

Founded 1925

Incorporated
by Royal Charter 1961*To promote the advancement
of radio, electronics and kindred
subjects by the exchange of
information in these branches
of engineering*

The Radio and Electronic Engineer

The Journal of the Institution of Electronic and Radio Engineers

WARC-1979*

M. MILI

ON 24th September 1979, the World Administrative Radio Conference (WARC) opened in Geneva, a conference which in ten weeks is to revise, bring up to date and harmonize the Radio Regulations drawn up by a similar conference in 1959. With its 1800 participants, its 14 000 proposals amounting to nearly 5000 pages and co-ordinated documentation in several volumes totalling nearly 3000 pages, no terms seem to be adequate to designate the largest conference ever organized under the auspices of the ITU.

It is no exaggeration to assert that this Conference will establish the framework that will regulate radiocommunications up to the year 2000. Moreover, since its task is to seek an equitable balance between everyone's most immediate needs and the long-term requirements of all services which will become operational during the next two decades, it will influence the planning and utilization of services well beyond the year 2000.

In our times, radio still seems, as it did in the beginning, to be one of the happiest and most productive discoveries of all ages. As we approach the year 2000 and an era when telecommunications, closely linked with computer science, will occupy a preponderant place in society, it is brought home to us that, from the dawn of the 20th century, radiocommunications have constituted a decisive factor in the prodigious development of the world. Have they not from the outset put an end to the isolation of seamen and of the most distant regions and, through broadcasting, have they not proved to be the most powerful of the mass communication media?

Unfortunately, we are all aware that the marvellous opportunities they offer us are dependent upon a frequency spectrum which is still limited, although its upper limit is constantly expanding as the result of technological advances.

It was very soon observed not only that radio waves knew no frontiers, but also that they were propagated over very long distances and could thus cause harmful interference. Radiocommunications therefore represent an area of activity in which the major problems could only be dealt with at the world level. There are, of course, specific requirements, regional and even national, and their disparity has led to the division of the world into three regions for the purposes of the Radio Regulations and particularly to a proliferation of 'footnotes' in the Table of Frequency Allocations.

These footnotes result in a less satisfactory use of the spectrum and introduce additional complications for frequency planning and co-ordination and for the standardization of equipment. It is therefore to be hoped that WARC-79 will make every effort to get rid of as many of them as possible.

Thus the Radio Regulations have not been revised in their entirety for twenty years. The mere fact that the Table of Frequency Allocations to the various services has not been reviewed as a whole since 1959 emphasizes the importance attached to this Conference.

A great deal has happened in the past twenty years. First of all, technology has made giant strides: in 1959 there was hardly any mention of large-scale integrated circuits and satellites were mainly for research. Moreover, for lack of suitable equipment, frequencies above a few gigahertz were scarcely used, and optical waves even less.

During those twenty years some far-reaching political events have also occurred. In 1959, the Union had only 96 Members and 5 Associate Members, some of them not yet fully independent. Today the membership is 154, making an increase of over 60%. The Union is thus faced with new obligations, especially since telecommunications are increasingly becoming the cornerstone of a country's infrastructure and consequently

* This guest editorial is based on extracts from the address to the inaugural meeting of the World Administrative Radio Conference by Mr M. Mili, Secretary General of the International Telecommunication Union. It is reproduced from the ITU's *Telecommunication Journal* for October 1979. It is hoped to publish authoritative technical reports on the Conference decisions in *The Radio and Electronic Engineer* early in 1980.

determine the success of national development as a whole. In many countries, radiocommunications of all kinds can be brought into operation more rapidly than any other methods of transmission and can be maintained more easily.

Referring to the International Telecommunication Convention, there are two basic provisions (in Article 33) which have to be borne in mind:

'Members shall endeavour to limit the number of frequencies and the spectrum space used to the minimum essential to provide in a satisfactory manner the necessary services. To that end they shall endeavour to apply the latest technical advances as soon as possible.'

'In using frequency bands for space radio services Members shall bear in mind that radio frequencies and the geostationary satellite orbit are limited natural resources, that they must be used efficiently and economically so that countries or groups of countries may have equitable access to both in conformity with the provisions of the Radio Regulations according to their needs and the technical facilities at their disposal.'

Particular attention will naturally be paid to such services as the mobile and broadcasting services, for which there is no economically viable solution other than the use of one or more frequency bands. But it is also important to take longer-term interests into consideration, by ensuring the allocation of adequate frequency bands to the various services and by adopting provisions whereby countries which are not in a position to do so today may, in the more or less distant future and at a time they consider appropriate, introduce new services to meet their national or international requirements. In such cases, they should be able to do so without giving rise to congestion or causing interference for the users of other services.

The agenda of the Conference emphasizes the importance of technical studies in the preparations for administrative conferences. It is thanks to these studies that radiocommunications have been able to expand their potential in quality and quantity since the beginning of the century. This perpetual race between demands on the spectrum and its capacity calls for increasingly elaborate technical measures. For many years, progress was made only with regard to a few specific technical points, such as propagation and modulation, and to equipment, particularly antennae. But now all the characteristics have to be considered together with a view to reaching the optimal solution, that is to say, maximum utilization of every band in the spectrum with a view to meeting the requirements and protecting the interests of all users.

'The Electronics Engineer'—A New Publication for All Members

During the last two years the Institution has been looking into the best way of providing members with a more up-to-date and complete service of news on matters both technical and non-technical than is practical with a monthly Journal. Members will recall that following a free mailing of ten issues of *Electronics Weekly* during the autumn of 1978, they were offered a concessionary subscription at half the normal price and several hundred members did in fact take this up. The number involved was not however sufficiently large to justify the Institution and the publishers taking the idea a further step forward by arranging for *Electronics Weekly* to carry Institution announcements and news on the complete and regular basis felt to be necessary.

The news content of *The Radio and Electronic Engineer* has of course been increased over recent months but the prerequisite of more frequent publication clearly cannot be overcome. Arrangements have therefore been made for members in Great Britain and throughout the world to receive a fortnightly newspaper—'*The Electronics Engineer*'—without any charge from January 1980. A newly-formed commercial publishing organization, Electronic Engineering Publications Ltd, will produce a tabloid publication of initially 8 pages containing Institution news and announcements, details of meetings, conferences and other activities, news of the industry and appointments advertisements.

The Society of Electronic and Radio Technicians is also participating in this venture and *The Electronics Engineer* will be sent on the same basis to its members. By this means 'vertical integration' of the electronics engineering profession as represented by its two leading specialized professional bodies will be substantially reinforced.

The Electronics Engineer will thus be incorporating the 'Forthcoming Meetings Bulletin' which has been sent out to members in the British Isles since September 1979. Twice-monthly publication can however provide much more

immediate notice of meetings and, most important, of any additions or alterations to the programmes. It is believed that members will particularly appreciate this, not least the local section committees who organize programmes of meetings each year and we are sure that they will be keen to co-operate in taking advantage of the swift response time of the new medium.

Publication dates for *The Electronics Engineer* will be the 1st and 3rd Thursdays of the month and the first issue will be sent direct to all members of IERE and SERT on 24th January 1980. The issues appearing on the 1st Thursday of each month will usually be sent out under the same cover as *The Radio and Electronic Engineer* and *Electronics Technology*. When an issue covers two months the newspaper will be sent out to members by itself in the month in which the main Journal does not appear.* For obvious reasons there will *not* be an issue on the 3rd Thursday in December!

