As with other technician level schemes scheduled for withdrawal when TEC replacements became available, the moderating committees have been maintaining control over the annual examinations and standards.

Mr Thomson's references to City and Guilds are inaccurate in that he has attended both meetings of the committee and, in addition to the 1976 letter referred to above, he was given an explanation of the position as recently as November last.

Technical authorship would seem to be a casualty of the changeover from City and Guilds to TEC provision. The Institute has continued to offer its examinations, which attracted only 36 candidates in 1979, in order that a recognized qualification might still be available to meet the needs of the profession but clearly these cannot continue on the present basis for very much longer. Provision has been made in our programme for 1980/81 to review industry's requirements for a City and Guilds qualification at an appropriate level but it is not possible at this stage to indicate what recommendations will emerge.

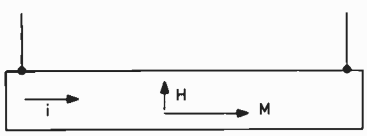
B. H. HENSON
Controller (Education and Training Services)

City and Guilds of London Institute,
46 Britannia Street,
London WC1X 9RG
*14th April 1980*

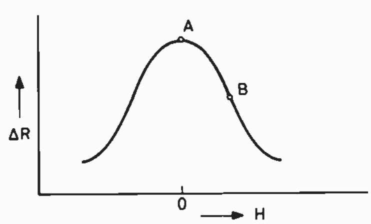
* *The Radio and Electronic Engineer*, **49**, p. 543, November 1979.
† *The Radio and Electronic Engineer*, **50**, p. 92, March 1980.

# A New Read Head for Magnetic Tape Recorders

As a result of magnetic feedback, magneto-resistance elements can acquire such a degree of linearity that it is possible, in principle, for them to be used for a number of magnetic recording applications. This is the result of recent research at the Philips Research Laboratories in Eindhoven. The principle on which magneto-resistance elements work is the change that takes place in the electrical resistance of a conductor under the influence of an external magnetic field. These elements have high sensitivity which is independent of the tape speed. Miniaturized versions can be manufactured by means of thin film technology, enabling read heads to be produced with a favourable signal-to-noise ratio and low cross-talk. This, in combination with the improved tape materials that are now available, makes it possible to increase the recording density. In itself, however, the magneto-resistance effect is far from linear and with the measures that have been taken hitherto it has been possible to compensate for this only to a certain extent. Magnetic feedback, however, brings about such an improvement in the linearity that it is even possible that magneto-resistance elements could be used in the read heads of high-fidelity audio equipment.




(a) Simple version of a magneto-resistance strip. The test current $i$ and the spontaneous magnetization $M$ which is due to anisotropy of shape have the same direction. $H$ is the external field being detected which brings about a rotation in the magnetization $M$ and therefore a change in the electrical resistance.




(b) The change in resistance $\Delta R$ due to the magneto-resistance effect as a function of the external field $H$. Unless further measures are taken, the operating point of a magneto-resistance strip is in the vicinity of point $A$ and there is substantial non-linearity. By applying a constant external field it is possible to move the operating point to $B$ and thus obtain reasonable linearity.

**Fig. 1**

The continuous improvement being obtained in the properties of recording tape is enabling the information density of magnetic recording to be increased. By using thin film techniques it is possible to manufacture the necessary miniaturized read and write heads. Miniaturized read heads using the magneto-resistance effect of a thin film of ferro-magnetic material (Fig 1(a)) are preferred to inductive read heads which, when produced by a thin film technique, are relatively insensitive. The magneto-resistance effect, however, is extremely non-linear as Fig. 1(b) shows. The change in resistance $\Delta R$ of a magneto-resistance element is plotted here as a function of the external field $H$. Unless special measures are taken, a magneto-resistance element operates round about point A where the non-linearity is at its maximum. If a constant magnetic field is now applied, the operating point can be moved to B and thus reasonable linearization can be obtained.

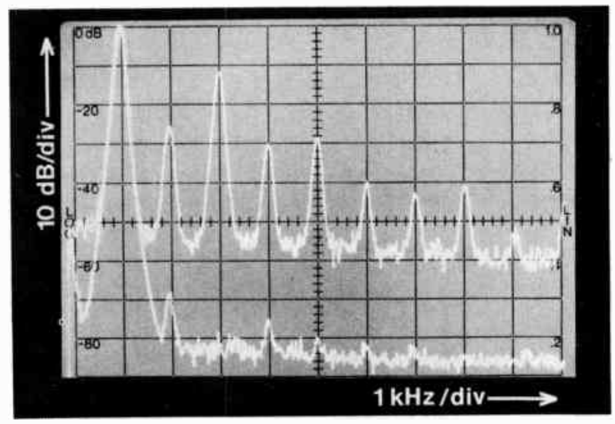



Fig. 2. Frequency response without (top) and with (bottom) magnetic feedback for excitation of a magneto-resistance element in an alternating magnetic field with a frequency of 1 kHz. Where there is no feedback, distortion produces a large number of harmonics, and a background is caused by Barkhausen noise. When magnetic feedback is used, the harmonics almost completely disappear and the noise is also considerably reduced.

For high-fidelity audio applications, however, even better linearity is needed and this can be obtained by combining shifting of the operating point with a magnetic feedback. To achieve this, part of the output current of the amplifier connected to the read head is fed back to a conductor which runs along the magneto-resistance strip giving rise to a field in the opposing direction to that of the field being read out.

The result of this feedback is shown in Fig. 2. The top curve gives the frequency spectrum of the output signal of a magneto-resistance element without feedback, for excitation in a magnetic alternating field with a frequency of 1 kHz. The large number of very strong harmonics are due to the non-linear behaviour. With magnetic feedback the bottom curve is obtained in which the harmonics have disappeared almost completely. A comparison of the two curves shows that in the case of feedback a noticeable reduction in the noise is also obtained. The noise concerned here is the 'Barkhausen noise' caused by abrupt movements of the magnetic domain walls when there is a change in the magnetization in a ferro-magnetic sample that contains more than one domain. The Barkhausen noise is reduced considerably by feedback because this noise, like the distortion, increases at more than a linear function of the magnitude of the alternating magnetic field being detected.

The constant magnetic field needed to shift the operating point can be obtained by feeding an extra direct current through the feedback line. The shifting can also be achieved by using the 'barber's pole' configuration of the magneto-resistance element which was developed some years ago in the Philips laboratories already mentioned.

# Programmable Transient Recorder

A new microprocessor-based two-channel transient recorder has been introduced by Datalab. The DL1080, stated to present a new concept in waveform storage, may be controlled and directed by a computer in an automatic system or used in a free-standing operator-controlled environment. In either mode, it is a powerful waveform acquisition and measurement instrument for one shot and low repetition events. It is designed for use in applications such as breakdown tests in electrical insulators, development of explosives and pressure vessels, impulse testing, high speed reaction chemistry and physics experiments, and power supply testing.

Amongst the many features offered by the DL1080 are an integral mini-cassette backup store and full compatability with GPIB interface protocol. DL1080 can simultaneously digitize and store waveforms, at sample rates of up to 20 MHz per channel, in each of two static r.a.m.s. of 4K × 8 bit capacity. Selection of recording parameters such as sweep time, trigger levels and input sensitivity can be achieved by means of a single control knob which provides an 'analogue feel' to parameter selection. This is considered to be superior to the 'keyboard only' approach in many modern digital instruments.

The Datalab DL1080 programmable transient recorder

Six recording modes are available on the DL1080. These include delayed sweep, dual-speed timebase and 'look back in time' pre-trigger modes together with the totally new zoom sweep recording facility. This new mode allows one channel to 'zoom in' at a high sample rate on an interesting region of the other channel.

The contents of both channels can be displayed simultaneously as analogue traces on an oscilloscope. Two bright spot cursors can be positioned on the display to provide constant brightness trace expansion and numeric display of amplitude and time information. Hard copy analogue readout of stored waveforms can be achieved by connecting a chart recorder to the DL1080. When an XY plotter is used, built in functions provide writing area calibration and maximized readout speed control, varied to the slow rate of the signal.

One of the disadvantages often encountered with digital instrument designs is that once power has been removed, memory data and front panel settings are forgotten. To overcome both these problems, in the DL1080 two user selectable non-volatile electrically alterable read-only memories are provided to store two complete sets of front panel settings and a mini-cassette recorder is built in to provide long-term waveform storage. Up to 14 complete memory records can be stored on the cassette together with file identification and recording conditions.

Applying power to the DL1080 will restore automatically all controls settings to the front panel from the primary non-volatile memory, and the settings in either memory can be recalled by push button or remote command at any time during operation. Waveform stored in the cassette can be replaced into the channel memories on command, from where they can be output in analogue or digital form in the normal way.

The flexibility and power of the DL1080 stems from its microprocessor based design. Among its many tasks the microprocessor provides the capability to sense remotely or select front panel conditions, to provide calibrated display and time expansion, to format and control data transfer to external equipment, and to control transfer of records to and from the integral cassette.

A central feature of the DL1080 is that all front panel functions, such as input sensitivity, sample rate, mode selection etc. are completely programmable by an external control unit. The built in GPIB interface means that the DL1080 can operate on the universally adopted IEEE 488 bus, making it compatible with a growing number of bus controllers, desk top computers and digital processors. In addition to GPIB, the DL1080 is capable of operating on serial data transmission systems requiring RS232/V24 compatibility. When fast transmission of stored data is essential, the unit also provide a direct memory access facility.

In operation the DL1080 can be used for waveform acquisition in a fully-automated system, as a free-standing instrument for fast analogue recording, or in an operator-controlled computer-related application. In a typical application, for example in an R & D or test laboratory, a number of waveform records may be captured, inspected and transferred to cassette. When the series is complete, stored data can be reviewed by an operator, before passing to a processor for manipulation. If necessary, data can be returned to the DL1080 for display or storage, again under automatic or operator control.

UDC 621.391.822: 681.844.086.7

*Indexing Terms:* Audio equipment, Sound reproduction, Phonograph records, Impulsive noise

# A system for reducing impulsive noise on gramophone reproduction equipment

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and

**I. BUCKNER,** B.Sc., Ph.D.†

**SUMMARY**

The paper describes a system for detecting and attenuating impulsive noise on gramophone records. Fast transitions and short duration are shown to be the essential characteristics of the impulsive noise waveform and a detector which uses these characteristics to distinguish between the noise and the recorded signal is described. Several methods of producing a noise-free approximation to the recorded signal are discussed and compared. A simple attenuation technique is shown to be as subjectively acceptable as more complex techniques. The designs of delay line and attenuator for optimum performance are considered and the subjective performance of the system is assessed.

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† *Formerly with Garrard Engineering Ltd., Advanced Development Unit; now with Hewlett Packard Ltd., South Queensferry Division.*

## 1 Introduction

A century after its invention the gramophone record is still the most widely used medium for domestic sound reproduction. This is because the gramophone record combines wide frequency and dynamic ranges and can be reproduced by systems of comparatively low cost. It is also eminently suitable for mass production. Against these advantages must be set the susceptibility of the vinyl material to accidental damage during handling and playing. Damage sustained in this way manifests itself as impulsive noise which is subjectively annoying because it can have a regular repetitive nature and is heard against an inherently low-noise background. A system for the detection and suppression of such impulsive noise can make a substantial subjective improvement to the performance of the gramophone medium. Such a system is the subject of this paper, and is shown schematically in Fig. 1. The system is designed to accept an unamplified signal from a magnetic gramophone pick-up and to deliver an equalized signal to the main amplifier after the noise suppression process. The signals from the left and right channels are amplified and fed into a delay line to a detector circuit. The detector, which recognizes the presence of impulsive noise, continuously monitors the signals from each channel. If the signals are noise-free they pass through the delay line and switchable attenuator and are equalized. When a noise impulse is detected the signal is attenuated for the duration of the impulse which is hence effectively deleted. The user can select the sensitivity of the detector to impulses and can switch the suppression circuitry in and out of circuit. In the Sections that follow we examine the features in the design of each part of the system which contribute to an effective and subjectively acceptable performance.

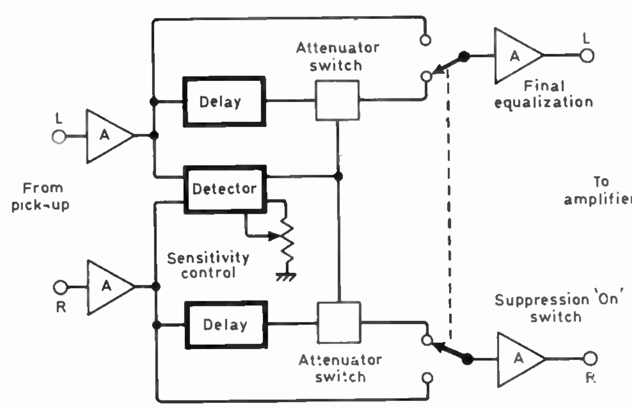


Fig. 1. The noise suppression system.

## 2 The Detector Circuit

The detector circuit must be able to distinguish between the noise waveform and the recorded signal. It therefore



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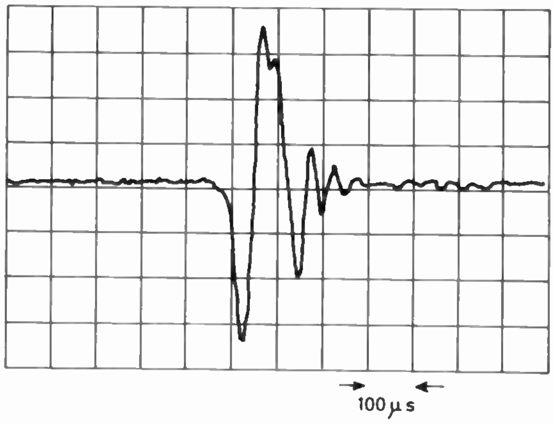
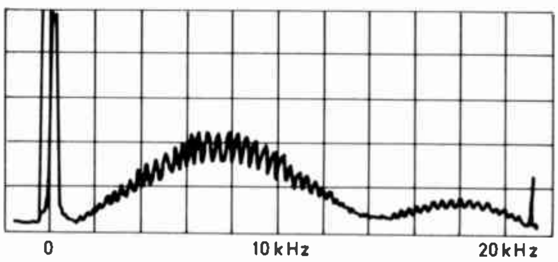
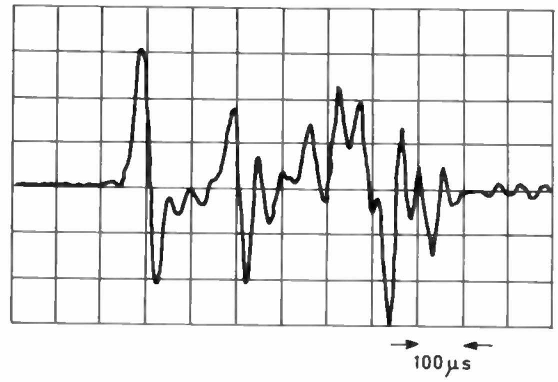


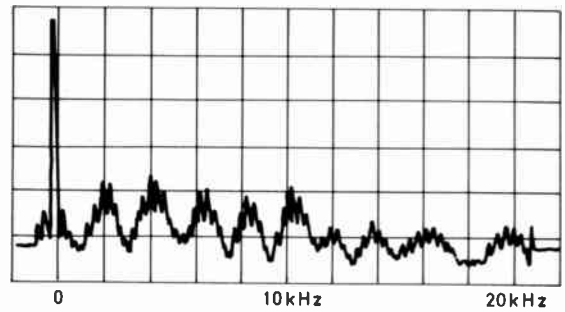

(a) Simple scratch, no equalization.




(b) Spectrum of (a).




(c) Complex scratch, no equalization.




(d) Spectrum of (c).

Fig. 2. Waveform and spectra of noise impulses (Shure M75ED2 cartridge).

is necessary to identify some feature of the noise which is not shared by the signal.

The first feature examined was the noise amplitudes in the sum $(L+R)$ and difference $(L-R)$ channels. The noise amplitudes were found to be somewhat larger in the difference channel and signal levels in the difference channel are less than those in the sum channel. Discrimination between the noise and the signal is therefore simpler in the difference channel.

The waveforms and spectra of a large number of noise impulses were examined and typical examples, shown in Fig. 2, show that there is a considerable difference both in waveform and spectrum between different noise impulses. Also the noise frequency components are well within the audio band and so the signal and noise cannot be separated by simple filtering. However, all the impulses share one common feature, namely fast transitions in the waveform between positive and negative peaks. The time taken by these transitions is about 50 µs. This is a typical figure, the actual value being dependent on the effective tip, mass and compliance of the pick-up cartridge. Observations of musical material showed that transitions of this sort were absent from the signals except at high levels.

The detector constructed (Fig. 3) compares the current signal amplitude in the difference channel $(L-R)$ with the amplitude 50 µs earlier by applying the current signal and a signal delayed by an all-pass active network to a pair of comparators. The sum signal $(L+R)$ is rectified and smoothed and the signal is used to control the threshold of the comparators via a potentiometer which the user can set. The comparators are arranged so that one triggers on a positive difference between current and delayed signal and the other triggers on a negative difference. The outputs of the comparators are combined in an OR gate which gives the detector output.

Under normal signal conditions the rate of change of the signal over a 50 µs period is insufficient to trigger the comparator. However, when impulsive noise with fast transitions of the type shown in Fig. 2 is present the comparator threshold is exceeded and the detector triggers. If the fast transitions occur in the signal these are accompanied by high levels in the sum signal which raises the comparator threshold and prevents false triggering. This differential input to the comparator when a typical noise impulse is present is shown in Fig. 4. The detector was found to discriminate satisfactorily between noise and a wide variety of musical material.

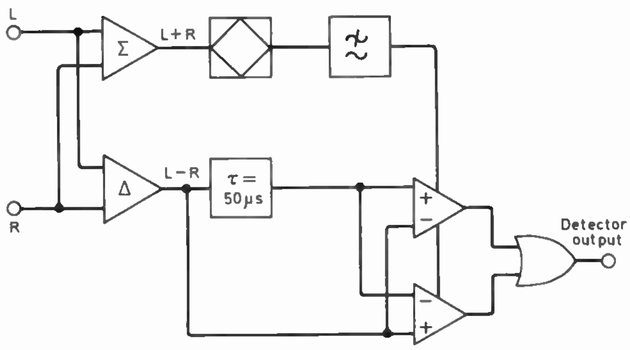



Fig. 3. The detector circuit.

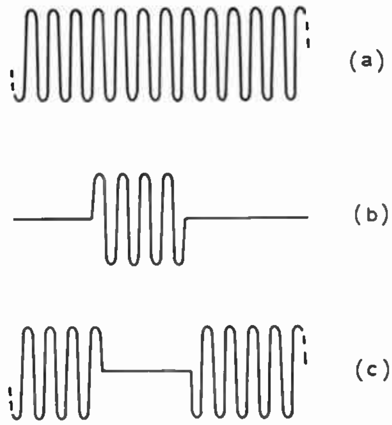



Fig. 5. (a) A continuous sinusoid.
(b) The error signal.
(c) The sinusoid after deletion.

## 3 Noise Deletion Techniques

### 3.1 Deletion Strategies

When impulsive noise has triggered the detector circuit the affected part of the signal must be processed to remove the noise. This can be done by a number of methods which include the following:

(a) track and hold the signal waveform constant for the duration of the noise;
(b) linearly interpolate between signal values before and after the noise;
(c) 'real-time editing' of the signal waveform;
(d) linear or non-linear low-pass filtering;
(e) simultaneous attenuation of the signal and noise.

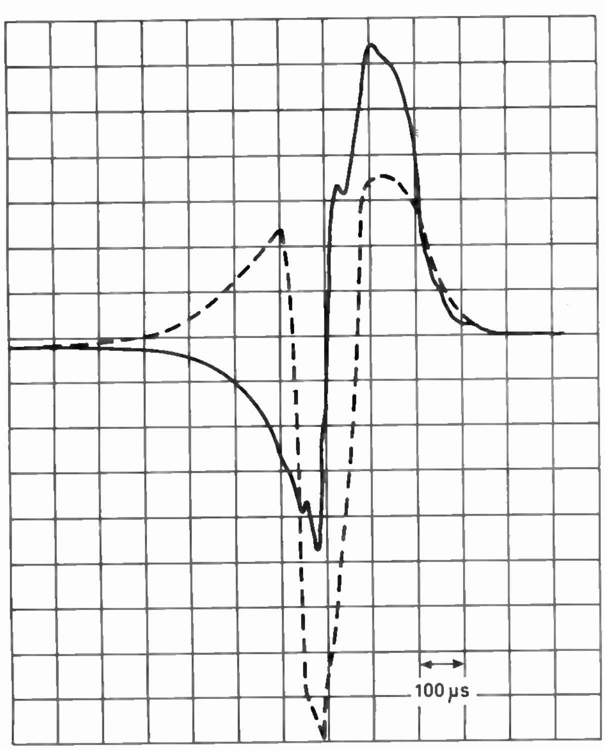



Fig. 4. Noise impulse and inputs to comparator.

Method (a) has been described by Craven and Gerzon[1] in connection with 78 rev/min records. The shortcomings of (a) and (b) have been discussed by Sacks and Bullingham[2] who favour method (c). This method consists of switching between the output and input of the delay line when noise has been detected and has therefore entered the delay. The switch is reset to the output of the delay line some time later when noise has passed through it. This method thus eliminates noise from the output but repeats a section of the output waveform when the switch is reset. The methods compared by the authors were (c) real-time editing, (d) low-pass filtering and (e) attenuation. Of these methods, (e) was found to be as subjectively acceptable as method (c) and more acceptable than method (d). Also method (c) requires that the output of the system is connected to the input of the delay line for at least 30 ms and Sacks and Bullingham themselves used 70 ms as a suitable time. The effect of this is that the system cannot remove any impulses occurring in this time which is a considerable disadvantage. Method (e) was therefore chosen for the system described.

The reasons for the good subjective performance of a technique which removes the original signal with the noise are worthy of attention. The effect of attenuating the wanted signal together with the noise impulse can be elucidated by considering the process not as attenuation but as the addition of a signal equal in amplitude but of opposite polarity for the period for which the original signal is deleted. The effect of this process can then be analysed in the frequency domain.