In announcing this exciting new venture, the Council of the IERE believes that it will go a long way towards meeting the well understood demand of the electronic and radio engineer for up-to-date news and views of his profession, the need for which has been widely expressed both at meetings and in letters. It is hoped that members will, for their part, make the fullest use of *The Electronics Engineer* and support it as an advertising medium, for that is what its success will depend upon. Members who need to engage electronic and radio engineering staff, whether Chartered Engineers, Technician Engineers or Technicians, are urged to use it for this purpose. Similarly, members who are seeking new appointments are asked to give particular attention to posts advertised in their own newspaper.

* Combined issues of *The Radio and Electronic Engineer* in 1980 will be January/February and November/December. Thereafter, until further notice, the combined issues will be July/August and November/December.

The 54th Annual General Meeting of the IERE

*Held at the London School of Hygiene and Tropical Medicine
on Thursday, 25th October 1979*

The meeting was opened at 6 p.m. by the President, Mr D. W. Heightman, when 68 members were present.

It was confirmed by the Secretary, Air Vice-Marshal S. M. Davidson, that due notice of the meeting, the eighteenth Annual General Meeting of the Institution since its incorporation by Royal Charter, had been sent to all members in the July/August 1979 issue of *The Radio and Electronic Engineer*.

The Secretary then reported that the minutes of the 53rd Annual General Meeting, held on 5th October 1978, had been published in the December 1978 issue of *The Radio and Electronic Engineer* (Volume No. 48, no. 12, pp. 631–634). No comment having been received on these minutes, it was unanimously agreed that they be taken as read and the President thereupon signed them as a correct record.

Annual Report of the Council

The President then presented the 53rd Annual Report of the Council for the year ended 31st March 1979 and said:

'The Annual Report of the Council of the Institution for the year ended 31st March 1979 was published in the September 1979 issue of our Journal. Once again the Council thought it proper to report in writing to the membership in considerable depth on all aspects of the Institution's affairs, so I need not go into great detail at this stage.

'I feel, however, that I would like to highlight several of the more important points covered in the Report before moving its adoption.

'First, overseas affairs. In the Executive Committee section of the Report you will have seen that the detailed review of arrangements for the management and support of the IERE membership in India announced in the previous Annual Report was duly completed and in the light of this Council decided to cease publication of the Indian Proceedings in July 1978 and to close the Institution's Bangalore Office in September 1978. The Institution's Indian affairs are now administered direct from 99 Gower Street, but there has been no change in the arrangements for learned society activities in the Indian regions, for which Dr P. K. Patwardhan (the last Indian Divisional Chairman) has willingly accepted responsibility as IERE National Representative in India.

'Whilst some of us must feel more than a tinge of regret at the run-down of our affairs in India we have to admit (and our commitment to the Commonwealth Engineering Council reflects this) that it would be churlish of us not to welcome the fact that our Indian colleagues are now able to turn to their own professional bodies for the support for which they have looked to us in the past. Fortunately we can take further comfort from the fact that as our representation and influence in India has declined so it has been growing in other areas of what we call the Far East. As the Report shows, we now have an enthusiastic and thriving local section in Hong Kong and our membership is steadily growing larger and stronger in both Singapore and Malaysia—all of these being fast developing areas of interest to the world-wide electronic engineering community.

'Second, concerning personalities, many of our senior



Do we detect a certain look of satisfaction on the Honorary Treasurer's face? Mr S. R. Wilkins waits for any comments from members on the sixth set of Institution Accounts he has presented since he was first elected to this post in 1973.

members will have learned with deep regret of the death in Canada on 13th January 1978 of Leslie H. Paddle, a Founder Member. It was typical of Leslie Paddle's generosity and his unfailing interest in our profession that he should include in his will a bequest of 50,000 Canadian dollars to the Institution to found a post-graduate scholarship in electronic or radio engineering. An educational trust has been set up and it is intended to invite applications for the award of the first scholarship towards the end of the current financial year.

'And third, I must refer with particular gratitude to the time spent by the Education and Training Committee and the Executive Committee in considering the implications of proposals made by other bodies for changes in professional standards and registration procedures. In particular there was the Institution of Electrical Engineers' "Merriman Report" as a result of which that body decided to raise, with effect from 1982, the academic standard for entry to C.Eng. status above that agreed and specified by CEI. In stating that no such change was at present needed or justified the IERE Council unanimously endorsed the conclusions of its committees, and this view was publicized in *The Radio and Electronic Engineer*. The response to this statement of policy on such a vital matter in the main strongly supported the IERE view, this support being particularly evident among those active in higher management on the British industrial scene.

'Next, I would like to draw your attention to the wide range of our professional activities referred to in the three sections of the Report headed Professional Activities Committee, Papers Committee and Institution Publications. As you read these sections I hope you will note with approval the progress we have made in the course of the year under report on the development of our Journal to widen its appeal without damaging its highly professional image, and the continuing strength of our conference work and local section meetings programmes. All of this is of great importance for the very good reason that it is in this area of our work that we see the prime purpose of the Institution, namely the advancement of the art, science and practice of electronic and radio engineering, being achieved in a most satisfactory manner.



The President hands to Mr Ahmed El-Badry his part of the Clerk Maxwell Premium. His co-author, Professor John Simmons, was unfortunately prevented at the last moment from coming over from the United States.

'And finally, I feel that I must draw your attention to the Table in the Membership Committee report, because I think it is both remarkable and encouraging that in a year of such great uncertainty concerning the Finiston Report and the future of the profession, and of such severe financial pressure on all personal and company accounts that we should have seen such stability—and even a modest measure of growth—in our membership. For myself I find it particularly heartening to see that growth strengthening the vertical integration of our profession in the manner which I discussed in my Presidential Address just a year ago, highlighting as it does the close association and harmony which must exist between the technician engineers and the chartered engineers in our profession, if we are to achieve maximum development and exploitation of the promise of electronics in the public interest.

'In conclusion, I suggest that the report on our 53rd year demonstrates the domestic stability, professional confidence and steady progress in pursuit of our aims and objects that is appropriate for our Institution in the light of our Royal Charter obligations. But before I move the adoption of the Report I would both invite and welcome comment.'

No questions were asked and the President moved from the Chair that the Annual Report of the Institution for the year ended 31st March 1979 be adopted and this was approved unanimously.



Dr Gunter Kompa was able to come over from Germany to receive his Heinrich Hertz Premium from Mr Heightman personally. The Institution's Editor, Mr Frank Sharp, who introduced the prize winners, is in the background.

Auditors' Report, Accounts and Balance Sheet

The President called upon the Honorary Treasurer, Mr S. R. Wilkins, to propose the adoption of the Institution's Accounts for the year ended 31st March 1979, who said:

'Examination of the income and expenditure account will show that the Institution ended the year with a deficit of income over expenditure of £8,165.

'When this figure is considered in relation to non-recurrent items of expenditure resulting from the preparation, printing, publishing and implementation of the recent By-law amendments, the costs arising from the delayed closure of the India office, which totalled over £8,000, and the loss of revenue previously enjoyed as a contribution from the National Electronics Council for publishing their Journal, then I think we can regard the year's results as satisfactory.

'Income for the year shows an increase of a little less than £8,000 largely resulting from an improvement in publication sales.

'Despite the inevitable increase in printing costs amounting to some £14,000, the overall expenses of the Institution have increased by under £2,000 or less than 1% of the total.