Figure 5(a) shows a continuous sinusoid while Fig. 5(b) shows a signal of equal and opposite polarity for the attenuation period. When added together they give Fig. 5(c) which is the original sinusoid with a section deleted. The spectra of these signals[3] are given in Fig. 6(a), (b) and (c). As can be seen from Fig. 6(c) the attenuator not only removes the existing signal but also produces other

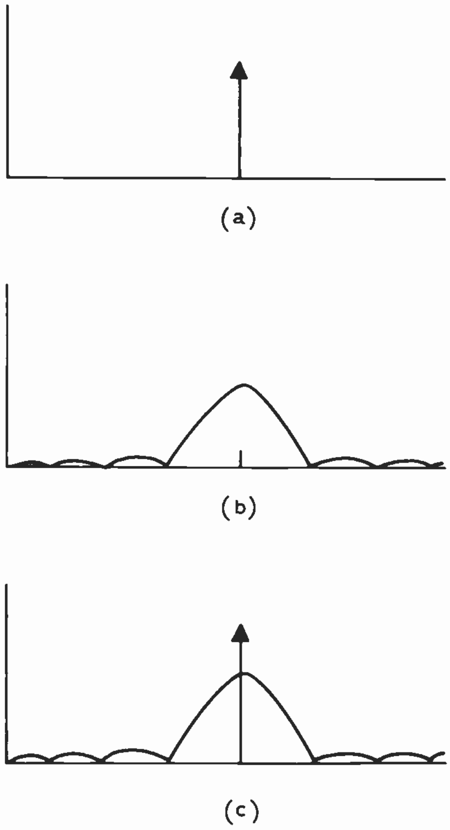



Fig. 6. (a) Spectrum of sinusoid (positive frequencies only).
(b) Spectrum of the error signal.
(c) Spectrum of sinusoid after deletion.

spectral components, which will clearly be noticeable. The number and magnitude of the spectral components produced will depend on the shape of the attenuation characteristic which is equivalent to the shape of the envelope of the error signal. The envelope and the spectral components generated form a Fourier transform pair, an effect which is the inverse of the windowing effect experienced in digital filter design. The optimum form of attenuation characteristic will be a inverse window which will minimize the spread of error components in the frequency domain. A suitable window would be the raised-cosine response:

$$W_H(t) = \begin{cases} 0{\cdot}5 + \cos 2\pi t/T, & -T/2 < t < T/2 \\ 0{\cdot}0, & \text{elsewhere,} \end{cases}$$

which may be regarded as an ideal for practical circuit.

The subjective effect of spectral components generated in this way will also depend on the presence of masking components in the original signal. For example, the effect of a short attenuation on pink noise signal is inaudible. Typical music signals usually contain sufficiently wide range of spectral components to render the attenuation inaudible. Other noise deletion techniques seek to preserve some part of the original waveform while attenuating the noise impulse. If these are analysed by the technique just described we find that the effect of increasing the size of the original signal is to reduce the size of the error signal, e.g. Fig. 7(a), (b) and (c).

If a significantly improved performance is to be achieved the error signal must be reduced substantially. However, a reduction of, for example, 10 dB in the error signal would require the preservation of 68% of the original signal waveform. Noting from the previous Section that most of the noise energy and signal energy occupy the same band and both have variable spectral characteristics, it is clear that filtering to reduce the noise to an acceptable level and to preserve the signal to an extent where there is a useful improvement over attenuation is extremely difficult. Hence the simplicity of the attenuation strategy is preferred.

### 3.2 The Attenuator Switch

From the foregoing it is clear that the attenuation characteristic has a marked effect on the performance of the circuit since this determines the envelope of the error signal and hence the new spectral components generated. Also the switching waveform itself can cause noise if it is coupled in to the signal channels through the attenuator switch. For example, when a junction field effect transistor is used as a switch attenuator, the switching waveform is coupled to the signal path by the gate-channel capacitance and this produces audible clicks when signal levels are low. While it is possible to use capacitor balance techniques[4] to reduce spike break-through problems in m.o.s. switches, it is difficult to set this balance accurately in volume production and to

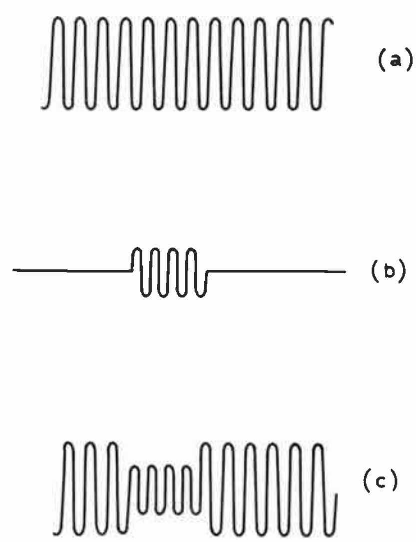



Fig. 7. (a) A continuous sinusoid.
(b) The error signal.
(c) The sinusoid after approximation.

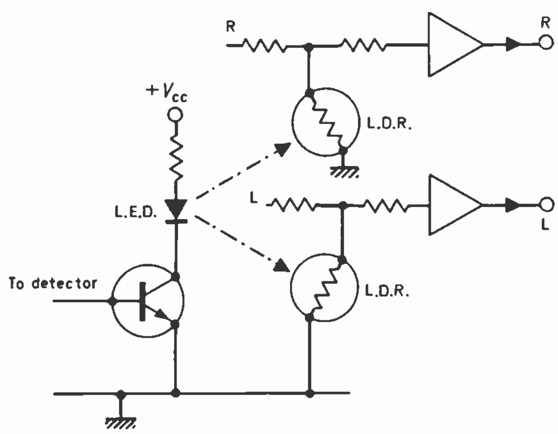



Fig. 8. The attenuator switch.

maintain balance when the equipment is subject to temperature variations since the capacitance required is only a few picofarads. These factors suggest that the switching waveform should provide smooth transition between the unattenuated and attenuated states, and the coupling between the switch path should also be minimized.

In order to achieve the switching characteristic and attenuation required the attenuator switch was constructed from light dependent resistors (l.d.r.s) and light emitting diodes (l.e.d.s). The resistors used have a high impedance ($>10^6\ \Omega$) in the dark condition which falls when they are exposed to light to a value which is dependent on the light intensity, typically 500 Ω. The switch is constructed (Fig. 8) so that light from the l.e.d. falls on to the l.d.r. and reduces its resistance. Under normal conditions the l.e.d. is off and the l.d.r. presents a high shunt resistance to the signal. When an impulse is detected the l.e.d. flashes, reducing the resistance of the l.d.r.s and thus attenuating the signal. The optimum operation of the circuit was achieved when the l.e.d. was on for 3 ms and the circuit was arranged to give a peak attenuation of 40 dB. Using this technique the attenuation occurs gradually, generating few extraneous spectral components, and complete isolation is achieved between the signal and switching waveforms.

## 4 The Delay

The function of the delay line is to allow the detector and attenuator time to operate before the noise impulse appears in the signal waveform. The detector requires 100 μs for operation and the switch 3 ms so that a total delay greater than 3 ms is required. This delay lies in the normal signal path of the system and any noise or distortion introduced by the delay will appear in the final output. The three types of delay considered for the system were:

(a) a continuous analogue delay using a series of all pass networks;

(b) a digital delay line using codecs and shift registers;

(c) a charge coupled device (c.c.d.) shift register.

The continuous analogue delay line cannot achieve the delay time required without a large number of sections which increases the cost and impairs signal quality. The digital delay can achieve the required delay time and quality but the cost is prohibitive for consumer applications. The c.c.d. delay line is low cost and can achieve the required delay time but the noise performance of these devices was initially considered inadequate. However, the noise performance of the system was improved by arranging for the standard R.I.A.A. de-emphasis from the record to be carried out in two parts.

The first of these was the 3180 μs time-constant giving the bass boost which was applied in the pick-up preamplifier. The second part was the treble de-emphasis (75 μs time-constant) which was applied after the delay line and attenuator. By this means a signal with pre-emphasis is applied to the delay line and de-emphasized after the delay giving a reduction in any high frequency components of the noise introduced by the delay line. Using this technique on a 256-stage delay line clocked at 85 kHz (giving 3·3 ms delay), an overall dynamic range greater than 85 dB was achieved with a total harmonic distortion of 0·1% at nominal output. Since this distortion is predominantly second harmonic the delay is suitable for use in a high quality system.

## 5 Systems Performance

The system as described has been tested extensively under laboratory and domestic conditions. The effect of the delay line on the signal quality can be ascertained by switching from the delayed to the undelayed waveform by means of the 'suppressor on–off switch' shown in Fig. 1. While a higher noise level in the delayed signal can be detected under no signal conditions it is not noticeable at normal listening levels. The small amount of distortion produced by the delay line is difficult to detect and is not subjectively significant. When the detector sensitivity is correctly set the system detects more than 95% of noise impulses and the operation of the attenuator is inaudible under most conditions. When a near continuous tone is being transmitted, such as an organ pedal note, the operation of the attenuator is discernible, but the subjective effect of the attenuation is more acceptable than the original noise impulse.

## 6 Conclusions

A system for deleting impulsive noise from the reproduction of gramophone records has been described and the major factors affecting the design of the system's components have been discussed. A commercial unit (Fig. 9) with an excellent performance has been developed and is in current production. In a domestic

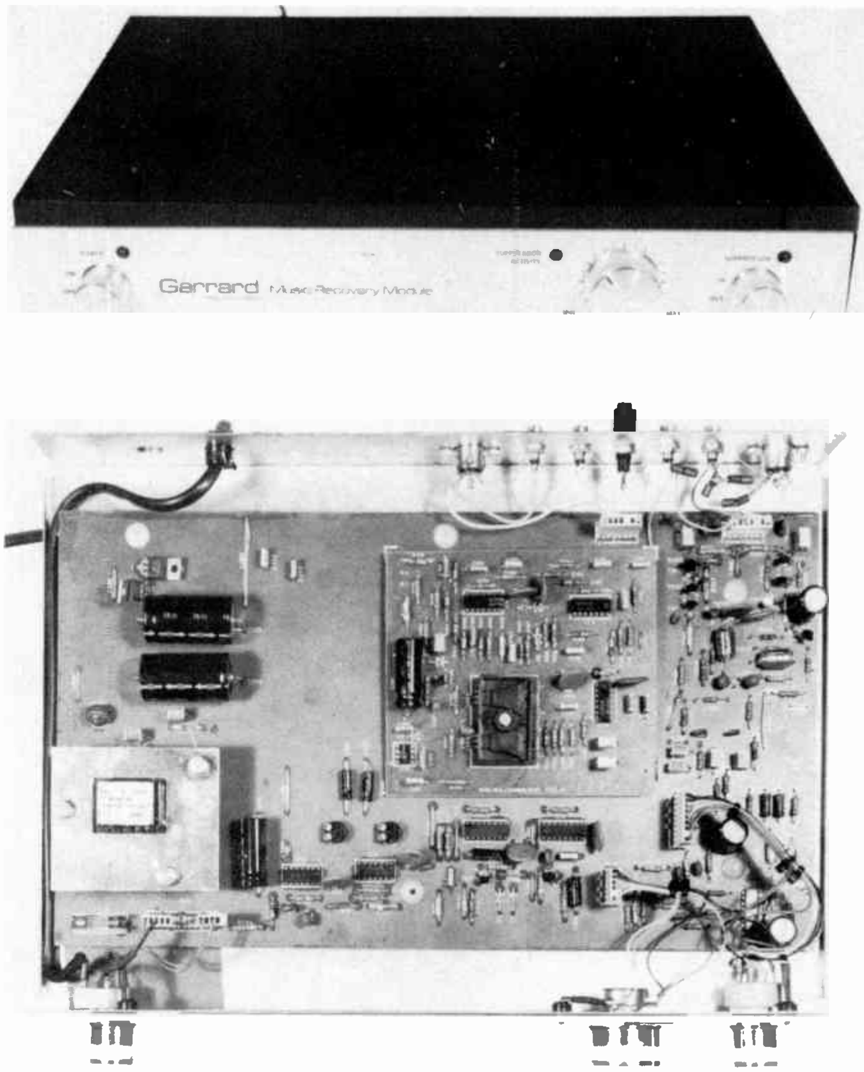



Fig. 9. (Top) The complete system.
(Bottom) The system showing detector and switching circuits mounted on the daughter board.

environment the unit gives a worthwhile improvement to the standard of sound reproduced from gramophone records.

## 7 Acknowledgments

The work upon which this paper is based was carried out during the development of the Garrard Music Recovery Module. The authors would like to thank Garrard Engineering Ltd. for permission to publish material in this paper.

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# Reliable high-speed data transmission from a mobile vehicle via a leaky feeder

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*Based on a paper presented at the IERE Conference on Land Mobile Radio held at Lancaster in September 1979*

**SUMMARY**

This paper establishes the feasibility of transmitting high integrity data, via a leaky coaxial feeder, at much higher rates than are normally considered possible over a mobile radio channel. Both signal variations and error rates are examined. The use of diversity techniques to improve performance are considered and possible system layouts are presented.

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## 1 Introduction

Over the past few years, there has been increasing interest in the application of digital techniques to the transmission of information over a mobile radio channel. Most of this interest has been aimed at low data rates of 1 to 10 kbit $s^{-1}$ (Refs. 1 and 2) and towards optimum use of the available bandwidths.[3,4]

Several authors have described studies of signal strength variations and their associated error rates[5] and have reported a Rayleigh distribution of signal strength in measurements at v.h.f. and u.h.f. in an urban environment.

Generally, even with these low data rates, the reliability is disappointingly low. As there are restrictions on the permitted transmitter power and bandwidths available, it is extremely difficult to achieve performance levels anywhere near that obtained on fixed data links. This paper addresses the problem of a much more demanding specification. It was required to transmit data at 100 kbit $s^{-1}$ from a vehicle to a fixed base station while the vehicle travelled around a 6 km track. This track was situated in hilly terrain with cuttings and tall trees present. The average error rate over this link was required to be better than 1 in 10,000 while the transmitted power was restricted to 250 mW.

With this demanding specification and unfavourable environment, free space propagation could not be used. Instead a leaky feeder system was proposed.[6] The arrangement consisted basically of a partially screened coaxial cable (leaky feeder), laid alongside the track, into which the signals transmitted from the mobile can 'leak' and then propagate with low attenuation to the base station. The use of this type of system is only possible since the mobile always remains within 15 m of the feeder.

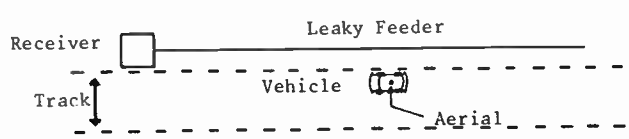


**Fig. 1.** Basic leaky feeder system.

When designing such a system, illustrated simply in Fig. 1, it is convenient to consider the overall propagation loss as having two separate components, cable attenuation and coupling loss. Coupling loss is the main variable; one definition of it is the difference in level between the signal applied to the mobile transmitter aerial and that in the feeder in its coaxial mode in one direction, at the nearest point to the transmitter. The coupling loss is affected by the height and supporting method of the feeder, by the transmitter polar diagram, by the distance of the vehicle from the feeder and also by the surrounding environment. Generally, it exhibits random or cyclic variations about a mean value with



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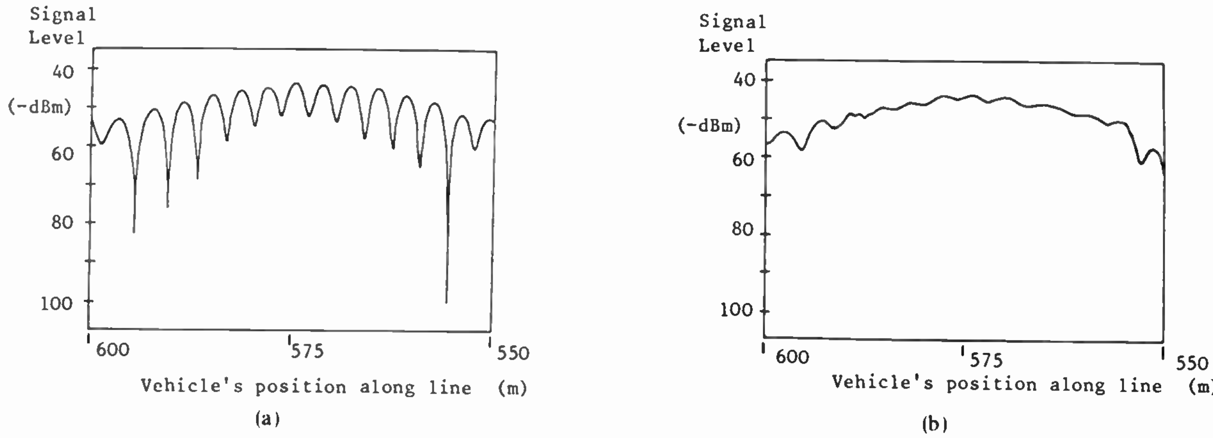



**Fig. 2.** Signal variations. (a) Far end unmatched. (b) Far end matched.

distance moved along the feeder, having localized signal minima or 'drop-outs' of 30 dB or more.

The characteristics of leaky feeder cables have been examined by Cree[6] and Martin.[7] From their results and taking into consideration the above mentioned points, a suitable cable was chosen, details of which are given in the Appendix.

The choice of frequency[7] involves a compromise between cable attenuation, which increases with frequency, and coupling loss, which tends to decrease with frequency depending upon design. The optimum for many feeders lies in the region of 30 MHz and for the present application a licence was obtained near this at 40 MHz. Hence, this frequency was used in the experiments described here.

To determine whether the specification could be met, measurements were made of the signal variations and the characteristics of data transmission using frequency shift keying for various system configurations. Also, the effect of the aerial position on the vehicle has been examined. Two forms of diversity, frequency and directional, have been investigated as a means of improving performance. Frequency diversity is a reasonable option because with leaky feeder systems conservation of spectrum is obtained through the low transmitter power possible with this technique. Finally, some consideration is given to the system layout and performance as a consequence of the presented results.

## 2 Signal Variations

To determine the signal strength variations, experiments were performed at 40 MHz using a layout similar to that shown in Fig. 1. Several parameters were varied in these tests, the most important being the feeder height, the mobile separation, the cable mounting method, and the mobile antenna. The results are summarized in Table 1, and a typical trace of signal strength is shown in Fig. 2.

As expected, the coupling loss increases as the feeder height is reduced and also as the separation of the mobile

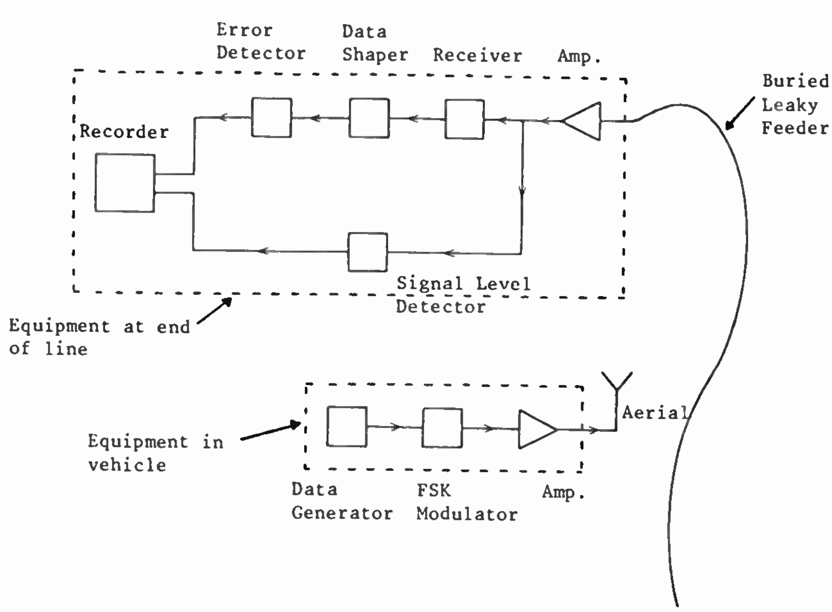



**Fig. 3.** Experimental arrangement.

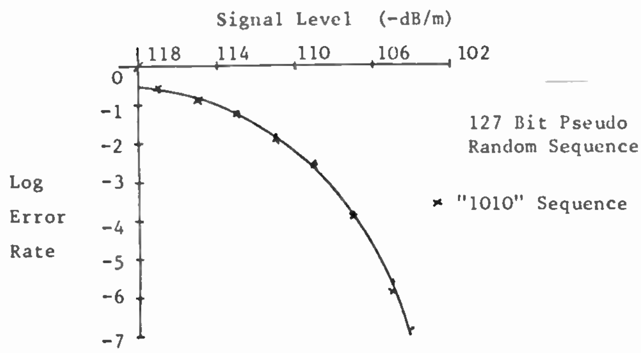



Fig. 4. Error rate performance.

from the line is increased. Conversely, however, the depth of the nulls and frequency of occurrence generally decrease in these two cases. It is also interesting to note that the coupling loss does not suffer greatly through burying the feeder a few centimetres under ground. Measurements under varying ground conditions have shown a variation in mean coupling loss, shown in Table 1, of ±4 dB for a buried feeder.

**Table 1**

Signal characteristics

| Separation (m) | Feeder height (m) | Null depth (dB) Max. | Null depth (dB) Typical | Coupling loss (dB) Min. | Coupling loss (dB) Typical |
|---|---|---|---|---|---|
| 4 | 1 | 50 | 20 | 56 | 60 |
| 10 | 1 | 30 | 15 | 61 | 65 |
| 4 | 0·5 | 40 | 15 | 61 | 63 |
| 15 | 0·5 | 40 | 15 | 71 | 73 |
| 4 | 0 | 50 | 10 | 65 | 70 |
| 15 | 0 | 40 | 10 | 79 | 84 |
| 3 | −0·01 | 50 | 10 | 68 | 74 |

Interference to the received signal caused by other passing vehicles is usually less than 10 dB, but if the transmitter is in a deep null then this null's position can move causing variation of 30 or 40 dB at that point. Interference can also result from metallic objects, such as signposts, situated alongside the track. These appear to be the major source of signal nulls, and in one case a null depth of 50 dB was observed.

Unless care is taken with the system layout, undesirable standing waves can occur, as shown in Fig. 2(a). The use of conductive suspension wire to suspend the feeder can result in exceptionally large standing waves. This also occurs if the end of the suspended leaky feeder is not correctly terminated in its single wire mode. To explain the reason for this, it is necessary to first consider the mechanism by which the transmitted signal is received. Radiation from the mobile sets up waves in both directions along the feeder in the single wire mode (that is a transmission line mode supported by the cable screen and the ground). These waves couple into the leaky coaxial cable, one of them producing the wanted coaxial mode propagating towards the receiver and the other a coaxial mode propagating in the opposite direction which, provided the far end is correctly terminated, is of no importance. However, if the second wave in the single wire mode is reflected at a discontinuity, it will set up a second, interfering, coaxial mode signal travelling towards the receiver. The end of the leaky feeder if unterminated in the single wire mode would provide such a discontinuity. The severity of the interference depends upon the nearness of the transmitter to the end of the line and the attenuation of the single wire mode, which decreases with increased feeder height.

The case of a suspended cable is made worse by the

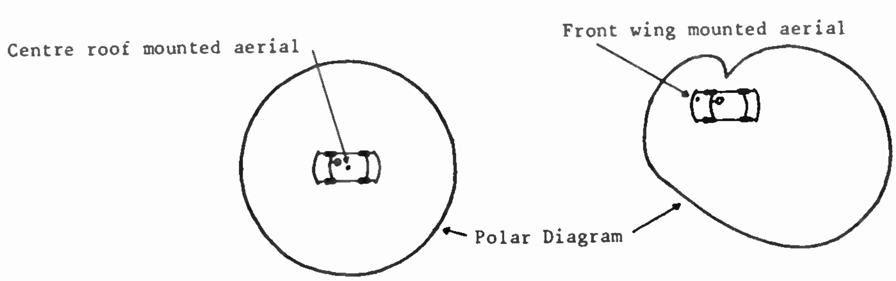



Fig. 5. Vertical aerial voltage polar diagrams (vertical polarization).

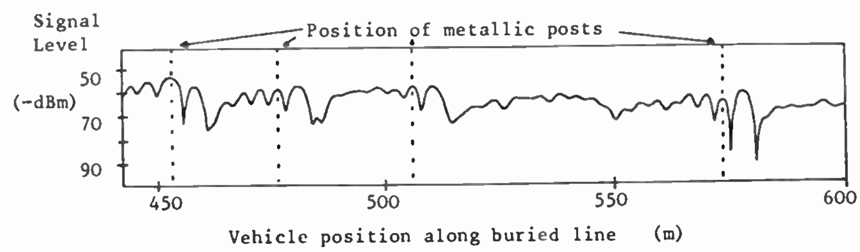



Fig. 6. Signal variations for mobile with centre roof-mounted aerial.

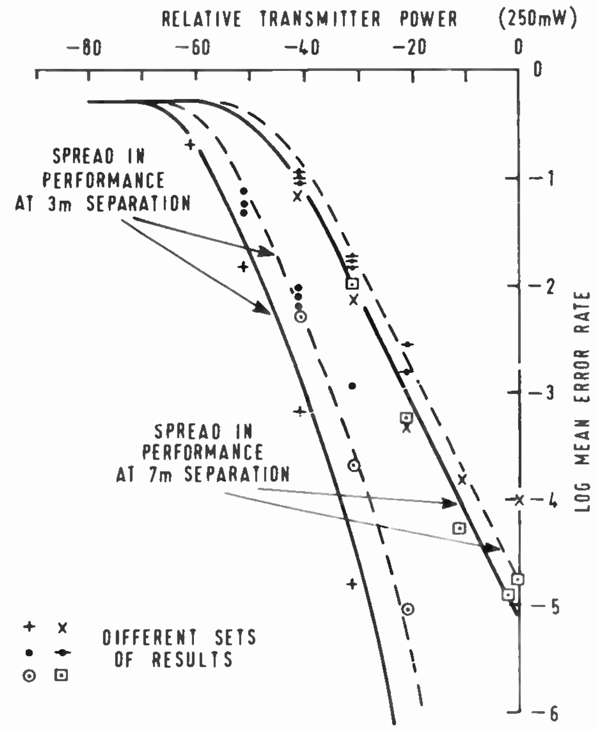



Fig. 7. Average error rate versus transmitter power—vertical aerial.

presence of a mode supported by the coaxial screen and the suspension wire. The interference in this case can be particularly severe because this bifilar mode is very low loss.

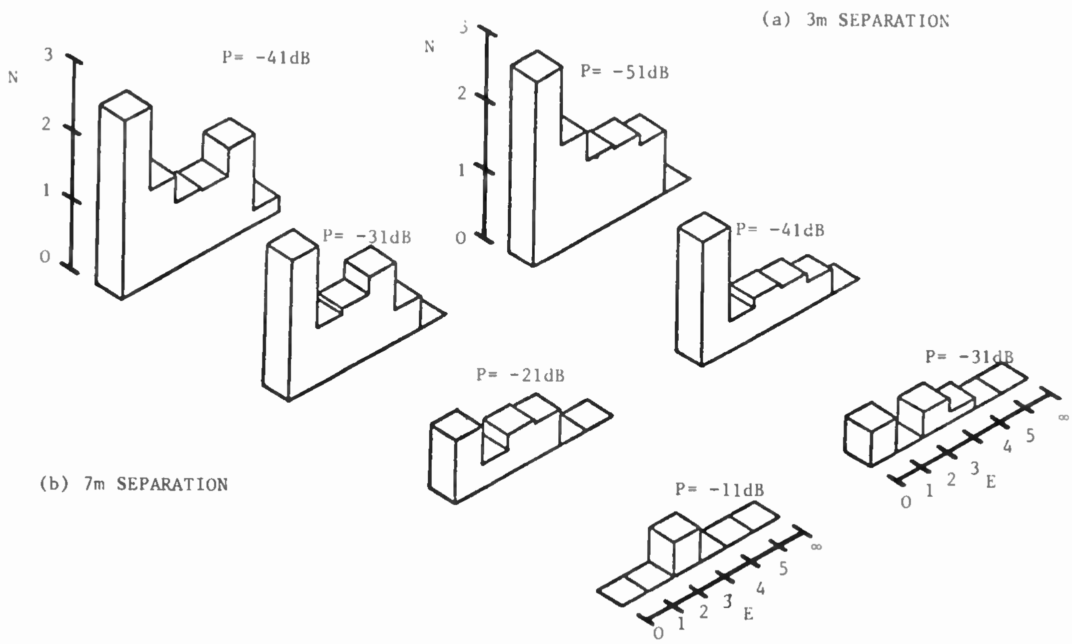
It is not easy to terminate these modes correctly, but one successful method is to extend the end of the line with non-leaky cable before bringing it gradually down to ground. This ensures that any end reflections are sufficiently attenuated before reaching the leaky section where coupling into the interfering coaxial mode can occur. Figure 2(b) demonstrates the effect of correctly terminating in this way. A second solution is to increase the loss of any non-coaxial modes, through, for instance, burying the feeder, or in particular to avoid using a conductive suspension wire.

## 3 Data Transmission

For performance, security and economic reasons the buried system appears the most promising. Hence a series of preliminary data transmission experiments have been performed using a buried line and the basic arrangement shown in Fig. 3.

Frequency shift keying was employed with a frequency separation of 100 kHz at a carrier frequency of about 40 MHz. The data rate was 100 kbit $s^{-1}$ and for experimental purposes the data sequence was a simple non-return-to-zero '1010' pattern, which allowed the use of a simple error detector at the receiver. The transmitter output power was 250 mW.

The effects of using the 1010 data pattern as opposed to a more conventional pseudo-random sequence were investigated in the laboratory under non-fading conditions. No discernible difference could be found in the error performance with these two data patterns, as shown in Fig. 4. It is thought likely that under frequency selective fading conditions the 1010 sequence will prove a sterner test of the system, but this will be




E = log (Error Block Size)
N = log (Normalized Number of Blocks)
P = Transmitter Power Relative to 250 mW

Fig. 8. Histograms of error block size—vertical aerial.

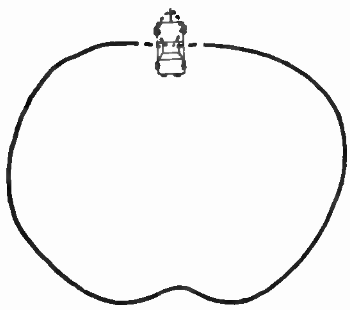


Fig. 9. Horizontal aerial voltage polar diagram (vertical polarization).

checked in the future.

The data transmission tests were performed with the mobile aerial in two positions on the vehicle: vertical, in the centre of the roof, where the majority of the measurements were made, and horizontal, as a tail, projecting from the back.

### 3.1 Vertical Aerial

The radiation pattern of the vertical aerial, both when mounted on the centre of the roof and also on the front wing, was measured with the results shown in Fig. 5. Using the centre roof-mounted aerial, a typical trace of received signal is as shown in Fig. 6. The major fluctuations in the signal are due to posts alongside the track.

Errors in the received data were recorded as the vehicle travelled along the track at varying speeds and distances from the line. Figure 7 displays these results. The lines are predictions from signal strength measurements. The points represent actual measurements. The day-to-day spread is shown, and is about 7 or 8 dB.

Certain observations can be made from these results. Firstly, no errors occurred with the vehicle 3 m from the line until the transmitter power was reduced to 2 mW. Secondly, at 7 m the gradient of the error rate versus transmitter power follows that predicted for Rayleigh fading—a 10 dB change in transmitter power gives a factor of 10 change in error rate; but at 3 m the required change in power is only 6 or 7 dB. Figure 7 also shows an effective loss of signal of 15 dB when the vehicle separation from the line is increased from 3 to 7 m. No change in error rate occurs with increased vehicle speeds, at least up to 36 km/h.

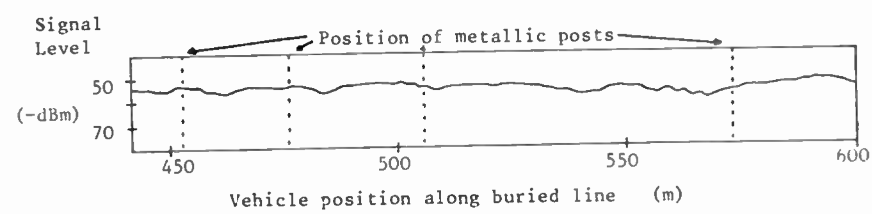



Fig. 10. Signal variations for mobile with horizontal aerial.

Histograms indicating the size of error blocks are shown in Fig. 8. The number of blocks has been normalized in each case to account for the journey time. It is interesting to note that around 50% of the error blocks are below 100 in length. The histogram for a pcwer of −11 dB shows the occurrence of bursts of errors in the range 100 to 1000. This effect is 'smeared out' at lower transmitter powers.

### 3.2 Horizontal Aerial

The radiation pattern measured for the horizontal aerial is shown in Fig. 9, which demonstrates its very directional nature. It has an advantage in the presence of vertical metallic structures since in this case they cause little interference, as demonstrated by Fig. 10 in comparison with Fig. 6.

The error rate when using a horizontal aerial depends upon the direction of travel. For a vehicle travelling (and therefore radiating) in the direction of the receiver, no

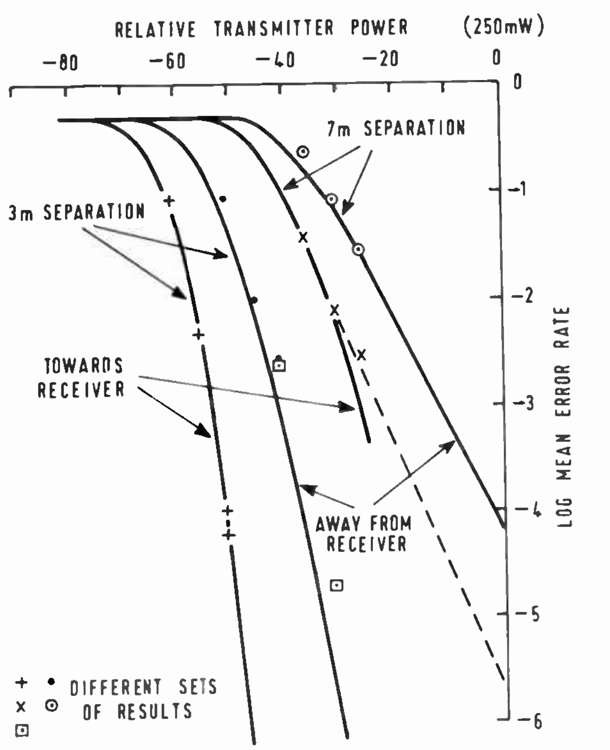



Fig. 11. Average error rate versus transmitter power—horizontal aerial.

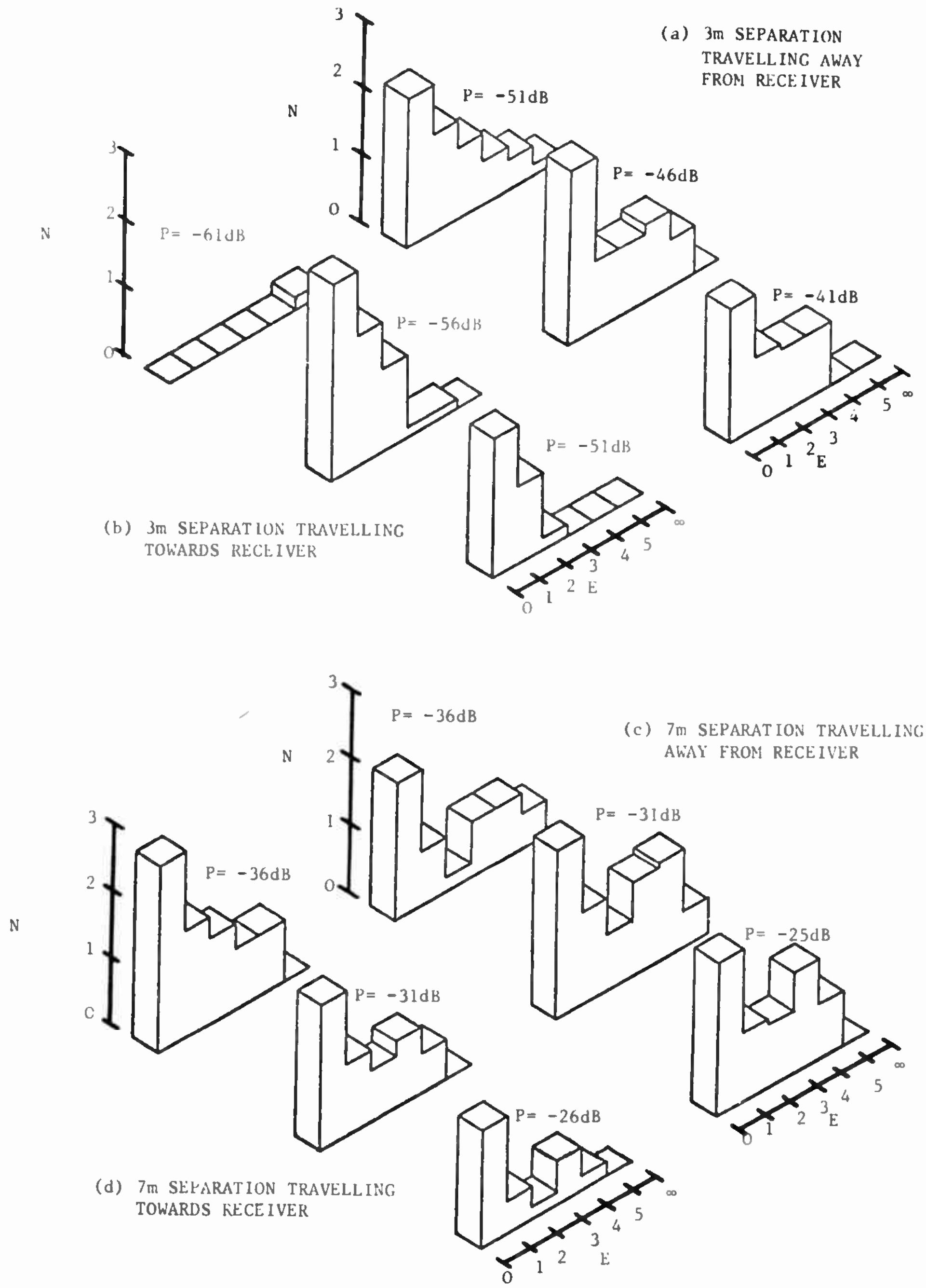



**Fig. 12.** Histograms of error block size—horizontal aerial.

errors at all were experienced with transmitter powers above 2 μW at 3 m separation from the line. For the opposite direction, and same separation, errors occur with transmitter powers below 200 μW. Figure 11 gives details of the error rate for both directions of travel. The solid lines show a prediction based on signal strength measurements. These can be unreliable when extrapolated because at low levels the measured signals become obscured by noise. The dotted line extrapolates according to a Rayleigh function.

These measurements can best be summarized by mentioning the transmitter powers required for an error rate of $10^{-4}$. At 7 m the power required is 250 mW, travelling away from the receiver, and 10 mW when moving towards the receiver. At 3 m the powers are only 50 μW and 2 μW respectively.

The distribution of error block sizes shown in Fig. 12 highlights the predominance of error blocks of less than 100 in length for the horizontal aerial. This is due to the signal variations being very small in this case.

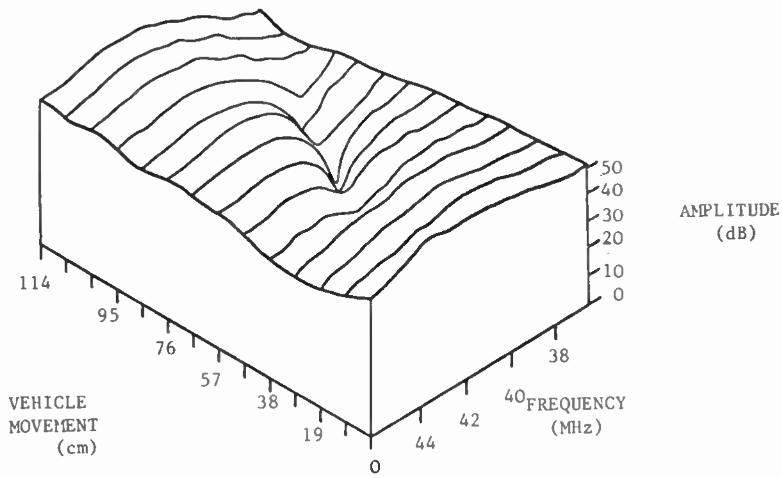



Fig. 13. Typical frequency and spatial characteristics for a buried line.

## 4 Diversity

The employment of diversity techniques should improve system performance. Both frequency diversity and directional diversity have been investigated, the latter being a form of spatial diversity using signals received at both ends of the feeder.

### 4.1 Frequency Diversity

The improvement possible using frequency diversity has been studied using a swept frequency transmitter and monitoring corresponding received signal variations at fixed vehicle locations. Typical signal traces obtained at consecutive vehicle positions around a signal null are shown in Fig. 13.

The conclusions from this work indicate that for a buried feeder the reflecting points are generally highly localized with wide frequency separations being desirable in a diversity system. A typical signal null caused by the presence of a nearby post is up to 30 dB deep and 2 MHz wide in the frequency domain, occupying about 0·5 m spatially. It is thought that a frequency separation of about 3 MHz should generally be adequate.

### 4.2 Directional Diversity

To determine the improvement possible through using both ends of the buried line, a receiver was placed at each end. A vertical centre roof-mounted aerial was used with 250 mW transmitter output power at 40 MHz. The signal variations at both ends were recorded simultaneously as the vehicle travelled along the line and in more detail around one particular interfering post, as shown in Fig. 14. This detailed recording shows an advantage of 21 dB and 3 dB can be obtained with this type of diversity in the case of two particular signal nulls at either end of the line.

. In more general terms for a complete run of the vehicle down the line, only 15% of the nulls (10 dB or greater in size) experienced at one end of the line coincide, within 20 cm of vehicle travel, with those at the other end. Of the remaining 85%, the typical spatial separation, or vehicle travel, to obtain corresponding signals nulls is 60 cm. The mean improvement in null depth available when reception can be switched to either end of the line is 8 dB with an additional mean improvement in signal strength of 8 dB. (These results are probably pessimistic because some of the nulls were 'buried' in receiver noise.)

## 5 System Layout

Upon consideration of the above results and other design criteria[7] two possible system layouts are proposed as shown in Fig. 15. They both employ repeaters and can be used with frequency diversity; however, the biline system also takes advantage of directional diversity. The

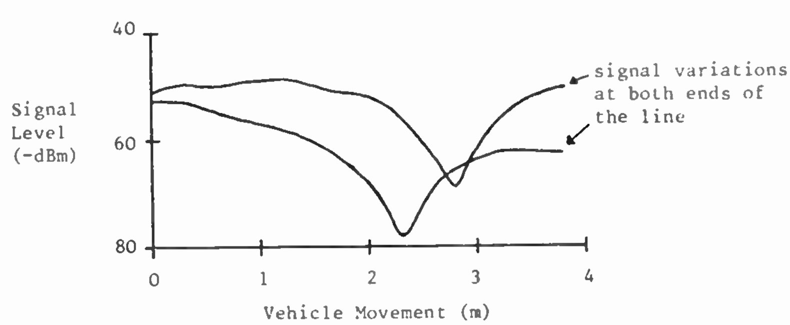



Fig. 14. Simultaneous signal variations at the two ends of the line around a signal null.

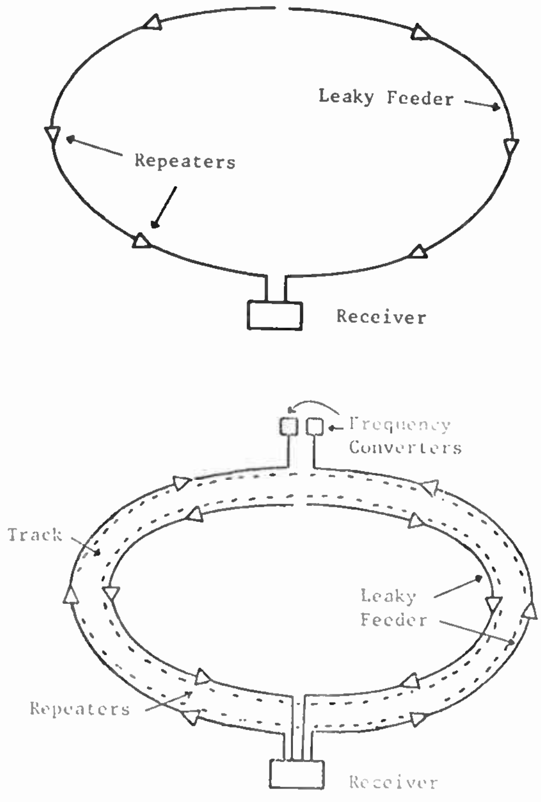



Fig. 15. Uniline and biline systems.

simpler uniline system can also gain from directional diversity if a repeater of the form shown in Fig. 16 is employed. Both systems are similar in that they are divided symmetrically into two. This is required in a circular system to ensure that the propagation delay does not suddenly change as the mobile moves around the track.

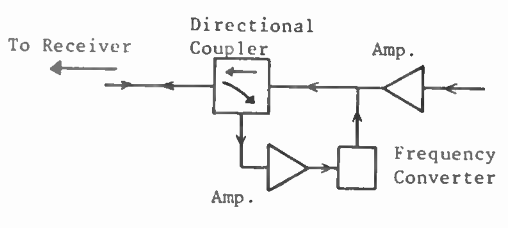



Fig. 16. Directional diversity repeater.

## 6 Conclusions

It has been shown that leaky feeders provide an excellent means of receiving high-speed data transmitted from a vehicle travelling along a fixed route. The required transmitter power is much less than is normally needed and the data rate is increased with greatly increased reliability. Further, employing the diversity techniques discussed here should significantly improve the system performance.

It is necessary to collect much more data concerning system performance, with and without diversity, before final statistical results can be obtained that are required in the successful design of a working system. This work is at present being undertaken.

It is worth finally mentioning that leaky feeder systems operate successfully with low transmitter powers and are less susceptible than conventional systems to interference from other co-channel transmitters outside their immediate vicinity. Thus, the same channel could be allocated to different users in much closer proximity than is possible with normal mobile radio systems. In view of this prospective saving in spectrum, it is desirable that consideration be given to the allocation of a dedicated channel for leaky feeder use somewhere in the region of 30 to 100 MHz.

## 7 Acknowledgments

The loan of the cable by British Rail Technical Centre, Derby, is gratefully acknowledged.

Mr. A. J. Motley was supported by a maintenance grant from the Science Research Council.