The Leslie McMichael Premium for a paper on management techniques is handed to Mr Eric Barnard, a chartered patent agent.

'Of particular note in this context is the cost of administration which accounts for nearly 50% of the total Institution expenses and which, despite the effects of inflation has been kept within a few hundred pounds of last year's figure. This highlights the tight management control to which I referred last year and reflects much credit on the Secretary and all the Institution staff whose efforts have contributed to this result.

'With the continued inflationary increase in costs that seem inevitable during the current year it would be naive to expect that there will not be a marked increase in the Institution's running expenses if the high standard of services to members is to be maintained.

'However, we have an excellent conference programme which should produce increased income from fees and publications, and of course there will be no recurrence of the exceptional items of expenditure to which I have already referred. Added to this is the additional income arising from the new subscription rates which take effect this year.

'With a continuance of the "good housekeeping" policies which have been established, we can confidently look forward to a year which will show a surplus of income over expenditure leading to a significant reduction of the adverse balance, with the consequential reduction in bank charges and improved financial credibility.



In presenting the Brabazon Premium to Professor D. E. N. Davies, Mr Heightman noted that Professor Davies's first paper to the Institution (in 1958) had also gained him the Brabazon Premium and he hoped that the present paper would not be the last we received!

'This brings me to the last point I wish to make in this report. Members will have noted from the accounts a significant drop in the figure for bank charges—a direct result of tight financial control and improved cash flow performance.



Mr Heightman is weighed down with the books which he had to present to Dr Nicholas Maslin for the Rediffusion Television Premium.

'They will therefore be somewhat surprised, as I was, to see that the overdraft figure quoted in the balance sheet is some £12,000 higher than last year. I am pleased to be able to tell you that this is somewhat of a "freak" figure, occurring only during the last few days of the financial year and in no way reflecting the operating level of overdraft throughout the year.

'It arose because such a large number of our members responded to our suggestion to pay their membership fees by variable direct debit—for which incidentally we are most grateful—that we were denied the income from fees paid in advance with which to offset the inevitable spate of end of year creditors. This situation was of very short duration however and within a few days of the new financial year was dramatically reversed.

'Mr President, this concludes my report on the accounts. With the help of our Auditors who are present this evening I shall be pleased to answer any questions that members might care to ask and then, if you will permit me, I would like, on behalf of the Council, to propose the adoption of the accounts as published, together with the Auditors report.'

There were no questions and Mr H. E. Drew (Fellow) seconded the Honorary Treasurer's proposal, which was carried unanimously.

Election of the Council for 1979-80

Confirming that there had not been any opposing nominations to those made by Council and circulated to corporate members in a notice dated 15th May 1979 in the



The President with the winners of the Dr Norman Partridge Premium, Mr Viktor Schiffer and Mr Alun Evans. This was the second Premium to have been received by Mr Evans.

May 1979 issue of *The Radio and Electronic Engineer*, the President said:

'We are therefore honoured that we should have as our new President Professor William Gosling.

'Mr H. E. Drew, Brigadier R. W. A. Lonsdale and Mr S. J. H. Stevens are re-elected as Vice-Presidents: Professor J. R. James, Dr P. K. Patwardhan and Mr J. Powell are elected to fill vacancies for Vice-Presidents. Colonel W. Barker and Mr R. Larry are elected to fill vacancies for Fellows; Instructor Commander D. J. Kenner, Mr B. Mann and Mr K. R. Thrower to fill vacancies for Members, Mr P. J. Hulse to fill a vacancy for an Associate Member and Mr M. W. Wright to fill a vacancy for an Associate.

'Mr S. R. Wilkins is re-elected as Honorary Treasurer, and the remainder of Council will continue to serve in accordance with the period of office laid down in Bye-law 48.



Commander Michael Ashworth is congratulated by the President on winning the J. Langham Thompson Premium. His co-author, Professor Dennis Towill, was abroad and could not be present to receive his sixth Institution Premium.

Members' Appointments

CORPORATE MEMBERS

G. D. Clifford, C.M.G. (Honorary Fellow 1978, Member 1939) retired on 18th October from the post of Honorary Treasurer of the National Electronics Council. Mr Clifford, former Director and Secretary of the Institution, had been closely involved in the Council since its inception in 1964 and was its first Secretary. A presentation on behalf of the Council was made to Mr Clifford by the Chairman, H.R.H. The Duke of Kent.

F. W. Wright (Fellow 1978) has been appointed Manager, Systems Engineering, in the TelSatCom Division of E-Systems, Inc., of Dallas, Texas. Mr Wright recently left Tehran, where he was General Manager of the Middle East Office of TAI Incorporated, an E-Systems subsidiary now absorbed in the TelSatCom Division.

Lt Col W. J. Crouch, REME (Member 1971) who was Officer Commanding Radar Division, Telecommunications and Radar Branch, REME, Malvern, has been appointed Commanding Officer, REME Technical Services, BAOR.

A. E. Fergusson-Nicol (Member 1966, Graduate 1961) has been appointed Director-General, Posts and Telecommunications

Department, Freetown, Sierra Leone. Mr Fergusson-Nicol studied at Huddersfield College of Technology and the Royal College of Advanced Technology, Salford between 1957 and 1961 and subsequently worked with electronics and communications organizations in Britain before taking up an appointment as a Telecommunications Engineer in the Sierra Leone P & T Department in 1962.

B. Garland, M.Sc., Ph.D. (Member) has been appointed a Senior Lecturer in the School of Electronic and Computer Science at Kingston Polytechnic. Formerly with Television Research in Jersey, Dr Garland was at the University of Bath from 1973 to 1969, initially as a Research Student and for the past year as a Research Officer with the Wolfson Foundation project on mobile radio.

N. O. Iube (Member 1973) who was a Principal Engineer with Nigerian Television, Benin City, has been appointed Chief Engineer at the Nigerian Television Authority Headquarters in Benin City.

Wing Cdr G. MacKinlay, RAF (Member 1974) has been appointed Officer Commanding, Royal Air Force Digby, near Lincoln. He has held a staff appointment at the Ministry of Defence for the past eighteen months.

Lt Cdr J. B. Sadler, RN (Member 1973, Graduate 1970) has completed his tour of duty as Weapons Electrical Officer in HMS *Sceptre* and has joined the staff of the Second Submarine Squadron as Staff Weapons Engineer Officer.

D. M. Pratt, B.Tech. (Member 1974, Graduate 1964) has been appointed as the Institution's representative on the City and Guilds of London Institute's Radio Amateur Examination Advisory Committee. Mr Pratt is a Training Advisor at the Northern Administration Centre of the Engineering Industry Training Board at Leeds.

K. Tempest (Member 1969) who was with the Canadian Marconi Company, has now joined NCR, Waterloo, Ontario as a Senior Engineer.

Lt Col V. Thyagarajan, B.Sc., Indian EME (Member 1977, Graduate 1970) is now Assistant Director Electrical and Mechanical Engineering (Equipment) at Army HQ, New Delhi. He was previously Officer-in-Charge of Development and Production Planning, 509 Army Base Workshops, Aggra.

NON-CORPORATE MEMBERS

S. G. Russell (Graduate 1970) has recently been appointed an Executive Engineer in the Management Services and Sciences Division of Post Office Telecommunications concerned with engineering efficiency assignments. For the past 4 years he has been with the Ministry of Defence Procurement Executive, concerned with data equipment.