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## 9 Appendix: Details of leaky feeder

Manufacturer's designation: BICC type T3522
Construction: PE sheathed apertured copper tape coaxial cable
Characteristic impedance: 75Ω
D.c. resistance: 4·9 Ω km$^{-1}$ outer, 4·1 Ω km$^{-1}$ inner
Surface transfer impedance: 2·1 Ω at 30 MHz
Diameter of outer conductor: 10·1 mm
Diameter of sheath: 13·1 mm
Velocity ratio: 0·87
Coaxial mode attenuation: 23 dB km$^{-1}$ at 40 MHz

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*Indexing Terms:* Railways, Communications, Radio telephones, Automatic exchanges

# SPARCS—a stored program automatic radio connection system

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*Based on a paper presented at the IERE Conference on Land Mobile Radio held at Lancaster in September 1979*

SUMMARY

A large area coverage radio system with telephone access is described. Two operator-controlled systems have been installed on British Rail. This paper describes the development of a stored program automatic radio connection system, which allows direct dialling between individual mobile radio users and access to the British Rail automatic trunk telephone network. Emergency override access from the mobile radios is provided.

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## 1 Introduction

In his paper 'Radio Communication on British Rail', Kessell'[1] introduces the British Rail National Radio Plan. This is a national V.H.F. radio network, available to the different engineering and operating departments of British Rail, enabling staff in the field, either at the trackside or in vehicles, to communicate with one another or the railway telephone network. Originally, the system was envisaged as an operator-controlled network with about 30 separate control centres. Since the system is for multi-departmental use, the operator has no staff management or despatching function and only makes the necessary connections. This paper describes the development of SPARCS, an automatic radio connection system for the National Radio Plan.

## 2 The Operator-controlled Radio Connection System

It is worthwhile to recall the principles of operation of the two original operator-controlled radio connection systems now functioning at London (Euston), covering an area bounded by London, Bedford, Northampton and Aylesbury and at Glasgow covering the Ayr Coast across to Glasgow and Edinburgh and South to Carlisle. Large area coverage was required and thus a number of overlapping fixed stations were necessary to cover each area. It was felt impractical to expect the mobile users to change channel as they moved around the area in order to communicate with their 'local' fixed station. Therefore a single channel was required for the whole area. Simultaneous operation of all the fixed station transmitters was rejected, because interference in the overlap areas of two adjacent fixed stations precludes separate conversations in each fixed station area. Quasi-synchronous operation of the fixed station transmitters can reduce the overlap interference, but only one conversation may be transmitted. This would mean that only one conversation could be handled at a time in the whole control centre area. There are also technical problems of quasi-synchronous transmitter operation over frequency division multiplex land line equipment, which may be used for controlling the remote fixed stations. As an alternative, a time-multiplexed system was proposed for outward signalling and only one fixed station would be selected for the conversation. This uses similar principles to systems described by Jewell and Philips.[2]

The signalling system to the mobile was chosen as five-tone sequential selective call (Selcall). The five tones are of 350 ms total duration (70 ms each) followed by an audio tone of up to 1·5 seconds duration to alert the mobile operator. In order to send a Selcall out over the whole area, a Selcall sequence (5 tones plus alert tone) may be sent out on each fixed station in turn. Each fixed station transmitter (normally off) is turned on only for the duration of the Selcall sequence. In this way only one



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transmitter is energized at any one time and overlap interference between adjacent fixed stations is avoided. This procedure can be improved by energizing a number of non-overlapping fixed stations simultaneously. Again, no overlap interference is received by the mobiles. In a typical case it has been found possible to reduce a large area coverage system using eight fixed stations to three non-overlapping Selcall groups. These groups are energized in sequence and the complete area can be Selcalled in less than 5 seconds.

When the mobile replies, having received his Selcall and been alerted, his transmission may be received by more than one fixed station receiver. Using a telemetry system, which measures the fixed station receiver incoming signal strength or signal-to-noise ratio, it is possible to indicate the 'best received' fixed station at the control centre. This is called a voting system. The operator then selects the 'best received' fixed station for the conversation. The other fixed stations are then available for further traffic, with the exception of those fixed stations which overlap with the selected fixed stations. These may not be used, since they could cause overlap interference to the mobile in conversation. Accordingly, they are disabled and excluded from the voting system. These disabled fixed stations are known as 'guard stations'. The guard stations for each fixed station have to be determined from practical survey results.

Busy indication is given to mobiles in the same fixed station area as the mobile in conversation by the fixed station radiating 'pip tone' when the mobile transmits. There are problems when a mobile in a guard station area requests service from the operator. If the guard station receiver has been excluded from the voting system, no indication will be received by the control operator. Consequently there will be no action from the operator, which will be very frustrating for the mobile calling in.

## 3 An Automatic Radio Connection System

The requirement for automatic connection of mobile radios and the railway trunk telephone network was discussed in 1976 whilst the Euston and Glasgow systems were under construction. Since 24-hour coverage was needed and nearly 30 systems would eventually be required, at least 90 new staff might be necessary. These figures were modified slightly, since in some locations the radio operators would perform other additional tasks. However, since this radio network was to be financially justified by staff savings or increased productivity, it was daunting to consider initially increasing the staff. For this reason the question was posed 'Could the network be adapted for automatic working?'.

The largest of the many problems concerned the action to be taken when a mobile in a guard station area requests service. A working party investigating the question of automatic operation was unable to find an immediate answer. The Research and Development Division of British Rail considered the problem further and 4 weeks later demonstrated a prototype control system. The basic principles were unchanged, using Selcall sequencing in non-overlapping groups and selection of fixed stations based on 'best received' signal (voting), as before. The following additional features were incorporated:

(i) Signalling from the mobile to the control system—'Touch Tone' (CCITT Rec. Q23[3]) was chosen as suitable encoders and decoders were readily available and not expensive.

(ii) Indication to the mobiles of system availability by using an idle tone—known as free tone.

This prototype system operated as follows: Free tone is radiated by each fixed station, which is not in conversation and not a guard station. Since adjacent, overlapping fixed stations may be free at the same time, the free tone is radiated sequentially in groups in a similar manner to Selcall. A mobile requiring to make a call, turns on his loudspeaker by overriding the Selcall muting circuit. He will not hear free tone if he is in a busy area (where another conversation is in progress) or in a guard area, or if he is out of range of the nearest free fixed station. When the mobile operator hears free tone, which may be from more than one fixed station, he transmits and his carrier may be received by one or more fixed stations. The voting equipment will detect the best received fixed station, but to avoid operation by spurious carriers, the mobile operator keys a touch tone '✱' character and then the best received fixed station is selected. This radiates a dial tone, indicating to the mobile that dialling may begin. Free tone stops on every fixed station. The mobile then keys in the required number. The first digit keyed in selects the route—radio or telephone.

Digit '9' selects a radio to radio call and subsequent digits fill a Selcall register. When the complete number has been entered, the Selcall is sequenced on all the free fixed stations, including the first selected fixed station and its guards, if they were originally free. Carrier and touch tone '✱' from the replying mobile cause the voting system to select another (or if the replying mobile is in the same area, the original) fixed station. The audio paths of these fixed stations are then cross-connected and conversation may commence. Free tone recommences on all the remaining fixed stations except the guard stations of the fixed station(s) in conversation. The conversation is terminated by either of the mobiles signalling with a disconnect character (touch tone '#').

The British Rail trunk telephone network[4] is similar to the British Post Office telephone system, in which the trunk routing equipment is selected with an initial digit '0'. When the mobile keys an initial digit '0', this is recognized as a radio to telephone call and telephone exchange trunk equipment is seized. Subsequent digits dialled from the mobile are transferred to the telephone system and a conversation may be established. The conversation is terminated by either the mobile signalling disconnect ('#'), or cleardown of the telephone subscriber. For calls from the railway telephone network, the radio control system appears to the telephone network as another telephone exchange. Fortunately there are spare trunk dialling codes for all the radio systems envisaged. Signalling information from the telephone subscriber is routed in the normal way to the radio trunk exchange relay set, where the last four digits appear locally. (The British Rail trunk network can only accommodate a four-digit numbering scheme). Instead of the four digits being decoded by two-motion selectors as in a local telephone exchange, the digits are decoded and stored in the Selcall register. When the register is full, a Selcall sequence takes place on all free fixed stations. The Selcalled mobile replies, is selected by the voting system and the conversation is established and terminated as before.

In 1977, the manually-operated radio systems at Euston and Glasgow came into operation and it was apparent that an automatic connection system would have a further advantage. On the manual schemes, the time taken to make a connection was quite long, because it involved the operator logging and checking back the information. The operator also checked the identity of the called mobile or telephone subscriber. In a manually-connected system it is not easy to log call information automatically and supervisory systems are complicated. Call logging has however been found necessary in order to identify radio channel usage. This may be required for allocation of further mobiles to a particular department or for justification of further frequency allocations. An automatic connection system allows the mobile user to monitor the progress of his call and re-dial if necessary (as on an automatic telephone system) and call logging is obviously not difficult. It was anticipated that an automatic connection system would reduce the signalling and connection time considerably.

As a result of successful demonstrations of the prototype automatic radio connection system in 1976 and early 1977, a specification was drawn up, which included a number of extra features. Timers were added to regulate the signalling and conversation periods and along with these, supervisory tones were introduced. These tones are very similar to those used in the telephone system.

An important financial justification for the radio system came from its use by engineers concerned with the railway electric traction power distribution and control network. At present a dedicated, omnibus, long line telephone system is installed with access at strategic points along the railway line. This telephone system is terminated at the Electric Control Room. The telephones are used for isolation and safety procedures and are very necessary, but unfortunately very costly both in installation and maintenance.

By using radio and having a direct land line connection from the automatic radio control equipment to the Electric Control Room, it has now been possible to avoid the installation of these electrification telephones when new electrified railway routes are constructed. The Specification for the automatic radio connection system therefore incorporates direct access to and from the Electric Control Rooms. Each control room will have two access lines.

A further safety feature has been incorporated, in that any mobile may initiate an emergency call without waiting for free tone. This is similar to the British Post Office '999' emergency service. Connection is made to a continuously-manned office, such as a main station telephone exchange, where the operator may arrange for the necessary emergency assistance. Because the railway telephone emergency number is '111', it was decided to use the sequence '1111' for a radio emergency. When the digit '1' is received four times consecutively by any fixed station, an easily recognized emergency tone is transmitted by all fixed stations receiving the '1111' and all fixed stations, telephone and electric control room circuits in traffic. After a short period, this tone stops and the mobile initiating the emergency call keys a '✱'. The control system, using the voting equipment, selects the best received fixed station and extends the call to the emergency operator. It is not possible to select the best received fixed station immediately '1111' is received, because the received signals from other mobiles already in traffic may well be stronger than that from the emergency mobile. An emergency call can only be cleared down by the emergency telephone operator and may not be extended directly into the telephone system.

The initial demonstration automatic radio connection system used relays for speech and tone switching, with a hardwired, integrated-circuit control system. The demonstration equipment was only capable of one conversation at a time, although indication was provided showing which fixed stations were available for further conversations. Since, as with telephone exchanges, conversations are asynchronous, i.e. they start and terminate randomly, it appeared that if a second conversation was required, a second control system would also be required with access only to the free fixed stations. A second voting or selection system would also be required. If these techniques are considered for more than two conversations, the control system becomes quite complicated.

At this point, it was appropriate to review the control techniques used in modern telephone exchanges and it was found that stored program techniques offered advantages.

## 4 A Stored Program Automatic Radio Connection System (SPARCS)

In a typical radio area there may be up to 15 fixed stations, up to four trunk telephone circuits and perhaps two Electric Control Room circuits plus the Emergency Operator's circuit. Even though up to four simultaneous conversations may be in progress, very few control operations are required at any instant. Since Selcall sequencing can only take place for one call at a time, the Selcall generator may be common equipment. It was also recognized that even though quite complicated time-out procedures were necessary and that these would be at different phases for different conversations, each conversation would require only one timer, which could be allocated to the different timing operations.

The basic rules for each conversation are identical. A stored program approach to the radio connection system involves using a basic procedure, which is repeated for each conversation. Hardware performs speech and tone switching functions and in the system described in this paper the hardware has its own connection memories, i.e. once a connection path is set up, it remains until the control system changes it. Since the number of changes to the speech or tone connection paths is quite small during the call, it is possible to control a number of simultaneous, asynchronous conversations with relatively few instructions. A stored program control system using a microprocessor is ideal for this application and has the added advantage that the control system operation may be changed by modification to the software alone.

This is particularly attractive to British Rail, where up to 30 systems are required and in each there will be a different combination of fixed stations, telephone and Electric Control Room circuits required. The ability to change the pre-determined guard station or Selcall group allocations or to change the timers or other features, in the light of operational experience, is a distinct advantage over a hardwired control system. For this reason it was decided to build a prototype Stored Program Automatic Radio Connection System (SPARCS) with the following features:

(i) Input signalling

'Touch Tone' signalling is used from the mobile radio (or Electric Control Room console) into the control system. This system uses pairs of audio tones in the range 697 Hz–1633 Hz to represent each digit.

The system also requires information on the quality of the received signal, in order to select the 'best received' fixed station.

SPARCS receives digits from the trunk telephone system at 10 i.p.s.

(ii) Output signalling

Standard 5-tone sequential selective calling is used to alert the selected mobile. Signalling to the telephone exchange is again at 10 i.p.s.

(iii) Speech connection

Solid state analogue switches are used to connect the selected fixed station or Electric Control Room onto one of up to four audio highways, each of which is associated with a dedicated trunk telephone circuit. Where more than one telephone circuit is provided in the system, equipment in the telephone exchange hunts for the next free circuit into SPARCS.

Analogue switches are also used to apply the various supervisory tones.

(iv) Control system

The control system has three main tasks:

(a) Input signalling recognition.

(b) Connection path establishment.

This includes initiating output signalling, updating the system status memory and completing the connection path when the called party answers.

(c) Call supervision

This includes monitoring the call progress and terminating the call in response to timers or clear down requests.

In SPARCS the control system is centrally located and uses control cards, for each fixed station, Electric Control Room and telephone circuit. Each card is connected to three system busses. (See Fig. 1).

(i) The audio bus

This consists of four highways, each of which is used for the transmit and receive audio interconnections for one call.

(ii) The tone bus

All the supervisory and Selcall tones used in the system are generated on one card and distributed to all the control cards by this bus.

(iii) The digital bus

This provides a bi-directional control path between the microprocessor and all the control cards in the system.

The microprocessor interrogates the Touch Tone decoder on each fixed station, or Electric Control Room control card, or the 10 i.p.s. decoder on each telephone control card at regular intervals via the digital bus. The microprocessor also interrogates a card, which has digitally encoded the voting signals from the remote fixed station receivers. A buffer store is provided on each card to retain the last incoming digit until it is accepted by the microprocessor.

The microprocessor uses the digital bus to instruct the Selcall encoder and to operate the required tone switches. For outgoing telephone calls, data are transferred to a 10 i.p.s. dialling encoder.

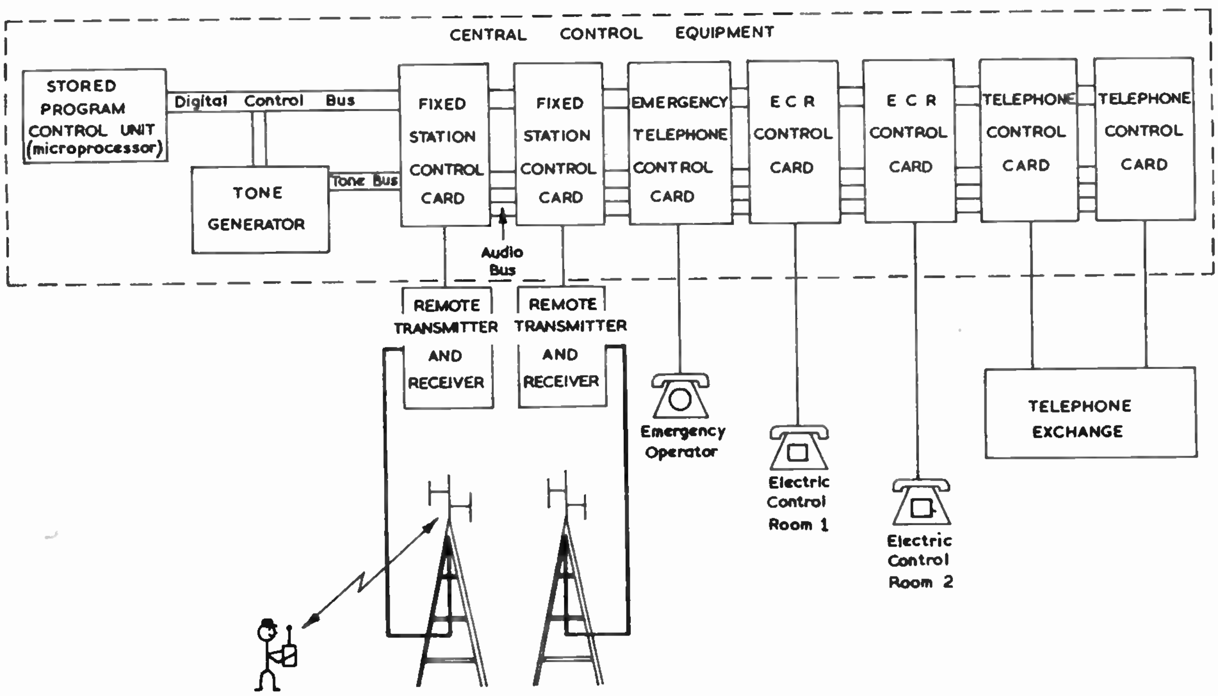


Fig. 1. SPARCS system schematic.

The microprocessor interconnects the control cards by operating their analogue switches to provide a speech path on a free audio highway. Supervisory tones may be added by operating the appropriate analogue switches. Since the telephone control card is allocated to a particular highway, talk-through connections may be provided by operating an analogue switch on this card to cross-connect the transmit and receive paths of the highway. A busy condition is applied to the telephone circuit to prevent incoming calls. This busy condition is also applied for all calls on that highway which do not involve the telephone circuit. Various special facilities have also been provided:

(a) Electric Control Room (E.C.R.) access
This is provided via a dedicated, fixed station control card with a land line direct to the Electric Control Room, where there is a control console, which appears to the system as a combined mobile/fixed station. In addition to the standard mobile facilities, extra system status information is provided on the console. Lamp indications are given of: 'system busy,' 'system free' and 'system partially busy' (i.e. some fixed stations are free). Each E.C.R. console has two lines into the control system to improve system availability, since there is usually more than one operator. Certain electric traction supply isolation procedures require continuous communication between two mobiles and the E.C.R. operator. Having established a connection with one mobile, the E.C.R. operator may initiate a second Selcall sequence. Since E.C.R. calls may involve safety of life, all calls involving the E.C.R. can only be terminated by the E.C.R. operator.

(b) Emergency operator
Access to an emergency operator is provided via a telephone control card, although the dial pulse encoding and decoding circuits are not used. Again for obvious reasons, the call cannot be terminated by the mobile.

(c) Automatic testing
An alarm output is provided on each fixed station control card, which may be operated by the microprocessor, when necessary, if a software controlled test should fail. This test involves one or more remote, transponding, fixed, mobile equipments, which can be accessed by each fixed station in turn under software control. This back-to-back test is made at regular intervals.

A modular constructional technique has been adopted, using as few different types of card as possible to minimize costs, facilitate easy expansion and simplify maintenence. Since all control cards of one type are completely interchangeable (all card addressing is set by two wire links on the card rack), only a few spare cards are required for first line maintenance.

A Motorola 6800 microprocessor system was chosen, because it is one of the current industry standards and British Rail already had considerable expertise available on this system within the Research and Development

Division. Development effort has been saved by utilizing 'off the shelf' microprocessor system cards and it is economic to provide a second complete microprocessor system for standby purposes.

## 5 SPARCS Software

In a stored program control system, there are two possible approaches for the acceptance of input stimuli:

(a) An interrupt system, in which the program only runs when necessary and is started or interrupted by an external stimulus.

(b) A scanning system, in which the program scans all inputs at regular intervals, actions any valid input requests and then returns to scanning operations.

The scanning approach was adopted for SPARCS for two reasons:

(i) There may be several requests for service originating nearly (but not exactly) simultaneously. An interrupt system would need to allocate priorities to each of these inputs, and complicated situations can develop.

(ii) The input data rate is relatively low ($<$ 100 ms/digit for both touch tone and telephone dialling inputs) and it is relatively easy to sample all inputs at a faster rate, whilst still performing any necessary processing before the next scan cycle.

A scan interval of 25 ms was chosen. Approximately 3 ms are required to interrogate up to 24 touch tone decoders, 24 encoded voting inputs (i.e. 20 fixed stations and 4 E.C.R.s), four trunk telephone circuit inputs and an emergency telephone input. The remaining time in the cycle is more than adequate to carry out all further processing, which may include sequencing of free tone or Selcall, updating the system status memory and setting up the necessary connections and signalling for the calls in progress. Traffic analysis and recording may also be carried out during this period, but may be interrupted by the start of the next scan sequence and resumed later. The 25 ms scan interval is synchronized to a signal derived from a crystal-controlled timing source and is also used both for the call timers and an internal (software) real time clock, used for logging purposes.

The software used in SPARCS is written in a modular form, i.e. each module of program has a specific task and may be modified if necessary without affecting any other program modules. Specific areas of memory are allocated to system status information. Program modules and memory areas are illustrated in Fig. 2 which also shows the main program flow. The individual modules are as follows:

(a) Initialization Program
This part of the program is used only when the system is first switched on and is responsible for presetting all the control cards to a known state. In addition, information is transferred from a read only memory into the working system status memory. This information is subsequently used to set the system timers and to set the desired pattern of guard stations for each fixed station. The configuration of fixed stations, E.C.R.s, and telephone circuits is also transferred from this r.o.m. into the working memory area. To change any part of the system configuration it is only necessary to change this initialization r.o.m.

(b) Input Scan Program
It was recognized early in the development of the program that many operations were of a routine nature and did not require a complex decision process. These operations include the input scan routine, the sequencing of free tone and Selcall

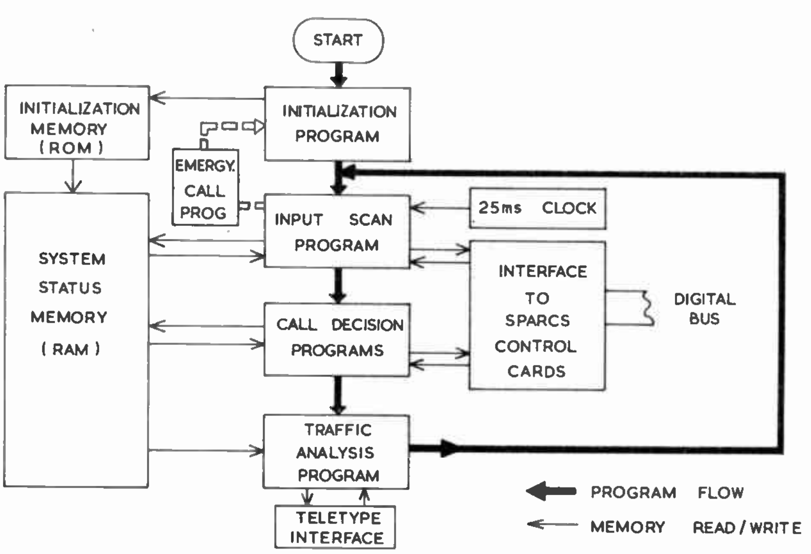


Fig. 2. SPARCS software organization.

and housekeeping routines such as the setting and resetting of guards. The implementation of these routines is controlled by reference to the system status memory, e.g. free tone is not applied to a busy or guard fixed station. The input scan program reads data from all the fixed station (and E.C.R.) touch tone decoders and encoded voting inputs every 25 ms. Any valid data is stored in the memory area allocated to that fixed station. This data is then sorted depending upon the system status and the status of the particular fixed station (or E.C.R.) at the time, and made available to the Call Decision Program. Incoming emergency calls are recognized by the input scan process and passed to the Emergency Call Control Program.

(c) Call Decision Program
This program performs output switching functions depending upon the system status, taking input data from the Input Scan Program module. The first free highway is allocated to process a new call originating from a fixed station or E.C.R. Subsequent input data relating to that call is recognized by this Call Decision Program and the appropriate action taken. This action may take one or both of the following forms:

(i) Direct instruction to the appropriate control cards to perform the required output functions, e.g. speech or tone switching, or
(ii) Updating the system status memory, from which it is possible to initiate routine operations such as selective calling, or the setting of guard stations, via the Input Scan Program in the next 25 ms cycle.

Since each telephone control card is associated with a particular highway, the allocation of a new call from the telephone network to a free highway will automatically be made by the telephone equipment finding a free telephone control card. The Call Decision Program reads the incoming signalling information directly from each telephone control card. Subsequent processing of this information is similar to that described for inputs originating from a fixed station or Electric Control Room.

The Call Decision Program maintains a location in memory, which is used as a timer for calls on each highway. The time, in seconds, required for a specific timer function is loaded into this memory location and decremented by the program once every second. This program also checks for a zero in the timer memory every scan cycle. When this occurs, appropriate action is taken.

(d) Emergency Call Control Program
When the Input Scan Program recognizes an incoming emergency call ('1111'), the appropriate fixed station, E.C.R. and telephone control cards are instructed to transmit emergency tone for a period as described previously. At the end of this period, the program flow is diverted to an Emergency Call Control Program, which selects the fixed station initiating the emergency call and alerts the emergency operator via the Emergency Telephone control card. Normal program flow is discontinued until the Emergency Call Control Program recognizes clear-down from the Emergency Telephone control card. The program is then restarted at the Initialization Program.

(e) Traffic Analysis Program
The time taken for the Input Scan and Call Decision Programs to run is variable depending upon the action required during each program cycle. This time is typically 10 ms, leaving 15 ms available for other purposes. This spare time may be used for traffic analysis and communication with external devices such as teletypes and data recorders. In addition to providing hard copy of traffic statistics, a teletype may be used to key in information to the system in order to make any temporary changes, which may be required 'in the light of operational experience'.

Changes may be made to timers, guard station allocations, free tone and Selcall sequence groups etc.

### 5.1 Summary of Program Flow

At switch-on, the system is initialized to predetermined conditions by the Initialization Program. The Input Scan Program then waits for the next 25 ms clock signal before starting the free tone sequence and scanning each fixed station and E.C.R. control card for input data. This data is validated by reference to the system status memory and passed on to the Call Decision Program, which scans the telephone control card inputs before making any necessary decisions and taking appropriate action. In the absence of any valid input data, the Call Decision Program steps on to the Traffic Analysis Program, which then returns to the Input Scan Program. The cycle recommences with the next 25 ms clock signal.

## 6 The Future

The prototype SPARCS system is operating at Derby using a number of fixed stations and two trunk telephone circuits. Railway telephone users anywhere in Britain can access mobile radios in the system and vice versa.

As a result of the development, a number of minor shortcomings in the specification have been recognized and the specification has been amended. Several stored program automatic radio connection systems are now being manufactured for installation in 1980 and it is anticipated that there will be approximately thirty systems in operation by 1986.

Development continues on traffic analysis methods and examination of the techniques required to maximize the advantages of using more than one channel in a radio area. These techniques include the use of a calling channel, channel marking, intelligent mobile radio equipments and off-air call set-up.

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(*Paper No. 1944/Comm 198*)

UDC 621.317.34: 621.376.56: 621.833.4

*Indexing Terms:* Pulse code modulation, Telephony networks, Distortion measurement, Quantizing distortion, Companding

# A review of distortion and its measurement in p.c.m. telephony systems

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*Based on a paper presented at a Measurements and Instruments Group Colloquium on Audio Distortion and Measurement held in London on 27th April 1978*

**SUMMARY**

Pulse code modulation systems as used in telephony networks are briefly described with particular regard to the inherent quantizing distortion (q.d.). Test methods for the evaluation of signal to q.d. ratio, from analogue input to analogue output, are reviewed. Three test stimuli are considered and their advantages and disadvantages discussed. The three stimuli are a sinusoid, a band-limited noise signal and a band-limited pseudo-random noise signal.

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## 1 Introduction

The use of pulse code modulation (p.c.m.) for transmission of signals in telephony systems is increasing throughout the world. Among important reasons for this increase are the inherent resistance of a digital transmission link to corruption from noise and the fact that transmission quality is almost independent of circuit length. Signals can therefore be transmitted long distances without any distortion being added. However, this is achieved at the penalty of some signal corruption, namely quantizing distortion, which is introduced in the analogue-to-digital conversion circuits. A compromise has to be made between an increase in the number of bits per channel to minimize distortion, and a reduction in the number of bits to keep transmission frequency low. This paper reviews the techniques presently used in the measurement of quantizing distortion to determine the quality of a p.c.m. system from analogue input to analogue output.

It should be noted that throughout the paper quantizing distortion (q.d.) is referred to as the measured quantity; it would strictly be more accurate to refer to the measured quantity as total distortion, but in general q.d. is the most significant part.

## 2 P.C.M. Systems

Before the measurement methods are described, a brief explanation of a p.c.m. system follows for those not familiar with the technique. More detailed explanations of the system[1] and an analysis of quantizing distortion[2] may be found elsewhere.

The description is based on a 30-channel system as specified in CCITT (The International Telegraph and Telephone Consultative Committee) G700 series of recommendations.[3]

Figure 1 shows a simplified diagram of the system with typical waveforms.

Each input from 30 speech channels is sequentially sampled at an 8 kHz rate. These samples are converted to 8-bit binary coded words representing the amplitudes of the pulses. This process is known as quantizing. As there are only 256 words (pulse amplitudes) available with 8-bit encoding, then the value of the binary code may not exactly represent the pulse amplitude, the difference being the cause of the distortion inherent in p.c.m. systems and known as quantizing distortion. The allocation of binary-coded words to levels is made on an approximation to a logarithmic law which gives a substantially constant signal-to-distortion ratio for a wide range of input levels.

The binary coded sample is interleaved with coded samples from the other channels in a time-division multiplexing manner and together with signalling and framing information in two other channels is transmitted over a cable system using a high density (HDB3) line code.[1,7] Frame words are inserted into the digital stream



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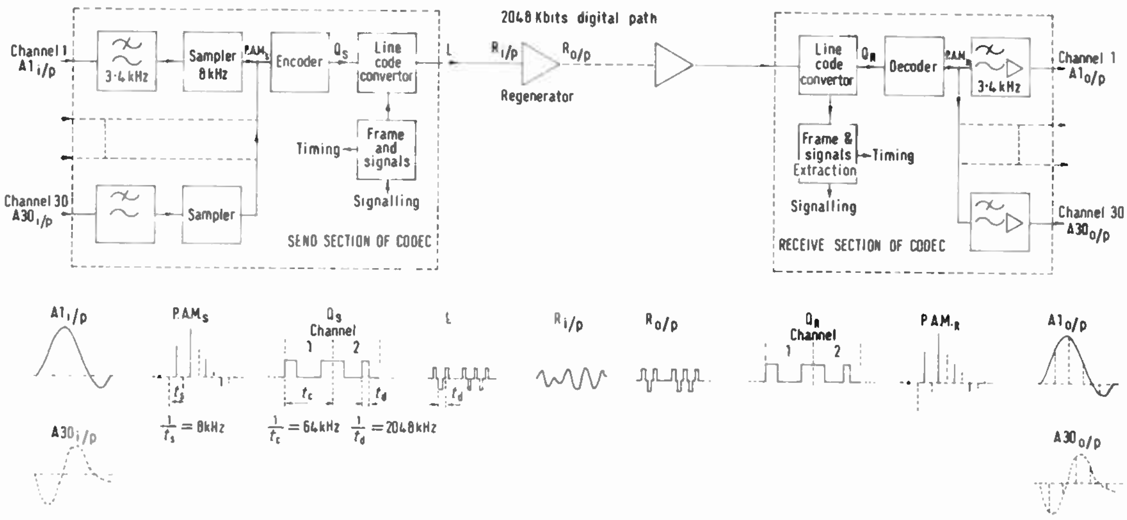


**Fig. 1.** Simplified schematic diagram of a 30-channel p.c.m. system with typical waveforms.

at regular intervals to enable the receive terminal to lock onto the stream and correctly locate the position of the channel words; a similar multiframe structure is necessary for the signalling on each channel.

The digital stream is distorted by the cable and at intervals of typically 1800 metres the signal has to be regenerated. As the regenerator has only to determine whether a 0 or 1 should be retransmitted, it has therefore a high noise immunity and, within the bounds of error rate and jitter, no further distortion is added to the coded speech signal. Typically the digital path distances are 15 to 25 kilometres using around 10 regenerators per route.

At the receive terminal a reverse process, converting the binary coded words into analogue signals, separating the channels and low-pass filtering, produces an analogue signal very similar to the original input. Another p.c.m. system in the reverse direction provides a return path. The receive equipment also monitors the frame structure in the receive digital stream to check for errors in the expected signals; in the event of a major fault condition being detected such as loss of frame alignment, the codec alarm will take the system out of service.

## 3 Sampling and Coding

The forerunner of p.c.m. was pulse amplitude modulation (p.a.m.) which involves sampling the signal at a rate greater than twice the maximum frequency to be transmitted (Fig. 2). The resultant frequency spectrum for a sampled speech signal is shown in Fig. 3, and it can be seen that the original signal can be reconstituted if the sampled waveform is passed through a low-pass filter. Having sampled the waveform it is possible to utilize the spaces between the samples for other channels—this is time division multiplexing (t.d.m.).

A signal of this format is not convenient because transmitting it over a cable would result in rapid signal degradation including phase and amplitude distortions caused by the line and added noise from crosstalk with other adjacent systems.

P.c.m. is developed from p.a.m. by quantizing each sample into a series of binary pulses. For example, using

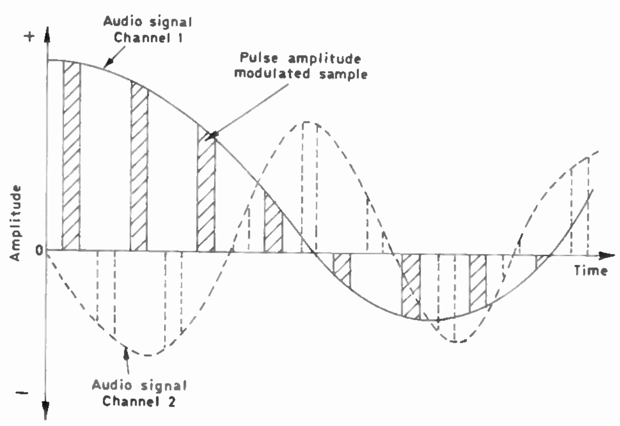


**Fig. 2.** Pulse amplitude modulation and time division multiplexing of two channels.

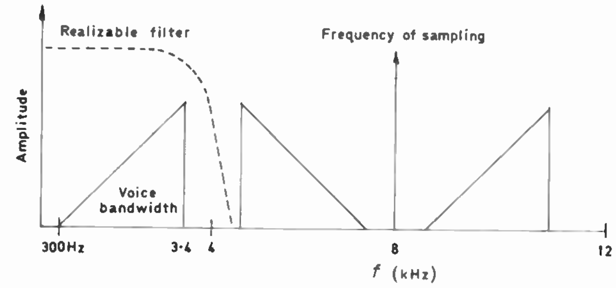


**Fig. 3.** Frequency spectrum of pulse amplitude modulated signal.

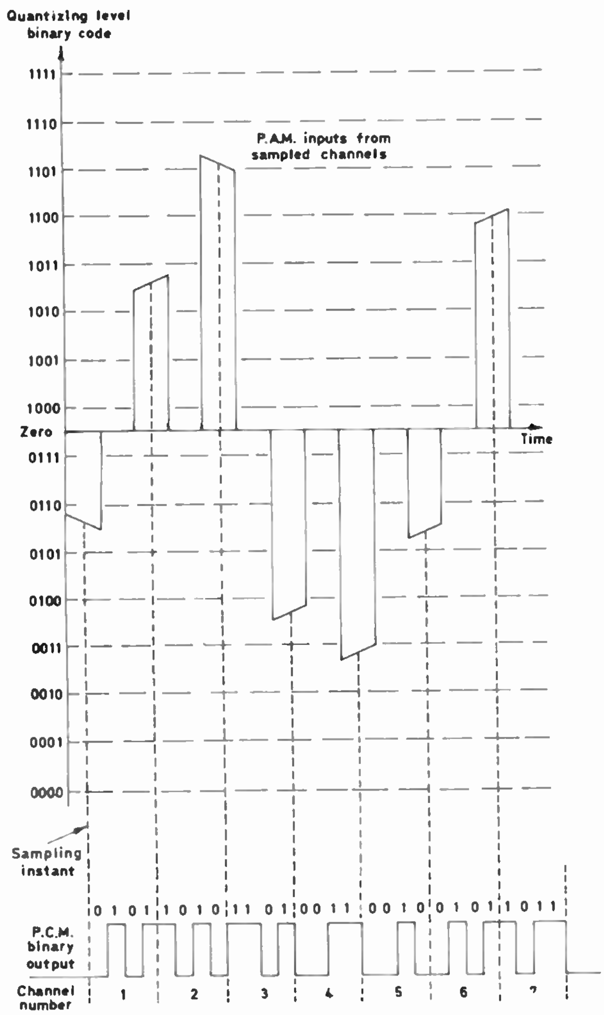


Fig. 4. Converting pulse amplitude modulation to pulse code modulation.

four bits per sample, 16 different amplitude levels can be specified (Fig. 4). Any sample not having a value corresponding to a specified level will be represented by the nearest quantizing level. When the signal is reconstituted these approximations will inevitably give rise to distortion which is known as quantizing distortion (q.d.).

It is unrealistic to represent the infinite number of amplitude values of speech with only 16 levels. The quantizing noise would be so great that the signal would be almost unintelligible. For speech a signal to q.d. ratio of at least 25 dB is desirable.

In order to achieve this figure for both loud and quiet talkers at least 1024 levels or 10 bits would be needed. However, if more levels are allocated to the quiet talker and less to the loud talker an acceptable result can be obtained using only six bits, in practice 8 bits are usual. This non-linear allocation of levels is known as companding and is shown in Fig. 5(a). If a logarithmic companding law is used, then an almost constant signal-to-q.d. ratio can be obtained (Fig. 5(b)) over a wide range of signal input levels.

The companding law used in the CCITT 30-channel system is known as an '$A$' law and is an approximation to the continuous function:

$$y = \frac{1+\log(Ax)}{1+\log A} \quad \text{for } \frac{1}{A} \leqslant x \leqslant 1$$

(max. input level = 1)

$$y = \frac{Ax}{1+\log A} \quad \text{for } 0 \leqslant x \leqslant \frac{1}{A}$$

where $A = 87{\cdot}6$.

A segmented approximation to this continuous function can be drawn, such that each successive segment

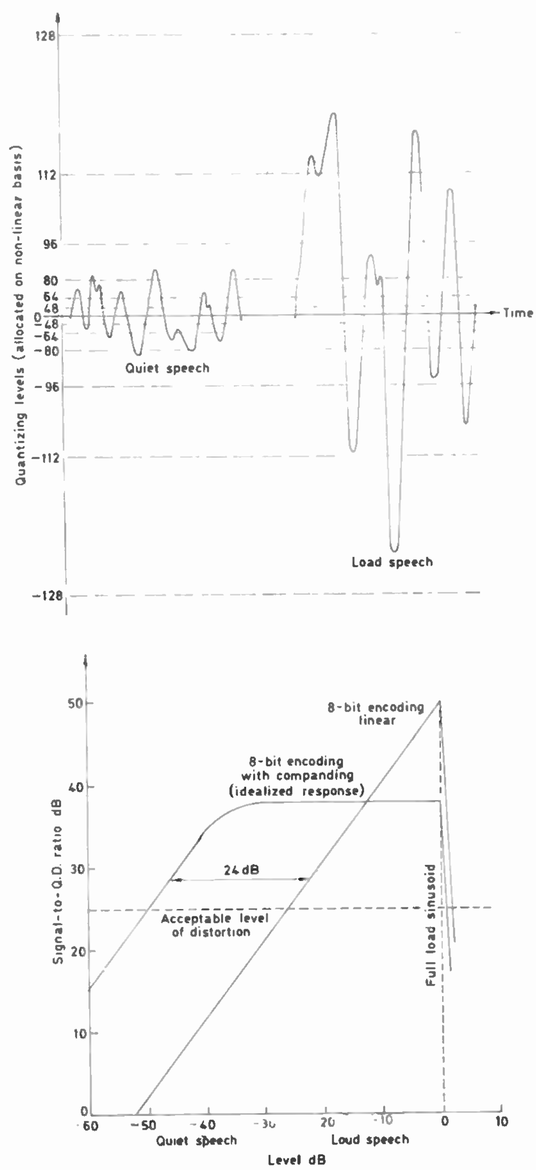


Fig. 5. Companding in p.c.m. system showing (a) companding of speech and (b) signal to quantizing distortion ratio against input speech level.

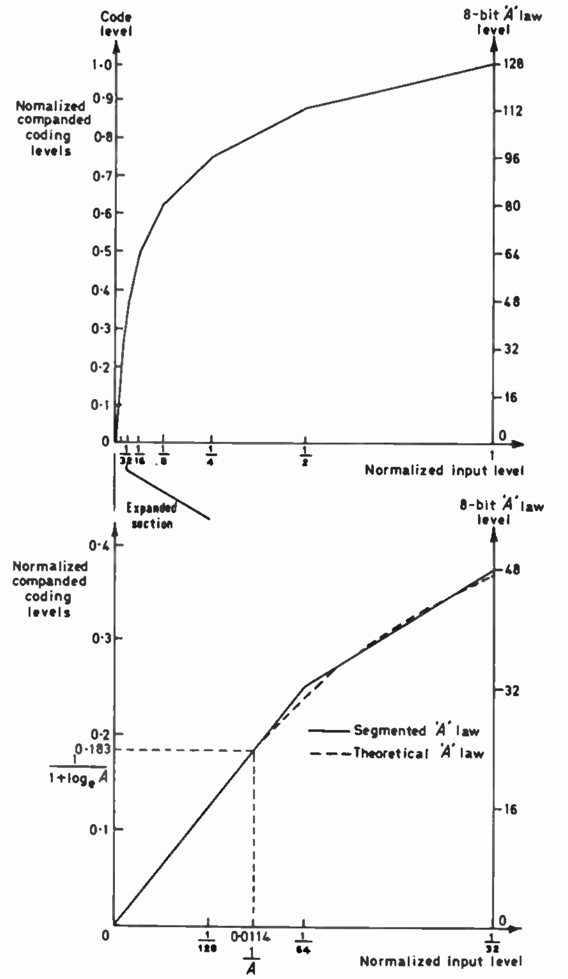


Fig. 6. *A* law companding.

changes its slope by a factor of 2. For positive values there are 7 segments, the first segment being twice as long as the other segments. Figure 6 shows the relationship between signal values at the coder input and the coder decision values. If both positive and negative values are considered there are 14 segments of which 2 (one each side of zero) are colinear giving 13 linear segments overall.

The $y$ axis in Fig. 6 represents the number of decision values resulting from the allocation of 8 bits to each encoded sample.

The positive input range is allocated 128 decision values requiring 7 bits ($2^7 = 128$). The negative input range requires a further bit, making a total of 8 ($2^8 = 256$). In practice one bit is used to indicate the polarity of the sample, while the remaining 7 bits are used for both positive and negative samples.

It will be seen that the relationship between the $x$ axis and $y$ axis is such that the range 0 to $\frac{1}{64}$ of the total input voltage is represented by 32 decision values (two co-linear segments). For the range $\frac{1}{64}$ to $\frac{1}{32}$ 16 decision values are used and the range $\frac{1}{32}$ to $\frac{1}{16}$ uses the next 16 decision values and so on up to the last segment where the range $\frac{1}{2}$ to 1 is represented by the last 16 decision values. In this way the relationship between relative input voltage and the number of decision values is changed by a factor of 2 for each successive segment. The relationship between a uniform coder and an equivalent non-uniform coder, both having the same dynamic range at the analogue input, is commonly known as the companding advantage.

This may be expressed as:

$$20 \log_{10} \frac{N}{n} \quad \text{dB}$$

where $N$ = number of decision values of the uniform coder,

$n$ = number of decision values of the non-uniform coder.

By way of example for the $A$-law characteristic ($A = 87{\cdot}6$) the relationship is:

$$20 \log_{10} \frac{4096}{256} = 24{\cdot}1 \text{ dB}.$$

The most accurate method of companding is done digitally. The sampled waveform is linearly coded into say 12 bits and then digitally compressed into 8 bits. At the receive terminal it is necessary to reform the signal by passing it through an expander circuit having a complementary characteristic.

The theoretical distortion is discussed fully by Cattermole.[2] However, the approximate signal to q.d. ratio ($S/N$) for a Gaussian signal is:

$$10 \log_{10} \frac{3N^2}{(1+\log A)^2} \quad \text{dB}$$

where $N$ is the number of levels.

Substituting

$$N = 256 \text{ for 8-bit coding}$$

and

$$A = 87{\cdot}6 \text{ for CCITT 'A' law}$$

this gives $S/N = 38{\cdot}2$ dB.

The selection of system parameters to give 38 dB signal to q.d. ratio rather than 25 dB previously mentioned allows for several systems to be connected in series whilst maintaining an acceptable distortion level. In practical systems an allowance has to be made for distortion added by the imperfections in the converters for send and receive sections, and the allowable q.d. is some 4 or 5 dB worse than theoretical. For example, in the 30 channel CCITT system the $S/N$ ratio allowable is 33·9 dB using a Gaussian test signal measured in a telephone channel bandwidth of 300 Hz to 3·4 kHz.

Other companding laws are used but none have any major advantage over the $A$ law companding and give similar results on q.d.

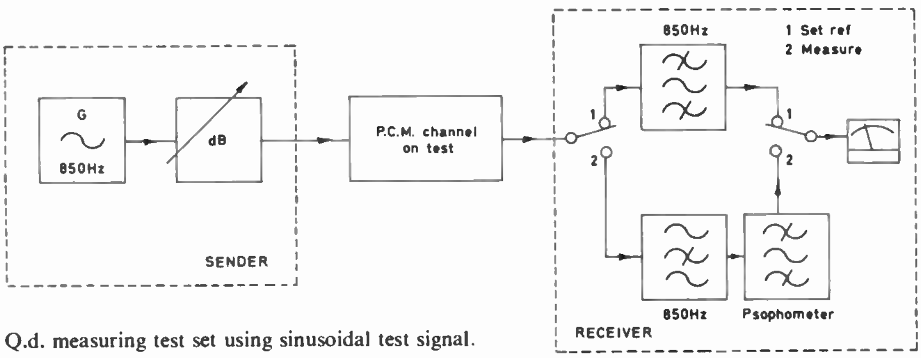


Fig. 7. Q.d. measuring test set using sinusoidal test signal.