Applicants for Election and Transfer

October Meeting (Membership Approval List No. 264)

GREAT BRITAIN AND IRELAND

CORPORATE MEMBERS

Transfer from Member to Fellow

HENOCQ, Cecil. *Brighton, Sussex.*

Transfer from Graduate to Member

FORDYCE, John Gill. *South Queensferry, Lothian.*
PETERSEN, David Alan. *Reading, Berks.*
PHILLIPS, Stephen Marcel A. *Aylesbury, Bucks.*
WESTWOOD, Thomas. *Paisley.*

Direct Election to Member

DHANOWA, Kashmir Singh. *Hornchurch, Essex.*
HOWELL, Peter John. *Camberley, Surrey.*
O'SULLIVAN, Barrington C. *Welling, Kent.*

NON-CORPORATE MEMBERS

Transfer from Student to Graduate

PERERA, Cyril Raomal. *Purley, Surrey.*

Direct Election to Graduate

GHURBHURUN, Kesswarsing. *London.*

Direct Election to Associate Member

JELOYSER, Francesco V. R. *Ingatstone, Essex.*
LITOBARSKI, Stuart L. H. *London.*
MURITU, Titus. *Shaoburyness, Essex.*
SMITH, Joseph Edoom. *Wembley, Middlesex.*

Direct Election to Student

DUNNE, Oliver Gerard. *Nenagh, Tipperary.*
FINN, David Joseph. *Dublin.*
FLOOD, Maurice. *Dublin.*
HEATH, Robert James. *Bath, Avon.*
KHAN, Shahid Hamid. *Heston, Middlesex.*
O'KEEFE, Michael James. *Swords, Co. Dublin.*
O'SULLIVAN, Michael William. *Swords, Co. Dublin.*
PORTER, Melvin Robert. *Cranfield, Bedford.*
RYAN, Francis P. V. *Janesboro, Limerick.*

OVERSEAS

CORPORATE MEMBERS

Direct Election to Member

CHIU, Yat Sing Raphael. *Kowloon, Hong Kong.*
CHUI, Gar Chung. *Kowloon, Hong Kong.*

NON-CORPORATE MEMBERS

Direct Election to Graduate

EZEANAKA, Gabriel C. *Enugu, Anambra State, Nigeria.*

Transfer from Student to Associate Member

AU-YONG, Wai. *Tokyo.*
THILAKARATHNA, Ranhoti B. *Sharjah, U.A.E.*

Direct Election to Associate Member

BOYNTON, David. *H.M.S. Diomedé.*
DAFE, James Erhinio. *Oba-Ile, Akure, Nigeria.*
NGAI, Charn Lam. *Hong Kong.*
SHEN, Chung Wah. *Kowloon, Hong Kong.*

Direct Election to Associate

LEONG, Laiman Raymond. *Nadi Airport, Fiji Islands.*

Direct Election to Student

FOK, Tung Leung. *Hong Kong.*
LEE, Kwai Kong. *Hong Kong.*
SI, Pui Chun. *Kowloon, Hong Kong.*
WONG, Chor Fung P. *Kowloon, Hong Kong.*
YAN, Siu Wah. *Kennedy Town, Hong Kong.*
YEUNG, Siu Wai. *Hong Kong.*
YING, Yiu Cheung. *Aberdeen, Hong Kong.*

Obituary

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

Kenneth Samuel Chapman, B.Sc. (Eng) (Fellow 1966, Member 1958, Graduate 1953) died on 8th July 1979 aged 52, leaving a widow, two sons and a daughter. Born and educated in Plymouth, Kenneth Chapman received his technical education at Plymouth and Devonport Technical College and at Northampton Polytechnical College, London, gaining the London University external engineering degree in 1953. From 1951 to 1957 he was with Marconi's Wireless Telegraphy Company, for most of the time as an engineer in the Radar Design and Test Department. In 1957 Mr Chapman moved into technical education at Mid-Essex Technical College, Chelmsford, now Chelmer Institute of Higher Education, and taught electrical and electronic engineering subjects for the CNA, College Associateship and CEI examinations. He was promoted Principal Lecturer in 1965 and Deputy Head of the Department of Electrical Engineering and Applied Physics at the end of 1978. With the reorganization of the Institute shortly afterwards, he had been appointed Head of the School of Electrical and Electronic Engineering.

In 1964 Mr Chapman published a book of 'Worked Examples in Advanced Electrical Engineering'. He was an assessor for Higher National Certificates and Diplomas in electrical and electronic engineering and a moderator in these subjects for the Technician Education Council. Just two days before his death he had been invited to join the Institution's Education and Training Committee.

William Edgar Corbett (Member 1936, Associate 1934) died in July 1979 aged 75. Between 1920 and 1928 William Corbett served as Sergeant Instructor in the Royal Corps of Signals and then entered the radio industry and in 1932 founded his own business in Liverpool. During the war he was commissioned in the Royal Signals and was demobilized with the rank of Major. He retired from his company, Cosmic Radio, in 1974 because of ill health.

Wing Commander Maurice Gordon Fountain, RAF (Ret.) (Member 1953) died on 3rd August 1979, aged 60. Born and educated in Horsham, Sussex, Maurice Fountain joined the Royal Air Force as an aircraft apprentice in 1935 and served as a wireless operator mechanic until he was commissioned in 1943. Over the next twenty-five years he held a variety of signal officer, instructor and staff appointments at home and abroad. On his retirement from the Royal Air Force in 1968, Maurice Fountain qualified as a teacher and was appointed lecturer at Yeovil Technical College, a post he held until January 1977 when ill health led to his retirement.

Frederick Livingston Hogg (Fellow 1944) died recently, aged 75 years, leaving a widow and two daughters. Livingston Hogg had a long and varied career in the radio industry ever since 1925 when he was concerned with the building of a broadcasting station in Iceland. For two years he was personal assistant to John Logie Baird and from 1929 to 1934 he was in business on his own account in Iceland. He then returned to England and for eight years was with Standard Telephones and Cables, serving as Deputy Head of the Inspection, Test and Installation Department for most of that time, and in 1943 he joined the Scientific Communications Company as Chief Engineer. In 1947 Mr Hogg set up as a specialist in the supply and installation and servicing of instruments in the communications industry. His company, later known as Livingston Laboratories, specialized in the hire of advanced instruments to research and development laboratories. He also formed associated companies including Livingston Research and Development and Livingston Controls which was concerned particularly with consultancy and supply of automation equipment. He retired as Chairman of the Livingston Group in 1972 because of increasingly poor health.

Patrick Noel Kirwan (Member 1972, Graduate 1967, Student 1956) died in Danver, Massachusetts, on 10th August 1979, aged 50, leaving a widow and one son. Born and educated in Ireland, Patrick Kirwan joined the Royal Air Force as a radar fitter in 1946 and rose to the rank of Chief Technician before retiring in 1968. He then joined the Marconi Company as a Systems Engineer at the Great Baddow Laboratories where he remained until 1970, when he went to the United States to join Fairbanks Morse as a project engineer. Since 1973 he had been a consulting engineer with the USM Corporation in Beverley, Mass.

James Arthur Ladd, B.Sc. (Graduate 1953, Student 1951) died on 3rd August 1979 aged 54 leaving a widow. After war time service in the Royal Navy, James Ladd joined the Ministry of Supply at the Explosives Research and Development Establishment, Waltham Abbey and following promotion to Experimental Officer in 1951 he transferred to the X-Ray Department. From 1954 to 1963 he worked in Canada and at the time of his death he was living in the Bristol area. For some years Mr Ladd was a member of the Committee of the South Western Section.

Horace Lysons, B.Sc. (Member 1954) died in Ottawa on 25th December 1978, aged 61. After graduating from Manchester College of Technology in 1939, he served three years as a college apprentice with a firm of electrical engineers at Openshaw. Between 1942 and 1946 he was a Technical officer at the

Telecommunications Research Establishment concerned with the design and development of ground transmitters and airborne interrogators. On the formation of the Atomic Energy Research Establishment, Mr Lysons worked on synchrotron design and later became responsible for extra-mural contracts placed with industry and universities for research and development of nucleonic instruments. He held this post until 1958 when he went to Canada to take up an appointment with the Defence Research Board of Canada.

Wing Commander Denis McDermott, RAF (Ret.) (Member 1945) died on 18th August 1979 aged 71, leaving a widow. Denis McDermott joined the Royal Air Force as an apprentice in 1924 and served as an electrical and wireless mechanic until he was commissioned as a Signals Officer in 1940, and served for several years in Coastal Command. After the war his appointments included that of Senior Technical Officer Aircraft Apprentices Wing No.1 Radio School at Cranwell and he subsequently was responsible for planning the layout of the new Radio School at RAF Locking. He retired in 1960 and took up an appointment with Decca Radar as Apprentice Training Supervisor with whom he remained until his retirement in 1973.

James William Page (Associate 1962, Student 1959) died recently at the age of 58, leaving a widow. At the time of his death he was Technical Director of ADL Technicare of Bourne End, Bucks, a company of which he was joint founder and which specialized in providing servicing and calibration for electronic equipment. Following war time service, in the Royal Navy as Chief Electrician, he had held appointments with Sperry Gyroscope and with Technivision as Chief Engineer and subsequently Technical Director.

Stuart Verrinder Rayner (Member 1959, Graduate 55) died last June at the age of 53 leaving a widow and one son. Stuart Rayner had been with the BBC for 37 years working first in the Studio Planning Installation Department and then for a number of years in the Communications Department. In 1973 he was appointed an Assistant to the Director of Engineering.

Harold James Vaughan (Member 1961) died on 6th September aged 54. Having joined the Royal Navy in 1942 and continued to serve as an instructor until 1947, he then joined C. A. Parsons and Co. as a research assistant and was for a number of years worked on the commissioning of reactor instrumentation at nuclear power stations. In 1960 he joined the Nuclear Power Company where he was concerned with control and computer applications until his retirement in 1976.

Career Selection Techniques

A Study of Data Processing Staff Selection

A three-year study, undertaken by The National Computing Centre, has thrown new light on the value of various techniques and yardsticks for the selection of candidates suitable for careers in computing. It also could have wider implications for selection of other types of staff.

An instance of particular interest is that of CSE Grade 1 holders. This qualification is generally regarded as equating with 'O' level grade C, i.e. below the average 'O' level; yet in the study, this sub-sample performed in the selection tests as well as the average 'O' level holder, but less well subsequently in the Threshold examinations.

'Data Processing Staff Selection—a Validation Study' is the result of the work of George Penney, a member of NCC staff, and A. J. Lazzerini of Nottingham University. It is based on the results of tests given to 1639 applicants for places in the Threshold Scheme, which takes school leavers between the ages of 16 and 19, regardless of school examination results, and trains them to become computer programmers or computer operators. Of the applicants 862 (52.6%) were male and 777 (47.4%) female. All had been either out of work or in dead-end jobs for periods varying from two to eighteen months.

The study showed that there are highly significant and substantial correlations between the three main factors studied—selection criteria (psychometric tests and interview); school examinations; and threshold assessments—but the remaining variation (which goes against these trends) shows that each assessment measure is tapping a slightly different combination of abilities.

The best single predictor of course assessment, and of success in finding a job, was number of 'O' levels. However, out of those who obtained programming jobs direct from the course, 9% had no 'O' levels at all, and nearly half had less than four 'O' levels. Out of the four 'O' level subject results which were monitored the best single predictor of course assessment was English Language. Maths and English Literature had about equal predictive value, still highly significant, but less so than English Language.

For the trainee sample as a whole, AH2 was the test which gave the best prediction of course results. This is a timed test of general reasoning ability at a level appropriate to school leavers, with sub-tests for verbal, numerical and perceptual (diagrams and pictures), but all three parts have some verbal component: duration 28 minutes. For Continuous Assessment, however, while AH2 provided the best predictor for the operators, the test which best predicted programmer score was one which measured clerical speed and accuracy. The test which best differentiated those who obtained programmer jobs from the remainder of the sample, was Raven's Standard Progressive Matrices. This assesses observation and clear thinking ability, and is entirely non-verbal and untimed.

A surprising result was the fact that the boys tended to score significantly higher than the girls in verbal ability. Earlier work had shown that boys surpassed girls in tests of spatial ability, but that girls were superior in linguistic ability. The report says 'Either the verbal ability of males has improved relative to females in the past 20 or 30 years, or we have, in our applicants, an untypical sample'.

An important finding of the study confirms a point that NCC has been making to schools and careers officers—but with seemingly little effect—for some years. It is that ability in mathematics is *not* the most important criterion in selecting data processing staff. Another surprising result was that success or failure in 'A' levels, or in particular 'A' level subjects, is 'irrelevant to programming'.

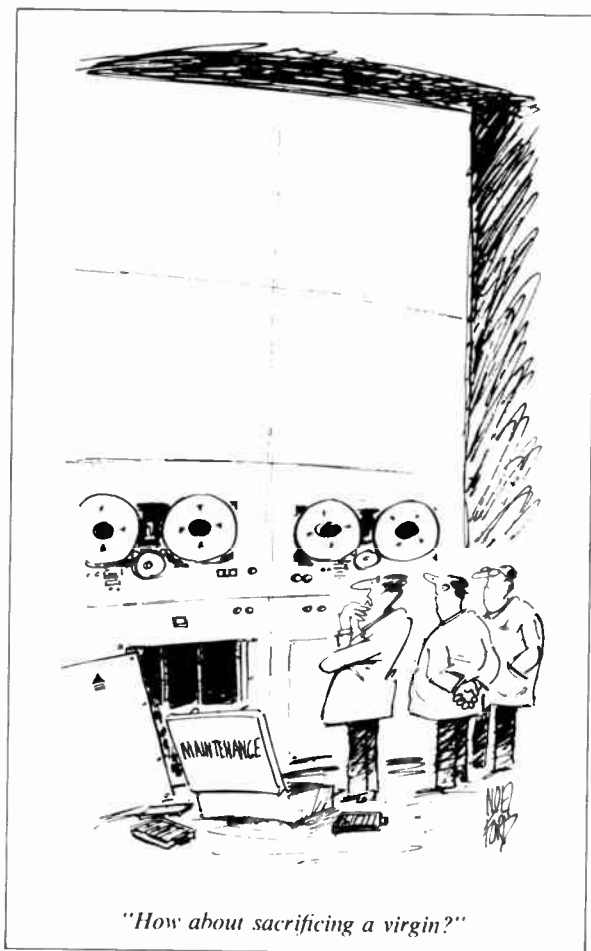
Leisure interests were also examined and the widely held view that interest in chess, bridge and crosswords would be a good indicator of ability at programming, received no support at all.