## 4 Measurement of Distortion

To test p.c.m. systems two main types of instrumentation are required; these are for checking codec performance and for checking digital paths.

Most tests of codec performance are made, at present, by looping the receive and send sections back to back and performing tests on each channel. Checks on performance include: gain, linearity, frequency response, distortion, noise and crosstalk. Testing is carried out using analogue test signals and measuring equipment common in many respects with that used for frequency division multiplexing (f.d.m.) systems.

The measurement of distortion will now be considered in detail with particular reference to three types of test stimuli, namely a sinusoid, noise and pseudo-random noise.

## 5 Sinusoidal Test Signal

The most obvious and simple test for distortion is to use a sinusoidal test signal and measure the distortion produced at the output of the p.c.m. system under test.

A block diagram is shown in Fig. 7 for the test system and results from such a set-up are shown in Fig. 8. The particular system tested had a 13-segment *A* law companding with 7-bit coding. The distortion produced clearly shows the discontinuities between individual levels and the changes of companding law slope at 6 dB intervals of input level. For 13-segment *A* law companding with 8-bit coding, the mean trend will be a 6 dB higher q.d. figure and the discontinuities will be doubled but the excursions at the discontinuities will be halved.

The test frequency indicated in Fig. 7 is 850 Hz but frequencies in the range 300 to 1100 Hz have been used by various telecommunication authorities, a generalization being that the frequency should be in the lower end of the voice frequency band.

This method is very exact for determining the performance of the individual quantizing steps and segment slope changes but has two major disadvantages:

(a) The sinusoidal test signal does not represent the normal speech signal applied to the p.c.m. link.

(b) The great number of measurements that need to be made to determine the exact q.d. to input level response. For example, with 7-bit coding (64 positive and 64 negative levels) and allowing 4 measurement levels per step to define the q.d. between steps accurately, then 256 $(=4\times 64)$ measurements will need to be taken, and with 8-bit coding this would increase to 512 measurements.

If fewer measurements were taken, for example 2 dB steps between input levels of +2 and −46 dB (25

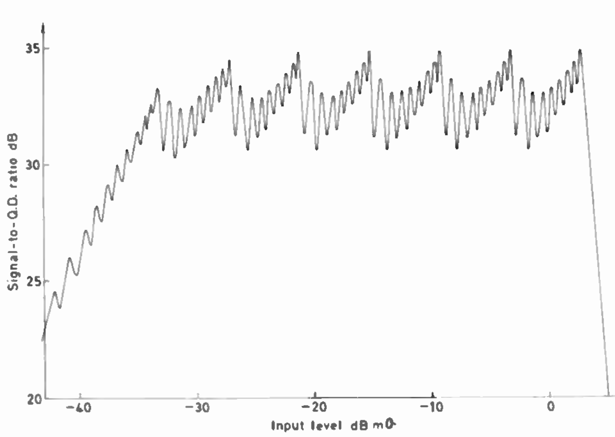


Fig. 8. Q.d. measurement with a sinusoidal test signal on a 24 channel p.c.m. system with *A* law companding (7 bits/sample).

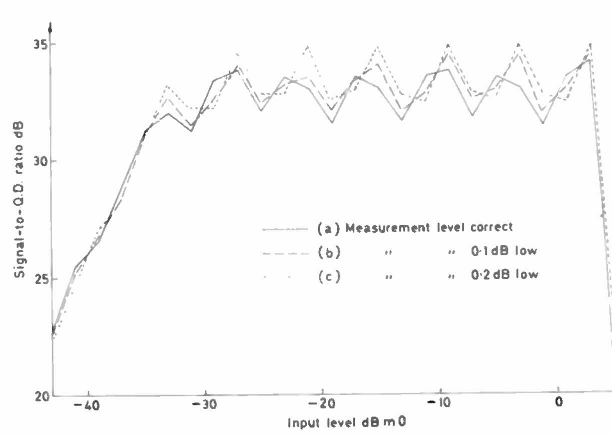


Fig. 9. Q.d. measurements as in Fig. 8, but with measurements taken at 2 dB intervals.

measurements), then a curve as shown in Fig. 9(a) would be obtained. If the results were repeated at some other time when the level had shifted by 0·1 dB or 0·2 dB because of test level change or small change in p.c.m. coder or decoder performance, then the curves of Fig. 9(b) and 9(c) respectively would be obtained. From these it can be seen that discrepancies of up to 2 dB or more in the q.d. measurement can occur. The results shown in Fig. 9 were taken under very tightly controlled test conditions as is apparent by the easily discernible repetition of results at 6 dB input level intervals. A set of results taken in a typical test situation, with minor variations due to attenuator inaccuracies, etc., would give an even more random scattering of results.

Because of the discontinuity changes, practical measurements taken on a day-to-day basis can give apparently different results unless a great number of results are taken, it is therefore desirable to use a test signal which will give a smoother q.d. to input level response and at the same time simulate speech more closely.

## 6 Noise Test Signal

Studies of the characteristics of speech have given rise to two main models for speech:

(a) A model using a noise signal with an inverse exponential distribution of amplitudes.
(b) A model using a noise signal with a Gaussian (or normal) distribution of amplitudes.

Theoretical studies using both models for speech applied to a p.c.m. system with a logarithmic companding law show a maximum variation between the two models of 0·3 dB at one part of the q.d. to input level response. The use of either speech model gives a q.d. to input level response which is a smooth curve when plotted. The theoretical q.d. for both models can be computed, but the more practical type of noise to produce is that with a Gaussian distribution and because of this the preferred test signal is a noise signal with a Gaussian distribution of amplitudes.

Figure 10 shows the method of test. The bandwidth of the noise will have no real significance on the test result providing it is small compared with the system input bandwidth and is typically 100 Hz to 200 Hz centred on 450 Hz. The use of a narrow band-pass filter in the receive side of the test set, to enable the input signal level to be measured, and a reference set, are only essential when large q.d. levels are present and need not be defined as closely as the send filter and measure filter. The general shapes of test curves using a noise test signal are shown in Fig. 11.

The advantages of noise as a test signal are:

(a) The q.d. to input level response when plotted gives a smooth curve and small day-to-day changes of test level or system performance will give a curve similar to the original measured response. A minimum number of measurements are required to define the response correctly: some 20 or 30 measurements over the entire input level range being more than adequate.
(b) The noise test signal has properties similar to speech (over a long period of time).

But disadvantages are:

(a) The essentially random nature of true noise having Gaussian distribution dictates that for the purpose of measurement a long integration time is required for the meter circuit. An integration time of the order of 100 seconds is necessary to reach a measurement certainty of 0·1 dB. The long integration time has distinct disadvantages in the practical measurement of q.d. A typical set of measurements on one channel could take up to 2 hours if an accurate response is required.
(b) A truly random noise source is not practical because certain limitations must be present and these ought to be defined if test sets are to be compatible. An obvious limitation is a limit to the peak-to-r.m.s. ratio with peak clipping on high peaks. Another is that although filtering of noise with Gaussian distribution should not theoretic-

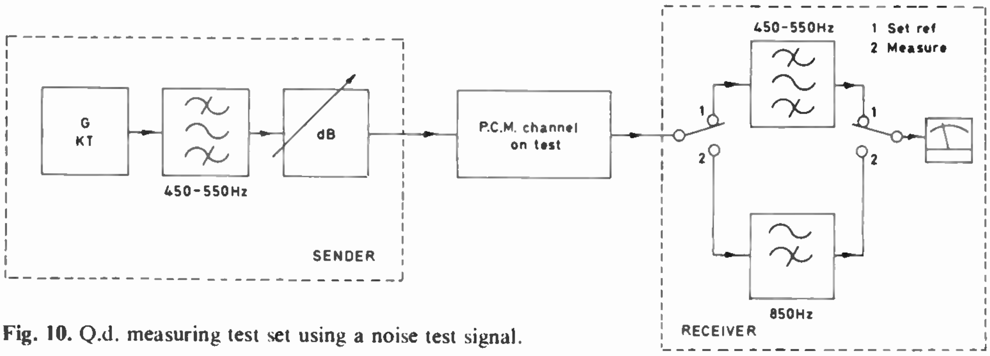


**Fig. 10.** Q.d. measuring test set using a noise test signal.

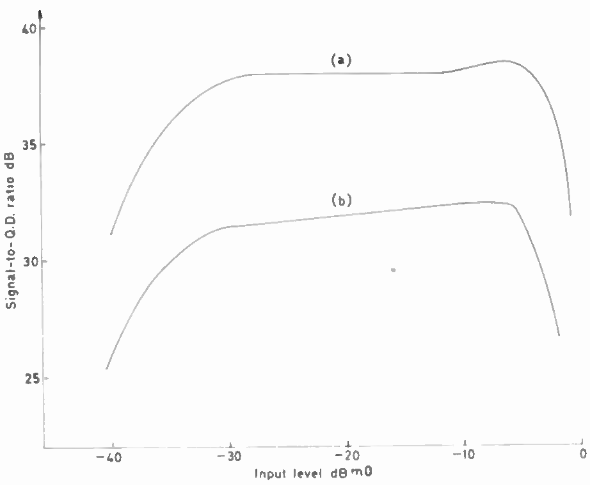


**Fig. 11.** Typical q.d. measurements with a true noise test signal.
(a) 30-channel p.c.m. system with $A$ law companding (8 bits/sample).
(b) 24-channel p.c.m. system with $A$ law companding (7 bits/sample).

ally change the distribution, the extraneous effects of large amplitude spikes on ferrite cored inductors is only one of the possible undesirable effects. To overcome the above difficulties pseudo-random noise is used as a test stimulus for p.c.m. systems replacing the true noise source shown in Fig. 10.

### 7 Pseudo-random Noise Test Signal

It is possible to produce a signal simulating noise by digital techniques. This is done by connecting a shift register $n$ stages long to produce a maximum length binary sequence $(2^n - 1)$ bits long with a clock frequency $f_c$. The output power density will be as in Fig. 12 with spectral lines occurring at $f_c/(2^n - 1)$ Hz intervals. The output is then filtered by a narrow-bandpass filter to produce a pseudo-random signal which approximates to the true noise signal previously discussed.

The use of pseudo-random noise has the following desirable properties facilitating the construction of a measuring set.

(a) It is easy to generate using a shift register connected so that it produces a maximum length pseudo-random binary sequence.
(b) If the supply voltage to the output stage of the shift register is stable the r.m.s. value of the output can be held constant.
(c) All measuring sets using a particular pseudo-random noise generator will apply the same test, the characteristics of the tester being repeatable.
(d) The time taken to make a reading compared with a true noise signal is considerably reduced, the noise being a defined repetitive sequence having a period typically less than 0·5 s for the requirement being considered.

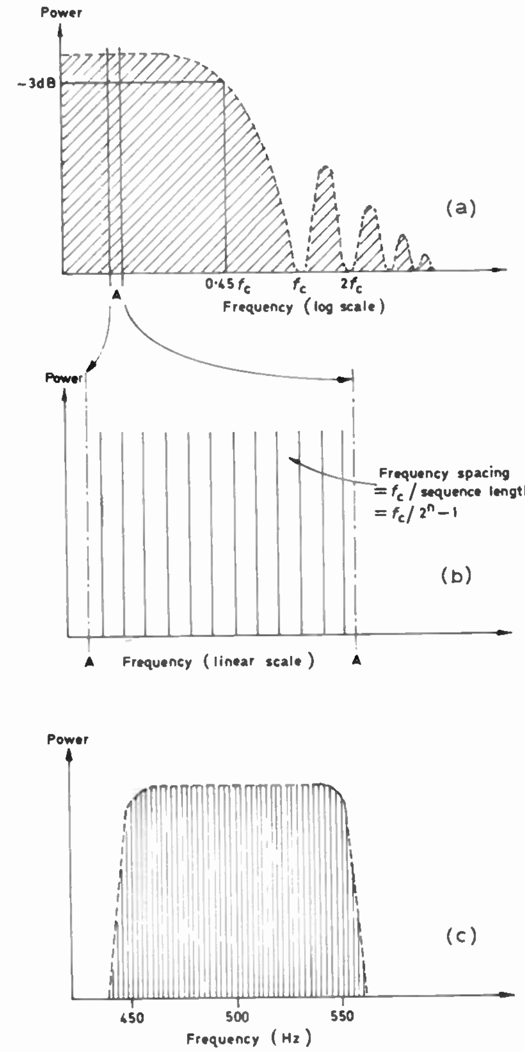


**Fig. 12.** Spectrum of pseudo-random binary sequence.
(a) Unfiltered sequence.
(b) Expanded section of (a).
(c) Output after filtering in 450 to 550 Hz bandwidth. ($f_c$ = 340 kHz, $n$ = 17 stages).

However, in order to utilize these desirable properties in a tester for measuring q.d. it is necessary to examine the requirements carefully. The required test signal should simulate a narrow band limited true noise signal. An examination of a true noise signal reveals three significant parameters:

(a) The amplitude probability density distribution of true noise is Gaussian.
(b) The theoretical peak-to-r.m.s. ratio of true noise is infinite, although in practice it rarely exceeds 13 dB and the published charts on Gaussian distribution show that the probability of this being exceeded is 0·002%.
(c) There are no dominant frequencies present within the frequency band considered.

It is therefore necessary when choosing a pseudo-random noise source to consider the peak-to-r.m.s.

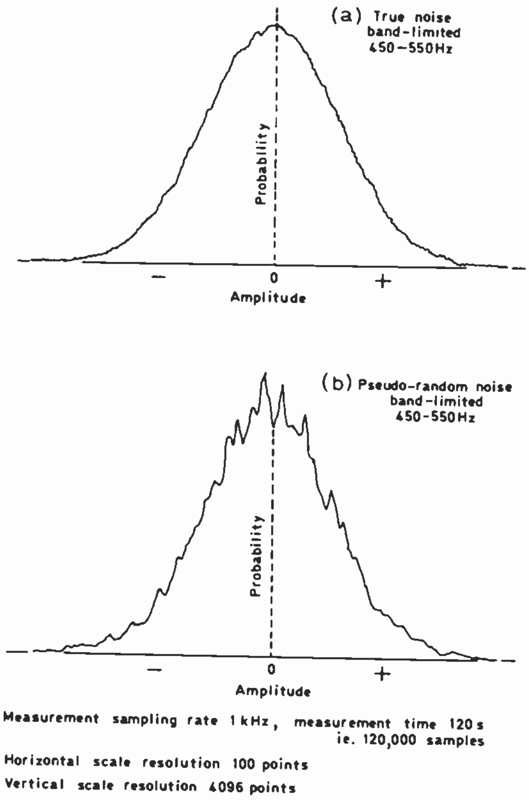


Fig. 13. Amplitude probability density distribution of pseudo-random noise measured over a period of 120 seconds. Noise source is the example given in Section 7.

ratio, probability density distribution and frequency distribution of the test signal produced. Extensive tests have been carried out to select the recommended pseudo-random signals now accepted internationally.

The requirements of an acceptable signal are:

(a) The word (sequence) length must be sufficiently short so as to allow the measurement to be made quickly. The word length is $(2^n - 1)$ multiplied by the clock period.
(b) The spectral lines (frequency components) must be close enough to simulate noise adequately and also have equal average levels.
(c) The peak-to-r.m.s. ratio must approximate to that of true noise.
(d) The amplitude probability density distribution must be Gaussian.

There is a conflict between requirements (a) and (b) as the spectral line spacing is the reciprocal of the word length, consequently a reasonable compromise must be made.

The vital statistics of filtered pseudo-random noise are very difficult to determine theoretically as there is no simple mathematical formula relating clock frequency, sequence length, feedback connection, filter characteristics and the resultant output. A numerical evaluation could be performed using a computer but the process would be lengthy, expensive and probably futile. An alternative method is to make a series of measurements using a voltmeter and oscilloscope to assess the peak-to-r.m.s. ratio and a probability density analyser to plot the distribution. The peak-to-r.m.s. ratio chosen was less than that of true noise because of the finite length of the noise 'word'. If a peak-to-r.m.s. ratio of 12·5 dB had been chosen it would be necessary to have a sequence with a length of at least 1000 seconds as there is only a probability of 0·0002 that this peak would occur with true noise.

The practical group of signals internationally accepted all have sequence lengths of about 200 to 400 ms and peak-to-r.m.s. ratio of 10·5 dB.

If the probability density distribution of any 300 ms of band limited (350 to 550 Hz) white noise is examined, it is extremely unlikely that it will be Gaussian. In fact it is almost safe to say that it can never be truly Gaussian. This is because there can only be a maximum of 165 cycles during this period and thus there is only a discrete number of amplitude levels. Pseudo-random noise with a defined word length, therefore, can only produce an approximation to true noise. In addition to this limitation the output will be Gaussian only if all the parameters are correct.

One example[4,5] of a sequence which gives an acceptable performance has the following parameters.

(a) A 17-stage shift register clocked at 340 kHz. This gives a word length $L$ in seconds of:

$$L = (2^{17} - 1) \times \frac{1}{340 \times 10^3}$$

$$= 385 \text{ ms}$$

and a spectral line spacing $S$:

$$S = 1/L = 2{\cdot}6 \text{ Hz.}$$

(b) A filter bandwidth of 130 Hz centred on 460 Hz.

pass band 395 Hz to 525 Hz

no. of spectral lines $= 130/S$
$= 50.$

(c) Amplitude probability density distribution—see Fig. 13.
(d) Peak-to-r.m.s. ratio 10·5 dB.

Figure 14 shows the results obtained using the above noise source to measure q.d. on a codec and results using true noise are also shown. The differences in results occur principally because of the limitations of the peak-to-r.m.s. ratio in the pseudo-random noise signal. The curve of both sources is predictable and measurement accuracies of better than $\pm 0{\cdot}5$ dB are achieved.

Compatibility tests with other sources have been made and a high degree of agreement of results has been

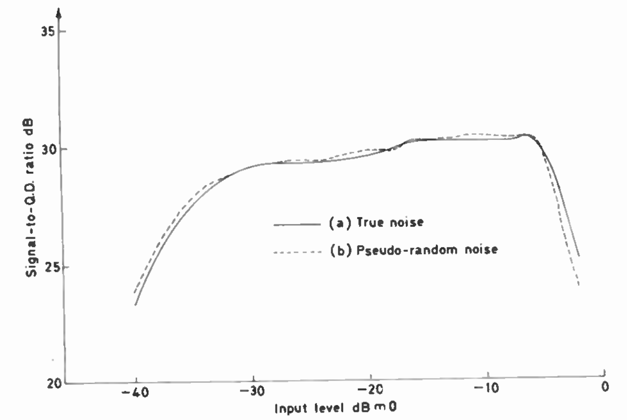


**Fig. 14.** Comparison between measurements using true noise and pseudo-random noise. 24-channel system with *A* law companding (7 bits/sample).

obtained (well within the expected measurement errors). These other sources generally use the same shift register length but clock frequencies and filter bandwidths have significant differences with filter bandwidths from 100 Hz to 200 Hz and clock frequencies giving word lengths of from 150 ms to 400 ms; all, however, have peak-to-r.m.s. ratios of 10·5 dB nominally.

## 8 Sinusoid versus Noise as a Stimulus

Although many telecommunication authorities now use pseudo-random noise for measurements, there is still a significant number who prefer a sinusoid.

The advantages of using a sinusoid are:

(a) Use can be made of test equipment such as distortion measuring sets which already exist for testing analogue systems.
(b) When some types of fault are present, such as missing steps in an encoder, a sharper indication of the effect is given. (See Fig. 15.)

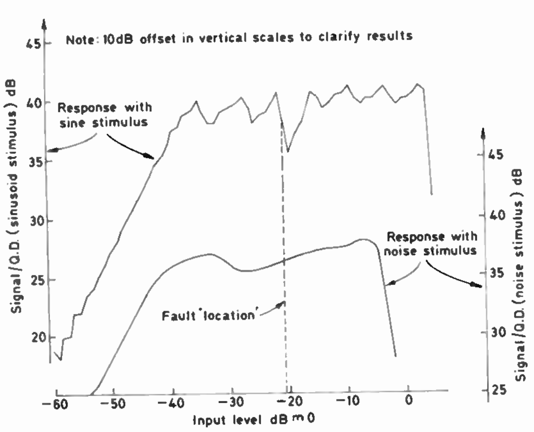


**Fig. 15.** Comparison of indications given by sinusoidal and noise test stimuli when measuring a codec with missing code steps.

The advantages of using a pseudo-random noise signal are:

(a) It more truly represents a speech signal.
(b) The results are more consistent and less open to misinterpretation by test technicians.

Because of the unresolved conflict between the two methods CCITT specify distortion measurements using both types of stimuli.[3] A recommendation[6] has been issued for a test set using pseudo-random noise and the example noise source given in Section 7 meets these requirements. Recommendations for a test set using a sinusoid as a stimulus are under consideration at present.

Figure 16 shows the results of measurements using both techniques applied to the same p.c.m. system, also shown are the CCITT performance masks. It is interesting to note that the performance of the particular

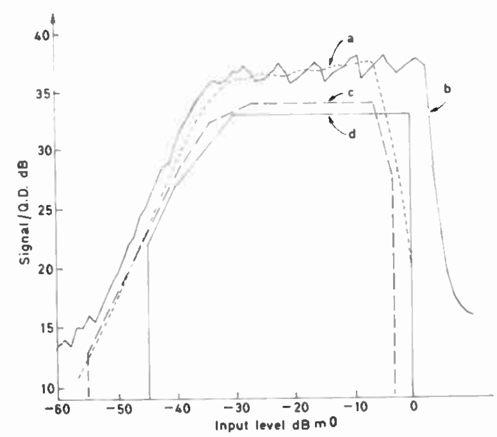


**Fig. 16.** Comparison of a set of results taken on a 30-channel p.c.m. system with *A* law companding (8 bits/sample) using (a) pseudo-random noise stimulus, (b) sinusoid (850 Hz). Also showing the CCITT performance masks for (c) noise stimulus and (d) sinusoidal stimulus.

codec measured was poor and the more stringent test with noise shows it to be out of specification at low input levels. This poor codec performance is not typical, most are nearer to the theoretically best results.

## 9 Conclusions

The two accepted signals for q.d. measurement are a bandpass-limited pseudo-random noise signal and a sinusoid. The choice between them is not resolved internationally although individual telecommunication authorities have decided which method suits their needs.

Most manufacturers of test equipment faced with the two standards for test signals are providing both methods in a single test set, combining these with other facilities to produce equipment capable of complete analysis of the analogue to analogue performance of p.c.m. systems.

## 10 Future

With the advent of all-digital exchanges the necessity to analyse the performance of the analogue to digital section of the system separately from the digital to analogue section is becoming an urgent requirement. Instrument design engineers are now working to produce measurement techniques and the ingenuity being shown in solving the problems should lead to an interesting review paper at a future date.

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*Indexing Terms:* Logic design, Multiplexers, Boolean matrices, Computer-aided design.

# A simple algorithm for logic design using multiplexers

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**SUMMARY**

Many algorithms have been suggested[1-5] in the past to design combinational logic circuits using multiplexers. An alternative method, based on the Boolean matrix approach, is suggested here. The method is simpler and faster compared to existing methods and can be easily programmed on a computer. Finally, the proper selection of multiplexers is suggested by applying simple empirical rules.

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## 1 Introduction

The design of logic circuits using standard gates, or integrated circuits, is well understood. The design algorithms are available in some of the recent books and journal papers related to switching theory.[1-7] Recent advances in m.s.i. technology have, however, produced an awareness of the inadequacy of these design methods. The number of logic gates, and the variety of the logic functions realized on a single m.s.i. chip, for example, multiplexers, demultiplexers, programmable logic arrays, etc., are increasing every year and these chips are now commercially available. Therefore, the design objectives are now different. For example, now one would like to utilize a given m.s.i. chip to its full capacity, or one would like to keep the number of the chips used to realize a given logical function to a minimum. In general, the design algorithms available in literature need to be modified, or sometimes discarded for newer ones, to match the advances in the technology.

In recent years, multiplexers have attracted a lot of attention from the design engineers. This chip is perfectly suitable for the design of the combinational circuits as it can generate all the $2^n$ combinations of an $n$-variable Boolean function. The use of this chip, as a universal logic module, to realize a logic function has given rise to a lot of activity in the area of design algorithms.[1-5] The available design algorithms, in general, aim at two objectives:

(1) The use of a minimum number of multiplexers to realize a given logic function, or alternatively, the use of a multiplexer to its full capacity.
(2) The modification of the available design methods to suit the commercially available multiplexers.

The main problem in the realization of a logic function $f(x_1, x_2, x_i, \ldots, x_n)$, of $n$ Boolean variables, using multiplexers is the proper selection of the 'data input' variable (see Fig. 1). It is obvious that it is convenient to have 'data input' variables as '0' (i.e. ground) or '1' (i.e. chip supply voltage) and an input bus $x_i$ (its complement $\bar{x}_i$ being generated easily). If the input variables are of too many types, e.g. the combinations of $x_1$, $x_2$, $x_3$, etc., then one is required to have three or more input buses. Therefore, the variable chosen as 'data-input' variable is the variable $x_i$ which is present in most of the terms of the function $f(x_1, x_2, \ldots, x_i, \ldots, x_n)$. Our aim in this paper is to present a simple algorithm to find out this variable and derive its values (as $x_i$, $\bar{x}_i$ 0 or 1) which are to be applied as 'data-input' variables. Sections 2 and 3 describe the basis and the algorithm developed respectively. Section 4 presents an empirical rule by which one can determine the number of multiplexers and their types required to realize a given function. The algorithm in Section 3 can be easily modified to suit the types of the multiplexers being used.



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## 2 Basis of the Algorithm

Given a logical function $f(x_1, x_2, \ldots, x_n)$ in the canonical sum-of-products (s.p.) form, it has been shown[6] that it can be put in a Boolean matrix form as follows:

$$\begin{array}{ll} \text{MT:} & [a_0 \quad a_1 \quad a_2 \quad \ldots \quad a_{(2^n-1)}] \\ \mathrm{A_{UNIT}}: & \begin{bmatrix} 0 & 0 & 0 & \ldots & 1 \\ 0 & 0 & 0 & \ldots & 1 \\ \vdots & \vdots & \vdots & \ldots & \vdots \\ \vdots & \vdots & 1 & \ldots & \vdots \\ 0 & 1 & 0 & \ldots & 1 \end{bmatrix} \end{array} \begin{bmatrix} x_1 \\ x_2 \\ \vdots \\ \vdots \\ x_n \end{bmatrix} = [f] \quad (1)$$

Here the minterm matrix (MT) contains coefficients $a_j$, $j = 0, 1, \ldots, 2^n - 1$ which have either value 0 or 1, showing the presence or absence of a minterm of $f$. $\mathrm{A_{UNIT}}$ matrix contains binary coding of the minterms in binary progression from left to right. It is also assumed that $x_1$ has the highest binary weight $2^{n-1}$ and $x_n$, the lowest, i.e. 1.

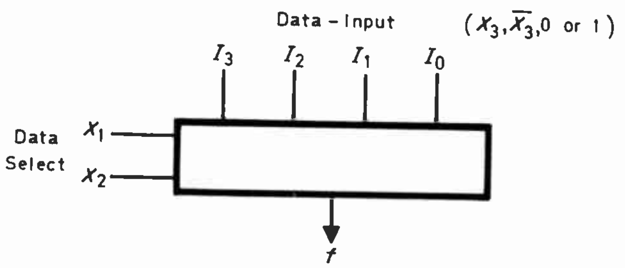


**Fig. 1.**

Now the best choice of a 'data-input' variable (see Fig. 1) will be that variable, out of $x_1, x_2, \ldots, x_n$, which is present in the maximum number of the minterms either in complemented or uncomplemented form. To obtain the 'best choice', we introduce here a function, $PFx_i$, which is called a 'pair-function with respect to variable $x_i$' and is defined as,

$$PF_{x_i} = f(x_1, x_2, \ldots, x_i, \ldots, x_n) \, . \, f(x_1, x_2, \ldots, \bar{x}_i, \ldots, x_n) \quad (2)$$

where '.' and '$^-$' denote the AND and COMPLEMENT operation, respectively. The number of those minterms which provide a simplification, of the type $Ax_i + A\bar{x}_i = A$, is given by the right-hand side of the equation (2). For example, if

$$f(x_1, x_2, x_3) = x_1 + x_2 x_3$$

then

$$\begin{aligned} PFx_2 &= (x_1 + x_2 x_3) \, . \, (x_1 + \bar{x}_2 x_3) \\ &= x_1 + x_1 \bar{x}_2 x_3 + x_1 x_2 x_3 = x_1. \end{aligned} \quad (3)$$

Since $f(x_1, x_2, x_3) = x_1 + x_2 x_3 = \sum 3, \; 4, \; 5, \; 6, \; 7$, therefore equation (3) can also be written as

$$PFx_2 = \sum 4, 5, 6, 7 \quad (4)$$

which precisely contains those minterms which contain $x_2$ or $\bar{x}_2$ so that these minterms, when paired, yield a simplification of the type $Ax_2 + A\bar{x}_2 = A$.

Another interpretation of equation (4) is that it contains only those minterms which differ by the weight of $x_2$, i.e. 2.

The pair function, as defined in equation (2), is difficult to calculate when the number of variables is large. The main aim of this paper is to simplify this calculation.

It is observed from the comparison of the matrix MT and matrix $\mathrm{A_{UNIT}}$, in equation (1), that each variable $x_i$, $i = 1, 2, \ldots, m$ partitions the MT matrix into $2^i$ partitions, denoted here as $\alpha_1, \alpha_2, \ldots, \alpha_p$, where $p = 2^i$. Further, those minterms which are grouped in odd-numbered partitions contain $\bar{x}_i$ in them, while those grouped in even-numbered partitions contain $x_i$. Therefore in a given function $f(x_1, x_2, \ldots, x_i, \ldots, x_n)$, the complementation of $x_i$ results in the rearrangement of the partitions $\alpha_1, \alpha_2, \ldots, \alpha_p$ (where $p = 2^i$) of the MT matrix. The new matrix MT′, which results from MT after $x_i$ is complemented, can be simply obtained by interchanging partition $\alpha_K$ with $\alpha_{K+1}$, where $K = 1, 3, \ldots, (2^i - 1)$.

Thus the pair function matrix $PFx_i$ can be written, in terms of minterms matrix MT, as

$$[PF_{x_i}] = [\mathrm{MT}] \, . \, [\mathrm{MT}'] \quad (5)$$

where matrix MT′ is obtained by the above procedure after complementing $x_i$.

Since both the matrices of equation (5) are row matrices and have 0 or 1 as their elements and since the ANDing is done element-by-element, the computer simulation of the equation becomes simple. Moreover the matrix $[PF_{x_i}]$ will contain only those minterms, indicated by the number of 1's in $[PF_{x_i}]$, which differ by the weight of $x_i$, or alternatively, which can be paired, as described by (4).

Therefore, the best choice for the 'data-input' variable is that variable $x_i$, for which $[PF_{x_i}]$ contains the largest number of 1's.

Once the 'data-input' variable is chosen, then the remaining variables $x_1, x_2, \ldots, x_{i-1}, \ldots, x_n$ are assigned to the data select variables in that order. It then remains to assign values to the 'data-input' variable so that the multiplexer realizes a given function. We now illustrate how this can be done using only the matrix MT.

Suppose that a variable $x_i$ has been chosen as the 'data-input' variable. It was seen above that it breaks the matrix MT into partitions $\alpha_1, \alpha_2, \ldots, \alpha_p$. It was also seen that all the odd-numbered partitions contain those minterms which contain $\bar{x}_i$ and the even-numbered those which contain $x_i$. Therefore the presence (or absence) of $\bar{x}_i$ in a given minterm is indicated by 1 (or 0) in that odd-numbered partition which contains this particular minterm. Similar remarks are valid for $x_i$. In addition, if two different minterms $a_r$ and $a_s$ contain $\bar{x}_i$ and $x_i$

respectively, such that $r-s = w_i$, where $w_i$ is the binary weight of $x_i$, then the 'data input' will have value 1. This is because $x_i+\bar{x}_i = 1$, in this case. These observations directly lead to the method to find the 'data input' assignments as shown below.

Rearrange the matrix MT in two rows as follows:

$$\begin{array}{c} \bar{x}_i \rightarrow \\ x_i \rightarrow \end{array} \begin{bmatrix} \alpha_1, \alpha_3, \ldots, \alpha_q \\ \alpha_2, \alpha_4, \ldots, \alpha_p \end{bmatrix} \qquad (6)$$

where $p = 2^i$ and $q = (2^i-1)$.

Now examine each column of the matrix as shown in (6). If both entries of a column are 0, then 'data input' has a value 0, while if both entries are 1, then its value is 1. Finally, if the entry in the first (second) row is 1 while in second (first) is 0, then its value is $\bar{x}_i$ ($x_i$).

### 3 Summary of Algorithm

*Step* 1

Write the given function in canonical s.p. form.

*Step* 2

Form matrix MT, assigning value 0 or 1 to $a_j$'s depending upon the absence or presence of $j$th minterm.

*Step* 3

Using (5) find $[PF_{x_1}]$, $[PF_{x_2}]$, . . ., $[PF_{x_n}]$. The $[PF_{x_i}]$ which contains maximum number of 1's yields the 'data input' variable $x_i$.

*Step* 4

Using eqn. (6), assign the value of $x_i$ as:

(1) If both entries of a column are 0(1), then $x_i$ has a value 0(1).
(2) $x_i \Rightarrow \bar{x}_i$ if only first row of a column is 1 otherwise $x_i \Rightarrow x_i$, where $\Rightarrow$ means 'is assigned'.

*Step* 5

The remaining variables $x_1, x_2, \ldots, x_{i-1}, x_{i+1}, \ldots, x_n$ are assigned to 'data select' inputs as follows:

(1) If only one multiplexer is used, then the variables are assigned *in that order* [see Fig. 2(a)].
(2) If several 'levels' are used, then start from level 1 onwards assigning $x_1, x_2, \ldots$, etc., *in that order* [see Fig. 2(b)].

### 4 Discussion on Function Realization

Once the design is complete, it then remains to see as to (a) what type of multiplexers and (b) how many multiplexers would be necessary to realize the given function? From a cost point of view, one would like to use a minimum number of multiplexers. Commercially, 4-, 8- and 16-input multiplexers are available and these can directly realize functions of 3, 4 and 5 variables,

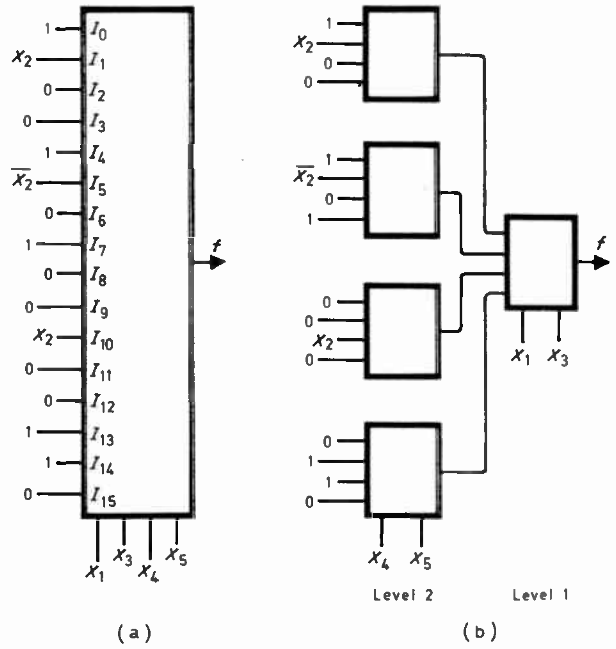


Fig 2. (a)
(b)

respectively. For $n > 5$, the following empirical formula is found to be satisfactory,

$$NMI\text{-}k = \sum_i 2^{[m-(n+1)]-i} \qquad (7)$$

where $NMI\text{-}k$ denotes 'number of multiplexers of $k$-input type',
$k = 4, 8$, or 16,
$m$ is such that $2^m = k$
and $i = 0, m, 2m, 3m, \ldots$, etc.

The following rules are necessary to utilize the relation of (7):

*Rule A*

If for a given $k$, $[n-(m+1)]-i = 1$ for some $i$, then the function cannot be realized using this $k$-input multiplexer.

*Rule B*

If for a given $k$ and $i = 0, m, 2m$, etc., some $i$ is found so that $[n-(m+1)]-i = 0$, then the sum on the right side of the relation gives the total number (minimum) of $k$-input multiplexers required for the realization.

*Rule C*

If for a given $k$ and $i = 0, m, 2m$, etc., some $i$ is found such that $[n-(m+1)]-i < 0$, then the series of equation (7) should be truncated after the first term having a negative power. Let this term denote a fraction $p (p < 1)$. Then the function can be realized using:

(a) $k$-input multiplexers whose total count is given by the sum of the series of positive powers of 2, plus
(b) one $(kp)$-input multiplexer.

*Examples:* The following examples illustrate the use of (7) and the rules.

(1) Let the function to be realized be of eight variables so $n = 8$.

If we use 8-input multiplexers then $m = 3$ and $i = 0, 3, 6, \ldots$, so that from (7) we get

$$NMI\text{–}8 = \sum_i 2^{(8-4)-i} = 2^4 + 2^1.$$

According to Rule A, $n = 8$ cannot be realized using the chosen multiplexer.

(2) Let $n = 7$. Let $k = 4$ so that $m = 2$ and $i = 0, 2, 4, \ldots$. Hence,

$$NMI\text{–}4 = \sum_i 4^{4-i} = 2^4 + 2^2 + 2^0 = 21.$$

As the last power of 2 is 0, according to Rule B, we need 21 multiplexers of 4-input type for realization. Instead of $k = 16$, then $m = 4$, $i = 0, 4, 8$, etc., so that for $n = 7$,

$$NMI = 16 = \sum_i 2^{7-i} = 2^2 + 2^{-2} = 4 + \tfrac{1}{4}.$$

Using Rule C, the series is truncated after first negative power and $p = \frac{1}{4}$.

Hence we need four 16-input and one 4-input multiplexer as $kp = 4$.

## 5 Illustrative Example

Let the function to be realized be

$$f(x_1, x_2, x_3, x_4, x_5) = \sum 0, 4, 5, 7, 8, 9, 12, 15, 21, 22, 26, 29, 30.$$

Step 2:

$$\mathrm{MT} = [10001101110010010000011000100110]$$

$$PFx_1 = [10001101110010010000011000100110] \cdot [00000110001001101000110111001001] = [00000100000000000000010000000000].$$

Similarly,

$$PFx_2 = [10001001100010010000011000000110]$$

$$PFx_3 = [10001000100000000000000000100010]$$

$$PFx_4 = [00000101000000000000000000000000]$$

$$PFx_5 = [00001100110000000000000000000000].$$

As $PFx_2$ has largest number of 1's, therefore $x_2$ is chosen as 'data input' variable.

Step 4. Therefore,

$$\begin{array}{c} \bar{x}_2 \\ x_2 \end{array} \begin{bmatrix} 1000\ 1101\ 0000\ 0110 \\ 1100\ 1001\ 0010\ 0110 \end{bmatrix}$$

Applying Rules A and B we get the assignments of $x_2$ to be applied to the input terminals as,

$$1, x_2, 0, 0, 1, \bar{x}_2, 0, 1, 0, 0, x_2, 0, 0, 1, 1, 0.$$

*Realization.* As $n = 5$ here, the realization using a 16-input multiplexer is shown in Fig. 2(a).

If we have only 4-input multiplexers, then equation (7) tells us that we need 5 multiplexers, 4 for the '2nd level' and 1 for the '1st level'. The realization is shown in Fig. 2(b).

## 6 Conclusion

This paper presents a simple algorithm which can be easily programmed on a computer to realize a given logical function using multiplexers. An empirical rule is also presented which gives an idea as to the number, and types, of the multiplexers required to realize the function and describes the alternatives available to the designer. The rule also yields the minimum number of the multiplexers of the selected variety, required to realize the given function. The algorithm is illustrated by an example.

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*Indexing Terms:* Digital techniques, Storage, Film, Holography

# Some experiments in digital holographic recording

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*Based on a paper presented at the IERE Conference on Video and Data Recording held at Southampton in July 1979*

**SUMMARY**

A holographic recording system, based on the use of low-power laser sources and high-resolution photographic film, is briefly outlined. It was devised principally for the archival storage of high-quality digital television signals and these requirements are reviewed. The development of suitable electro-optic transducers is difficult, largely because of the high data-transfer rates involved, but some progress has been made. Experimental versions of the transducers have been assembled to form a recording apparatus for writing and reading investigations at real-time rates. The results of some initial experiments are given which illustrate various problems that have been encountered, particularly with regard to the recording medium.

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## 1 Introduction

Following the introduction of laser sources, various optical methods were proposed for digital-data recording including the use of holographic techniques. Instead of recording narrow tracks of data in serial form, where each bit of data occupies a small but exclusive area of the recording medium, data is grouped into pages and a hologram of each page is formed.[1,2] By this means, each bit of information is spread over a larger area of the record but not exclusively. It is an attractive concept because the mechanical tolerances required for retrieval can be relaxed considerably and there is better protection against surface dirt and blemishes on the record than in the bit-by-bit approach.

Early proposals for optical computer stores envisaged pages containing $10^4$ bits of data, but, for real-time recording at high transfer rates, the problem was to develop the sophisticated transducers which could compose, record and recover such large blocks of data. For digital television, which involves data transfer rates of about 100 Mbit/s, it was necessary to think initially in terms of compromise arrangements using much smaller holograms containing only a modest number of bits (up to 100, say).

Several encouraging preliminary investigations have led to a possible method of holographic recording, using low-power laser sources and black-and-white photographic film, which is directed towards the archival storage of digital television signals.[3,4,5] Experimental recording and replay apparatus has been partly constructed to test the feasibility of the method.[6,7,8]

## 2 Requirements for Digital Television Recording

It has been assumed that high-quality digital television recording would require a data transfer rate of about 100 Mbit/s. Although bit-saving techniques exploiting the signal redundancies can dramatically reduce the transfer rate required, this can be regarded as a bonus and the spare capacity used to reduce the consumption of film and/or provide better protection against drop-outs. At the full rate, a 50-minute television programme requires a serially accessible store with a capacity of $3\times10^{11}$ bits.

In a system using sequentially recorded holograms, with $M$ bits per hologram, the recording and replay rate required is $(100/M)$ holograms per microsecond.

High information storage density is an important factor in an archival application of digital recording. Using optical media it should be possible to achieve storage densities approaching $10^7$ bits/cm$^2$ with an acceptable error rate on recovery. At this density, a $3\times10^{11}$ bit store would therefore occupy $3\times10^4$ cm$^2$ of recording medium and this could be accommodated, for instance, on 500 metres of 8 mm film.

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## 3 Basic Arrangement for a Holographic System

The system which is being investigated is based on the sequential recording of two-dimensional micro-holograms, each containing 50 bits of information, on high-resolution (holographic) film. Either amplitude or phase types of hologram can be used, but the prime interest is in amplitude holograms because it is believed that these offer better prospects for archival storage and that film processing is less critical.

### 3.1 Format

At the present stage of development, the holograms are approximately $50 \times 50\ \mu m$ and are arranged in rows across the film as indicated in Fig. 1(c). In a complete recorder it is envisaged that there would be 128 holograms per row; this is roughly equivalent to storing one television line per row. The complete record forms a track 6·4 mm wide which could be accommodated on unperforated 8 mm film, or as two tracks on 16 mm film.

If a 10 μm guard band between rows is used, the film will be required to run continuously at 94 cm/s in order to record a 100 Mbit/s data stream in real-time. The storage density is approximately $2 \times 10^6$ bits/cm$^2$ with these hologram dimensions.

### 3.2 Recording Process

Figure 1(a) shows the basic recording arrangement. The principal components are a low-power (10 mW) laser, a hologram page-composer, a single-axis deflector and a film transport mechanism. The laser beam is split by a partially reflecting mirror to derive a reference-beam and a signal-beam. The latter is expanded and illuminates a multi-port page-composer which generates and independently modulates a 'page' of digit-beams. In the present system there are 50 ports, each of which is connected to an electronic switch controlled by the incoming data. The light emerging from the page-composer consists of a two-dimensional array of narrow, collimated digit-beams. At this point the reference-beam is recombined and forms a central, collinear beam in the array. The array of digit-beams and the reference-beam pass through a deflector before they are brought finally to a common focus on the recording film by a converging lens.

When one hologram has been recorded, a new page is composed while the deflector redirects the beams slightly, so as to record the hologram of the new page at the next position in the row. This process continues until a complete row across the film has been exposed, when the deflector returns to the start of the scan.

### 3.3 Replay Process

After the film has been developed, it can be replayed using the arrangement shown in Fig. 1(b). If a laser beam is focused on a recorded hologram, an image of the original page of data is reconstructed on the far side of the film. This image, which is not localized but a diverging pattern of digit-beams, is intercepted by an array of photodetectors. These are geometrically disposed so as to correspond to the known pattern of digit-beams in the page. By this means, the presence or absence of a digit-beam in the page is detected.

The incident laser beam is caused to scan over a row of holograms by a single-axis deflector, so as to interrogate each hologram in turn; an acousto-optic deflector with its scan direction across the film is suitable for this purpose.