The study concludes by stating that:

'The results so far appear to give grounds for confidence in the selection procedure, as well as in the teaching of the course. There can be little doubt that most of the young people who benefit from the course would not have found any other way to escape the drudgery of work quite inappropriate to their level of mental ability. The social benefit of this alone can hardly be overestimated.

'We are aware that our findings raise a number of psychometric and educational questions which are outside the immediate purpose of the study, and which our resources and specific objectives will not permit us to follow up. We shall welcome offers of collaboration from interested bodies which can provide human and/or financial resources for this purpose.'

Copies of the report may be obtained from NCC Publications, Oxford Road, Manchester M1 7ED, price £5.00.



By courtesy of Punch

New and Revised British Standards

Copies of British Standards may be obtained from BSI Sales Department, 101 Pentonville Road, London N1 9ND.

SAFETY AND RADIO AND TELEVISION TRANSMITTERS

BSI has revised BS 3192 **Safety requirements for radio (including television) transmitting equipment** (6.40). First published in 1968, it is now identical with IEC Publication 215, the revised specification relating only to transmitters and auxiliary apparatus operated under the direction of skilled personnel.

BS 3192 deals with precautions against electric shock, skin burns, high temperature and fire, explosion and implosion, harmful radiation and other miscellaneous hazards. Design and construction requirements are given to ensure the safety of personnel both under normal and certain abnormal operating conditions and when carrying out routine adjustments, fault finding and repair operations.

STABILIZED POWER SUPPLIES

Recommendations for assessing certain performance characteristics of voltage stabilized or current stabilized power supplies have been issued by BSI in the forms of BS 5654 **Stabilized power supplies, d.c. output**. This new standard is published in three parts, namely **Part 1 Terms and definitions** (£6.40), **Part 2 Method of specifying rating and performance** (£4.80), and **Part 4 Tests other than radio frequency interference** (£8.80).

BS 5654 relates to stabilized power supplies designed to provide d.c. power from either an a.c. or d.c. source for various applications such as computers, telecommunications, laboratory and industrial equipment etc, but does not cover supplies used for electrical measurement purposes. Efforts have been made to minimize the need for specialized equipment, and no specific instruments are recommended. Test for radio frequency interference are under consideration.

All three documents are identical with the equivalent parts of the corresponding international standard IEC 478.

ELECTRONIC VOLTMETERS AND CONVERTORS

The British Standards Institution has published BS 5704 **Method for specifying the performance of digital electronic d.c. voltmeters and d.c. electronic analogue-to-convertors** (IEC 485) (£8.80). This new Standard is applicable to voltmeters and convertors, as well as accessories when these have been associated with them at the time of calibration. Some parts of the Standard may also be applicable to other types of analogue-to-digital converting apparatus, e.g. digital ohmmeters and ammeters, but the standard is not intended to provide complete technical requirements in these cases. Non-linear conversion is not explicitly considered.

The Standard contains definitions, general requirements for statements on functional performance, general requirements for tests, constructional requirements, marking and symbols pertaining to apparatus and accessories. It discusses operating and intrinsic error, influence errors and variations, and stability errors, and gives some further requirements concerning statements and tests on apparatus. Details for the expression of characteristics of digital voltmeters and analogue-to-digital convertors are included.

SAMPLING FOR ECONOMICAL QUALITY CONTROL

The British Standards Institution's newest contribution to the field of quality control is BS 6002 **Sampling procedures and charts for inspection by variables for percent defective** (£10.70). This British Standard specifies sampling plans and procedures for inspection by variables. The object of the methods laid down in this standard is to ensure that lots of an acceptable quality have a high probability of acceptance and that the probability of rejection of inferior lots is as high as possible, in accordance with the principles of quality control established over 40 years ago in BS 600. Although variables sampling schemes have been little used in industry in general to date, it is anticipated that with the advent of modern and relatively inexpensive calculators variables sampling schemes will be more widely used.

In four sections BS 6002 gives comprehensive guidance, operational instructions, tables and diagrams providing three alternative methods of sampling inspection indexed by Acceptable Quality Level (AQL) that take account of test data variability and enable relatively smaller sample sizes to be specified, compared with complementary attribute plans given in BS 6001.

The publication of this standard thus enables AQLs to be specified for various product characteristics with reference to both BS 6001 and BS 6002. Consequently users may choose either the simpler attributes scheme, a more economical variables scheme (in terms of sample size), or a combination of both with respect to a particular product. Dimensional or 'simple' functional requirements might be assessed in accordance with BS 6001, and destructive or 'expensive' test requirements in accordance with BS 6002, since maximum advantage can be taken of the quantitative information provided by the inspection or test results. Accordingly, application of the new standard is particularly recommended where the inspections or test methods employed are expensive or destructive.

EQUIPMENT RELIABILITY TESTING

Applicable to all types of equipment such as electronic, electromechanical and mechanical, the newly published DD 57 **Methods of equipment reliability testing** (£7.80) gives the general principles, as well as the methods and procedures, for equipment reliability testing.

Three distinct test areas are described:

- (a) Reliability determination test To determine the value of a reliability characteristic of the equipments tested.
- (b) Reliability growth test To expose and eliminate weaknesses, usually carried out as part of the development process.
- (c) Reliability compliance test To show whether or not the value of the reliability characteristic of the developed equipments complies with the specified reliability requirements.

Guidance is also given in the preparation of a reliability test procedure document.

The tests are not intended to replace existing functional and environmental tests. Further evidence is required before these tests could be incorporated into a British Standard and consequently users are urged to write to BSI giving details of their experience in applying this publication.

Automatic Call Recording Equipment

Trials of a new Post Office equipment have recently started at the Eastbourne telephone exchange which it is expected could save up to 20% of an operator's time in connecting calls. This would lead to savings in operating costs if applied nationally to all inland operator calls, and enable operators to give more time to helping telephone users.

The new equipment, known as ACRE (automatic call recording equipment), employs microprocessor techniques to help telephonists connect calls. It cuts out the time-consuming task of writing call details on the paper 'ticket' at present used for transferring the charge for the call onto the customer's phone bill.

Whenever a customer dials 100 to ask the operator for help with a call, the telephonist who answers has two main jobs to do: she has to put the call-charging details on a printed paper ticket, and then she has to set up the call. This means repeating the number she has already marked on the ticket, by pressing the buttons of her keypad (now used instead of a dial). She also sets a timer ready to start timing the call after the called person answers.

Then for a few seconds, she can do nothing but wait—until she hears the ringing tone which tells her that the phone at the other end is being rung. At this point she switches out of circuit.

On average, an experienced operator takes about 100 seconds to make out a ticket and set up an ordinary call in this way. And she still has to keep an eye on the call indicator light so that when it goes out—showing the call has ended—she can record the duration.

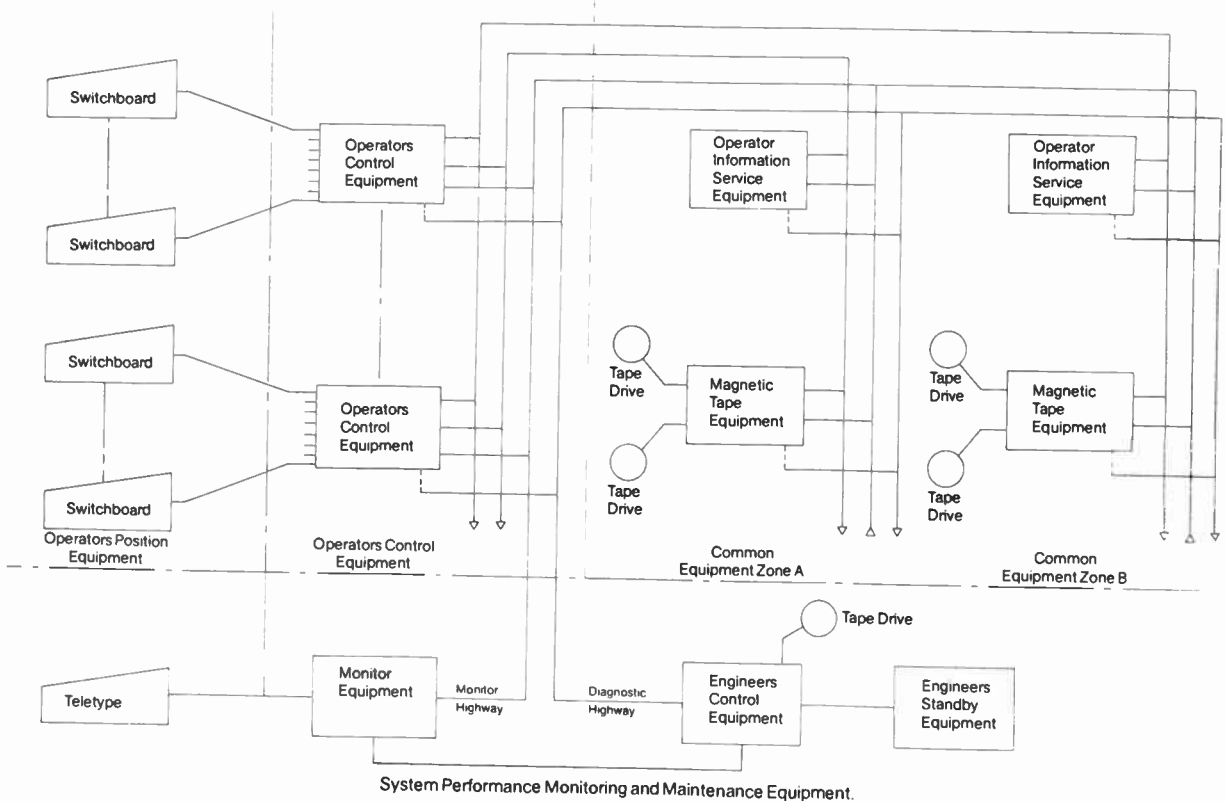
A special facility call—such as credit card, or transferred charge—means extra work. There is more information to put on the ticket—such as the card number for a credit card call—and the operator may have to call back afterwards with ADC (advice of duration and charge), if requested.

These, then, have been the problems.

With ACRE, the ticket is eliminated and all the call details are recorded automatically, at the touch of a button. At an ACRE switchboard, the operator keys the number only once on her keypad, and ACRE's microelectronic 'brain' automatically logs the number for billing purposes. While the call is going through, the operator is busy keying in any other details required. ACRE has its own clock, and a complete store of all charging information; it records charging rate and call duration automatically. And it even gives the operator an automatic reminder to call back if ADC was requested.

Developed wholly by the British Post Office, Automatic Call Recording Equipment (ACRE) is thus designed to automate many of the tedious and time-consuming tasks which have to be undertaken by switchboard telephonists. Its main feature is to record call details on magnetic tape for subsequent data processing. This eliminates the need for the call ticket and improves operator efficiency with corresponding beneficial effects on customer service.

The system uses advanced distributed microprocessor techniques for maximum flexibility, security and long-term reliability. Its performance is monitored continuously and full



Simplified schematic of automatic call recording equipment

diagnostic and maintenance aids enable faults to be readily located, identified and repaired.

Excluding system performance monitoring and maintenance facilities, ACRE has three main components:

- the operator's position equipment (the switchboards),
- operator's control equipment, and
- common equipment.

The accompanying schematic shows the basic system architecture. Each operator's control equipment normally services up to eight switchboards but may handle up to 16 under fault conditions. This protects the switchboards against failure of their associated control equipments by automatic changeover to the standby ports of adjacent equipments. The common equipment—duplicated for security—can handle up to 24 control equipments. A total system may therefore serve from 16 to 192 operators.

Both common and operator's control equipments are designed around a standard micro-computer module developed especially for ACRE. This in turn is based on the Intel 8080A microprocessor, and considerable use is also made of low-power Schottky TTL and i.c. relay drives.

Focal point of the ACRE switchboard layout is a direct-current electroluminescent 80-character alpha-numeric display panel on which call information appears as it is keyed into the system. The board includes a standard telephone keypad and control keypads used to input specific call information or instructions. Great care has been taken to ensure optimum ergonomic layout.

Rack-mounted equipment associated with the switchboards may be installed remotely, with the microprocessor units up to 200 m from the switchboards. Connection between the switchboards and the racks involves a specially developed multiplexing system using no more than a single pair of wires.

The operator's control equipment provides the control instructions for a wide variety of switchboard facilities. It tests the validity of information keyed in at the switchboard, and of line supervisory signals, before taking action, and it stores information for the switchboards—sending it continuously to them to activate the display, light signal lamps, send routing digits, etc.

The capability of the operator's control equipment is augmented by duplication of the common equipment, both sets of which are accessed simultaneously by each control unit over dedicated highways. This provides two-way security: on the one hand, control equipment failure cannot impair common equipment performance; on the other, the control equipment continuously compares information from each common equipment zone, and takes remedial action if they disagree. Moreover, if both common equipment zones fail, the control equipment has recourse to recovery procedures and service is not interrupted.

The common equipment houses most of the system's dependent data, enabling all control equipments to be simultaneously up-dated. Moreover, each common equipment zone can be up-dated independently of the other, to avoid service interruption.

The basic ACRE system is supported by three further microprocessor-based units: monitor equipment, engineer's control equipment and engineer's standby equipment.

The monitor equipment receives reports of system performance. If this departs from normal, the monitor automatically undertakes a pattern search to identify the fault, raises an alarm and provides a printout of the fault information. In addition, it collates operational statistics to aid switchroom management.

The engineer's control equipment supplements the monitor by pinpointing faults. Its repertoire of diagnostic instructions

enables non-specialist staff to localize a fault to a printed circuit board assembly.

The standby equipment provides a full range of functional printed circuit boards as maintenance spares for suspect units. Its store of boards is continuously tested by the engineer's control equipment to ensure their immediate availability. Suspect units may be tested in the standby equipment to confirm whether or not they are actually faulty.

Software reliability and a high standard of documentation are prime considerations in the ACRE design. The need for software queues has been virtually eliminated by the distributed microcomputer architecture, which ensures that inputs are queued in hardware and each microprocessor is lightly loaded. This results in a particularly simple software structure in which programs are event-driven and simultaneous access to data avoided.

The ACRE software has been developed using the high-level programming language PL/M, structured programming techniques, decomposition of the programs on a functional basis and a high standard of commenting.

Total program size is of the order of 30 40 K lines of high-level language. Operator's control equipments account for 8 K lines, common equipment some 12 K lines and the balance is taken by the maintenance and monitor equipment. There are also comprehensive diagnostic programs which may be loaded when required.