Ideally, the binary signal output from the photo-dectectors, after amplification and processing, is a

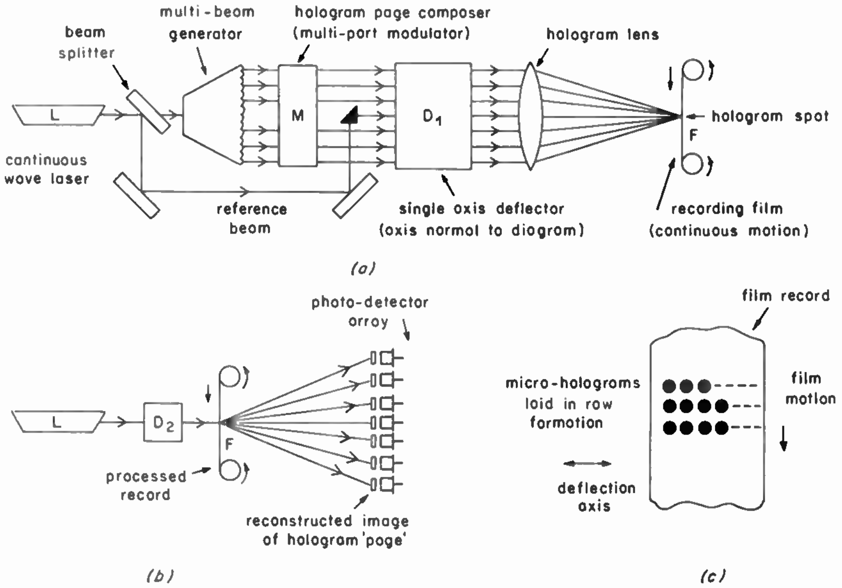


Fig. 1. Basic system arrangement: (a) Recording process (b) Replay process (c) Recording hologram format.

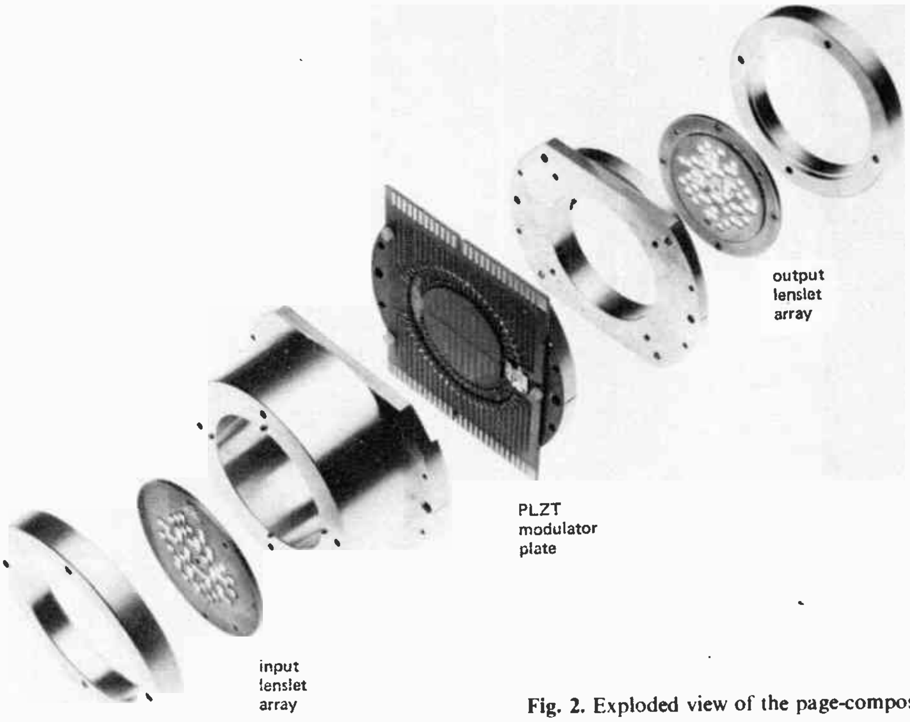


Fig. 2. Exploded view of the page-composer.

faithful replica of the digital signals applied to the electronic switches driving the ports of the page-composer in the recording process.

The film moves continuously as in recording, but means are required for controlling the film speed and ensuring that the scanning spot moves approximately along the centres of the holograms in each row.

## 4 Transducer Development

The average scanning rate required for the recording and replay deflectors is 2 holograms/μs, using 50 bits per hologram. Thus the maximum time available for exposing each hologram in recording, and for interrogation on replay, is 500 ns. Moreover, a new page of data must be composed in a small fraction of this time if significant cross-talk between consecutive holograms is to be avoided. Switching times for the ports in the page-composer of about 100 ns are required.

### 4.1 The Page-composer

After considering several possible methods for constructing a suitable page-composer, one based on the quadratic electro-optic effect in lanthanum-doped, polycrystalline lead zirconate-titanate (PLZT) was believed to offer the best solution. A linearly polarized beam of light passing through a thin polished slice of PLZT has its plane of polarization rotated through 90° by a transverse electric field at 45° to the direction of polarization. The field-strength required can be obtained by applying a voltage, usually between 200 and 250 V, across a pair of conducting electrodes on the surface of the slice. The electrodes are separated by a small gap or port (~ 100 μm) through which the incident beam is directed.[9] By depositing a number of electrode gaps on the PLZT slice, with provision for connections to external drive units, various page arrangements can be produced.

In operation, it is normally arranged that with no voltages applied to the ports the digit-beam polarizations are at 90° to that of the reference-beam. Under these conditions the digit-beams are, in effect, switched 'off' and they would not appear in the subsequent reconstruction of the hologram. However, when a voltage is applied to a port the polarization direction is rotated to coincide with that of the reference-beam and that digit-beam is turned 'on'.

The form of construction adopted is shown in Fig. 2, which is an exploded view of an experimental 50-port version. It comprises an input array of 50 small lenses (lenslets) for focusing the incident light through the ports, a PLZT modulator plate as described above, and a corresponding output array of lenslets for recollimating the digit-beams. The optical axes of the output lenslets are arranged to be coincident with those of the input set, respectively, and the lenslet arrays are axially separated by the sum of their focal lengths to produce an afocal system. Thus each pair of lenslets (one input and one output) generates and defines the position of one digit-beam in the page.

The geometrical arrangement of the page used in the experimental version is shown in Fig. 3, where the crosses indicate the centres of the digit-beams. The

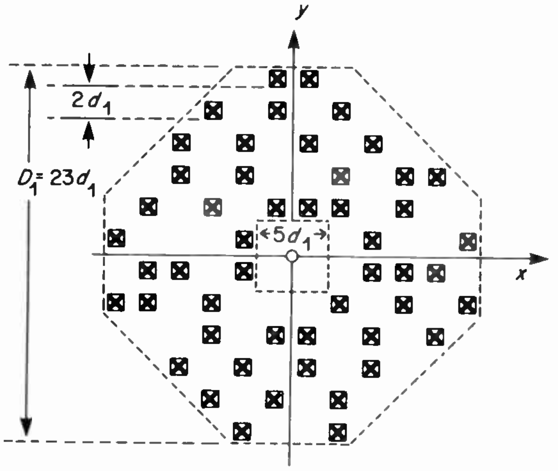


Fig. 3. Page arrangement of digit-beams.

reference-beam, when introduced just downstream of the page-composer, occupies a central position in the array of digit-beams (i.e. at the origin (0,0) in Fig. 3). The minimum spacing of the digit-beams is $2d_1$, where $d_1$ is the cross-sectional dimension of the beams ($d_1 \sim 1{\cdot}5$ mm). The digit-beams are irregularly positioned on a grid of possible locations, which is offset from the origin, but with the constraint that there are no pairs of beams related by point symmetry through the origin. This arrangement leads to a low-bandwidth (on-axis) type of hologram with reduced variance of the light-intensity distribution at the recording plane and with spatially-separable intermodulation products in the reconstruction to avoid spurious inter-beam cross-talk effects.

One of the constructional difficulties with this form of page-composer is the precise alignment of the lenslets in the output array and special techniques had to be devised for individually positioning and fixing them.[7]

Electrical inter-channel cross-talk due to capacitive coupling was eliminated by paying special attention to the electronic switch circuits and the lead-in arrangements to the ports.

Each port is required to handle a 2 Mbit/s data stream and, at this switching rate, there is a problem due to the dielectric heating effects. The sensitivity of the electro-optic effect decreases with temperature and can be only partly compensated for by increasing the switching voltage. Fortunately, the transition times and the stability of the zero-voltage state are unaffected. Cementing a cover glass to the electroded surface of the PLZT slice was found to assist the heat dissipation at the ports and lower the temperature. However, with many of the available optical cements, electrical breakdown or deterioration of the cement layer occurred causing premature failure of the ports.

### 4.2 Recording Deflector

There are various possibilities for constructing a suitable deflector, but mechanical solutions based on rotating mirrors[10] etc. were not considered. A more elegant approach is afforded, in principle, by a solid-state digital light-deflector based on electro-optic crystals and Wollaston prisms.[11] Earlier studies had indicated, however, that there would be difficult design problems to overcome if this type of deflector is required to operate at high scanning rates. On the other hand, the method is

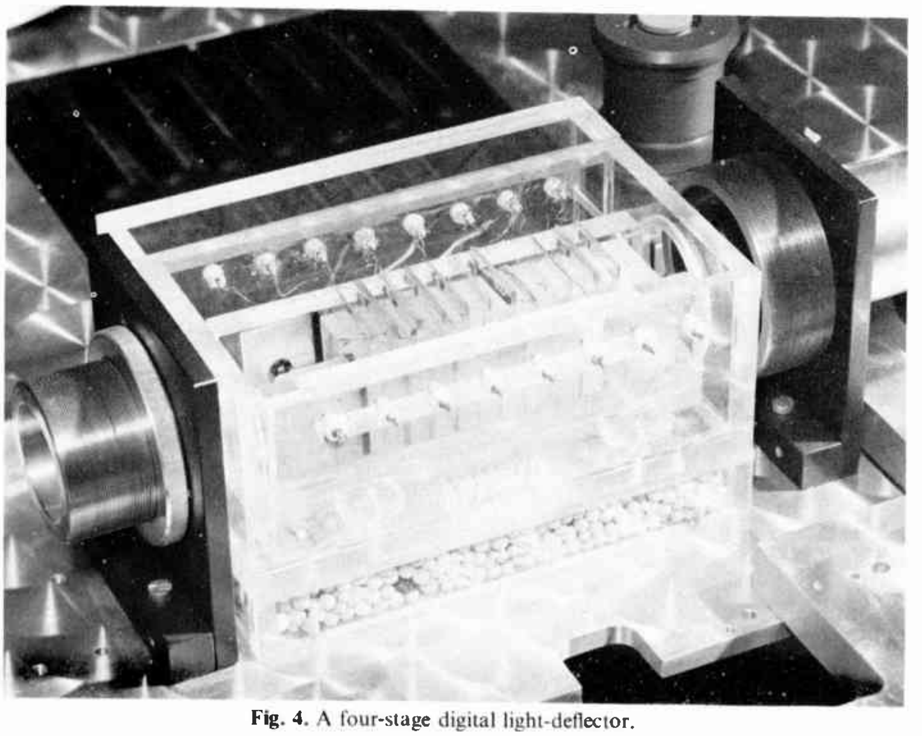

Fig. 4. A four-stage digital light-deflector.

attractive for a number of reasons including the fact that there is no relative motion in the direction of scan between the hologram exposing light and the recording film. This avoids the need to 'freeze' the exposure of each hologram during recording by pulse modulating the laser source.

A digital light-deflector comprises a number of deflection stages in tandem, each stage consisting of a Wollaston prism preceded by an electro-optic polarization switch. The prism deviates the incident digit-beams through a fixed angle, $\pm\beta$, the sign depending on the direction of polarization. This direction can be rotated through 90° by applying a suitable voltage (the half-wave voltage) to the electro-optic switch. If the deviation angle of each stage is made precisely twice that of the preceding stage, a deflector with $2N$ angular output positions can be constructed, where $N$ is the number of stages. Regular scanning is achieved by synchronously driving each stage with a square-wave voltage whose period is proportional to its angle of deviation, $\beta$.

An experimental four-stage deflector was constructed (Fig. 4) using deuterated dihydrogen phosphate crystals (KD*P) for the polarization switches and quartz for the Wollaston prisms. Except for the first stage, the switches have a large aperture (20 × 20 mm) to accommodate the complete array of digit-beams and are of the longitudinal type with the applied electric field in the same direction as the beams. Basic cells were fabricated by cementing electroded glass plates to both surfaces of a KD*P crystal. The thin-film electrodes took the form of ladder grids, through which the digit-beams pass unobstructed, with the gaps parallel to the deflection plane. Each polarization switch contains two basic cells in tandem; this reduces the applied voltage required for switching to about 2 kV.[7]

The electro-optic crystals are connected to 2 kV electronic switch units[12] which are controlled and synchronized using clock-generated square-waves. For a complete recorder, a seven-stage deflector would be required (128 output positions), each stage driven by a high-voltage square-wave; the fundamental frequencies would then range from 1 MHz, for the fastest stage, to 16 kHz for the slowest.

Ideally, the 2 kV electronic switches should have rise and fall transition times in the region of 100 ns. This has not been achieved, due to the limitations of existing high-voltage transistors, and the best transition times obtained so far have been between 200 and 250 ns, depending on the switching rate and the capacitive load. Also, there is a heat dissipation problem in the transistors at the highest switching rate.

Figure 5 shows some of the optical switching waveforms obtained with a KD*P polarization switch when driven with a square-wave voltage at various fundamental frequencies. The optical switching characteristics are good in the faster stages, but the effects of bulk mechanical resonances in the KD*P crystals due to piezo-electric coupling begin to be troublesome below about 62 kHz.

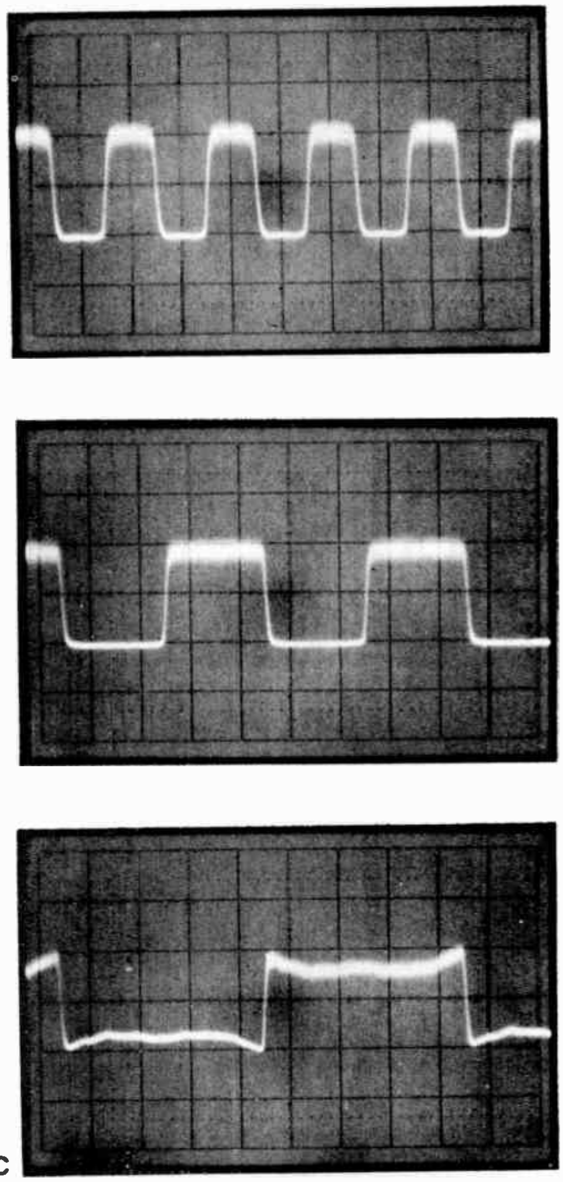
a
b
c

Fig. 5. Optical switching waveforms obtained with a KD*P polarization switch for three frequencies of applied square-wave voltage: (a) 250 kHz (b) 125 kHz (c) 62 kHz.

### 4.3 Photodetector Array

Although an integrated, two-dimensional, photodiode array would provide the most compact arrangement for detecting the reconstructed hologram images, more flexibility is obtained at the development stage by using discrete photodiodes.

An optical probe was constructed consisting of a bundle of flexible light-guides. One end of each guide is optically coupled to a small-area silicon photodiode. The free ends can then be arranged to correspond to the

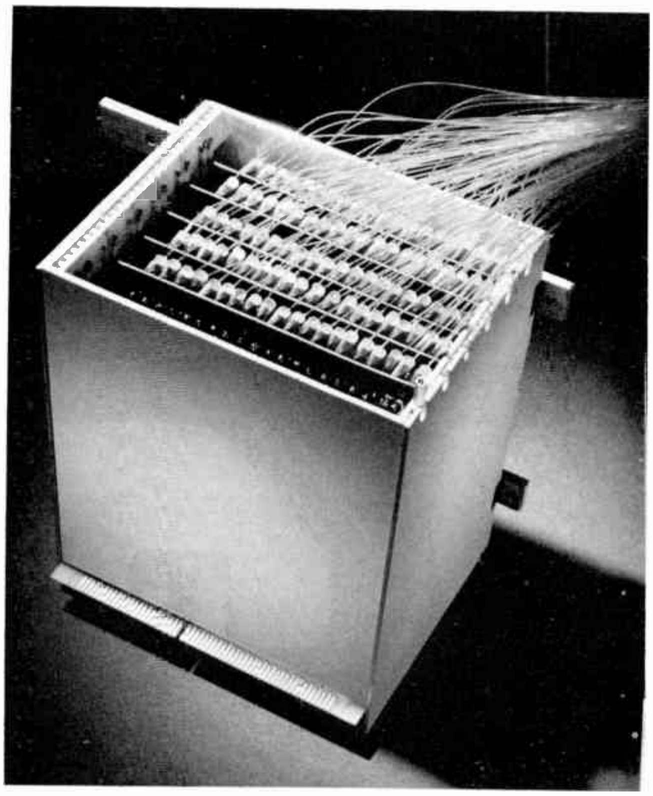

Fig. 6. Photodetector array/amplifier unit with fibre-optic link.

known page configuration at the detection plane by inserting them in a suitably drilled block. The photodiodes are electrically coupled to low-noise f.e.t. input amplifiers, followed by signal processing circuitry to optimize signal-to-noise performance. Figure 6 shows an experimental photodiode/amplifier assembly containing 100 photodiodes, together with the bundle of light-guides which links the array to the detection probe.

The overall sensitivity of each detection channel is typically 200 mV/μW. A signal-to-noise ratio of 26 dB was measured in a 1 MHz bandwidth for a digit-beam power of 40 nW incident on the probe. This performance is considered to be adequate for a replay system using a 5 mW helium-neon laser.

## 5 Some Initial Results

Pending completion of the experimental recorder, it has been possible to carry out some initial performance tests and produce various trial recordings, in order to make a realistic assessment of those areas where further development is required.

Fig. 7. A typical track of recorded holograms (holograms 60μm apart).

### 5.1 Recording

Photometric measurements at the recording plane showed that the digit-beams could be switched 'on' or 'off' in 100 ns with on/off contrast ratios greater than 10 : 1.

After assembling the fastest deflection stage, tracks containing one hologram per row were recorded on moving holographic film with an exposure time of 500 ns per hologram. Figure 7 is a microphotograph of a typical experimental track of holograms recorded in this manner; the holograms are approximately 60 μm apart.

Using a helium-cadmium laser operating at a wavelength of 441 nm the measured overall transmission of the recording apparatus was 5%; this is substantially lower than was anticipated.

It was found, by experiment, that the best recordings were obtained when the reference-beam to signal-beam intensity ratio at the recording plane was about 6 : 1.

### 5.2 Reconstruction

Test strips from the trial recordings were either placed in the replay head and selected groups of holograms repeatedly scanned at the full rate (2 holograms/μs), or the reconstructed images of individual holograms could be measured statically.

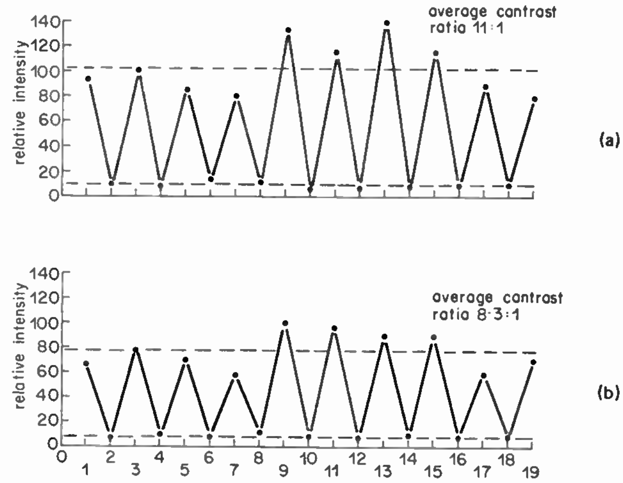


Fig. 8. On/off contrast ratio measurements: (a) Optimum size of reconstruction spot (b) Spot-size increased by 20%.

In one experiment, a track was recorded dynamically with the page of digit-beams switched off on alternate holograms. A typical result is shown in Fig. 8, where the relative digit-beam powers at a given location in the page is plotted for a series of consecutive holograms in the track. In Fig. 8(a), the size of the incident reconstructing spot, which has a Gaussian radial intensity profile, has been adjusted to give the most efficient reconstruction. In Fig. 8(b) the same group of holograms was measured, but the spot size was increased by 20%. A surprising feature of this and other similar experiments was the large variation (up to 2 : 1) in the peak intensities, which correspond to the 'on' holograms, bearing in mind that their exposures were nominally identical during recording.

5.3 Hologram Quality

In addition to the unpredictable variations in the reconstructed digit-beam powers from hologram to hologram, there was a large and irregular variation within the page. The measured digit-beam powers near to the centre of the page were commensurate with theoretical estimates of the holographic efficiency of the process, but there was an irregular fall-off in efficiency towards the edges greater than anticipated.

One possible reason for these unpredictable variations is that the holograms are not pure amplitude holograms, as intended, but have an unwanted phase component introduced by dimensional changes in the gelatin when the film is processed. It has been shown, for example that there is a surface relief pattern covering the hologram. This uncontrolled phase component perturbs the wavefront phase uniformity sufficiently to cause distortion of the diffraction process controlling reconstruction.

## 6 Conclusions

The project has not advanced sufficiently to be able to assess the overall feasibility of the proposed method. However, some progress has been made in the development of suitable transducers capable of handling the high information transfer rates involved in digital television recording.

Further work is required to overcome some of the practical difficulties encountered. In particular, problems associated with the recording film and its processing need elucidating.

Meanwhile, there has been further progress in other digital recording techniques and holography must be evaluated against the current status of these other possible methods of archival storage.

## 7 Acknowledgments

I wish to thank the Director of Engineering of the BBC for permission to publish this paper and also those of my colleagues who have assisted in this project.

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## The Author

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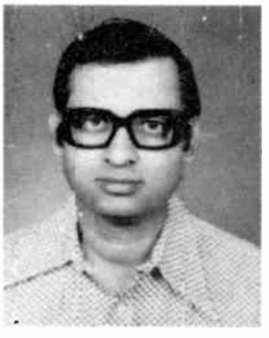

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**Michael Carey** was awarded a Post Office Scholarship to Imperial College, London in 1967, receiving the B.Sc. degree in 1971. For the next five years he worked at the Post Office Research Centre, Martlesham Heath, on the computer control of small telephone exchanges and later on digital encoding and filtering for speech transmission in digital telephone exchanges. He was awarded the Ph.D. degree from Imperial College in 1978. Since 1976 he has been Lecturer in Electronics at the University of Keele where he is leader of a group investigating the applications of digital signal processing. Dr Carey has acted as consultant to commercial organizations on gramophone noise suppression, digital video techniques and filter design. Current research interests include digital-to-analogue conversion techniques for high quality applications, speech signal processing and digital filter design on which he has a number of publications.