Standard Frequency Transmissions

(Communication from the National Physical Laboratory)

Relative Phase Readings in microseconds
NPL—Station
(Readings at 1500 UT)

OCTOBER 1979	MSF 60 kHz	GBR 16 kHz	Droitwich 200 kHz
1	-3.3	4.0	56.3
2	-3.3	3.9	56.0
3	-3.3	3.9	55.7
4	-3.3	4.5	55.5
5	-3.3	4.2	55.2
6	-3.2	4.7	55.0
7	-3.3	4.1	54.8
8	-3.1	4.8	54.5
9	-3.1	5.0	54.2
10	-3.3	4.7	53.9
11	-3.1	5.2	53.7
12	-2.9	4.7	53.5
13	-3.1	4.6	53.2
14	-3.1	5.2	53.0
15	-3.1	5.3	52.8
16	-3.1	5.2	52.5
17	-3.0	4.9	52.2
18	-3.0	5.4	51.9
19	-3.0	5.2	51.5
20	-3.2	4.2	51.2
21	-3.1	7.7	51.0
22	-3.1	6.4	50.7
23	-3.2	10.2	50.5
24	-3.3	8.0	50.3
25	-3.2	8.7	50.1
26	-3.3	8.5	49.9
27	-3.0	8.2	49.7
28	-3.1	8.7	49.5
29	-3.0	9.9	49.3
30	-3.2	9.5	49.1
31	-3.1	9.2	48.9

Notes: (a) Relative to UTC scale (UTC_{NPL}-Station) = +10 at 1500 UT, 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 10¹¹ per day.

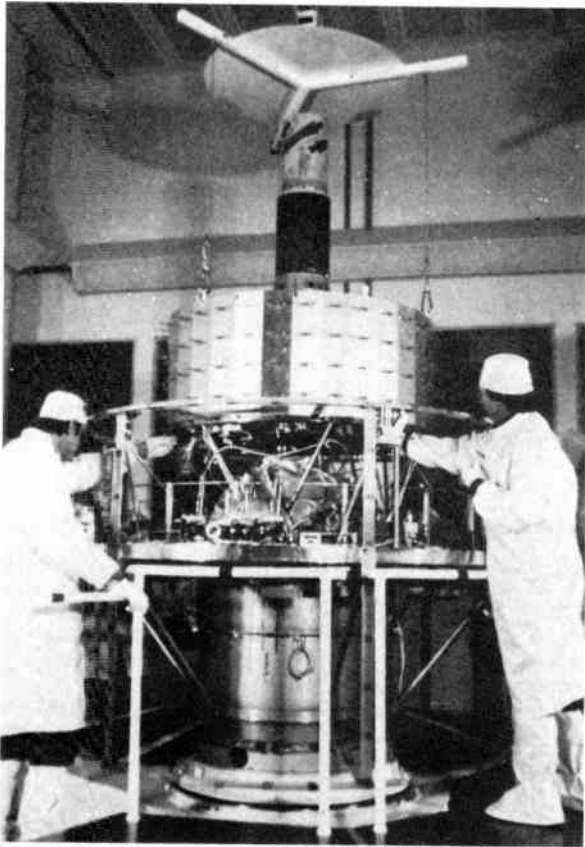


Fig. 3. A photograph of *Meteosat* taken during laboratory tests.

to measure both the incoming radiation and that upwelling from the Earth and atmosphere and hence the net radiation balance of the earth-atmosphere system as a whole.

The satellite also carries a dedicated communications

system. Raw image and other meteorological sensing data and spacecraft housekeeping data are transmitted on an S-band channel to the Data Acquisition, Telemetry and Tracking Station and thence to the *Meteosat* Control Centre, both located near Darmstadt. Processed images and other data from the Control Centre and Computer System are then transmitted back to *Meteosat* for onward relay to other user ground stations. However, many of the meteorological products such as winds, cloud- and sea-surface temperatures, radiances, etc., produced by the two large ICL 2980 computers in Darmstadt, are sent by land-line to national meteorological centres. The satellite system also provides for two-way u.h.f. communication with up to 66 automatic data collection platforms such as ships, aircraft, drifting buoys, freely-floating balloons and other satellites and for the transmission of the information received to the Control Centre—see Fig. 4.

During the Special Observing Periods (SOPs) of the Global Weather Experiment (GWE) an average of about 6600 wind vectors per day were produced by the five geostationary satellites of which about 2250 per day were over the equatorial tropics (10°N to 10°S). Winds determined by two different satellites in regions where they overlap were in excellent agreement but there were discrepancies in the assignment of cloud-top temperatures and pressures, of up to 10 K and 50 mb respectively, the reasons for which have not yet been resolved. A map showing the coverage of low-level wind vectors including a large proportion from the US geostationary satellites and some from *Meteosat* is reproduced in Fig. 5.

The US Tiros 'N' Satellites

Tiros 'N' is the third generation of US operational polar-orbiting satellites and incorporates several major improvements over the earlier series. It provides pictures of greater resolution both in the visible and infra-red parts of the spectrum to give day and night cover, it also provides improved vertical temperature and water vapour profiles using both

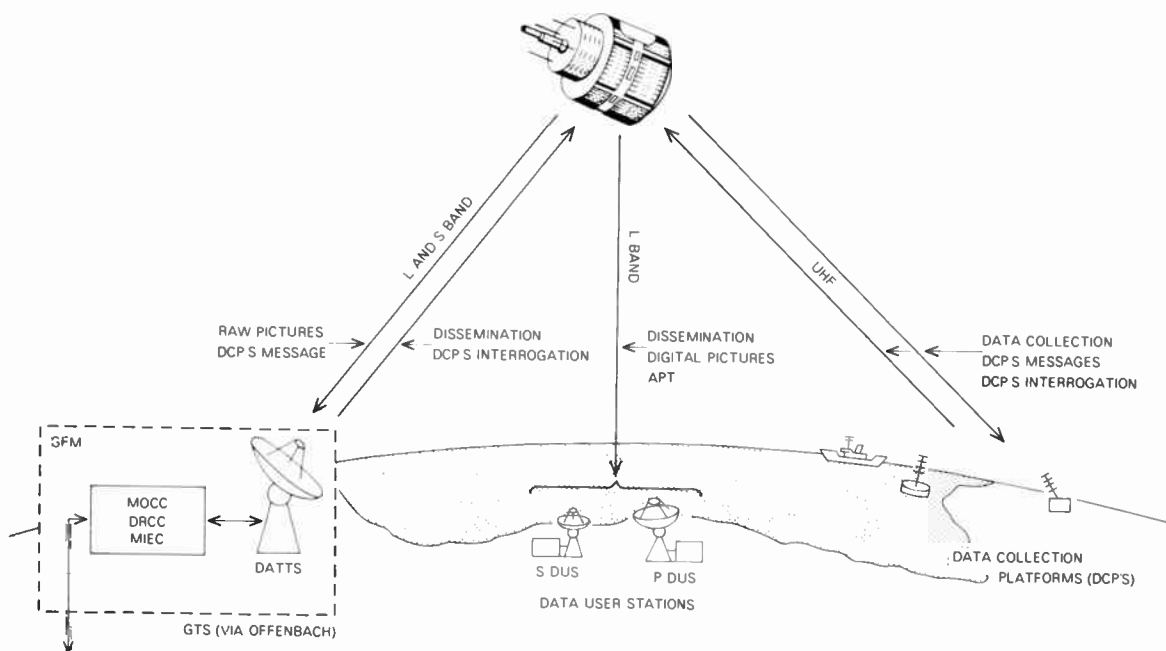


Fig. 4. A diagrammatic representation of the complete *Meteosat* system.

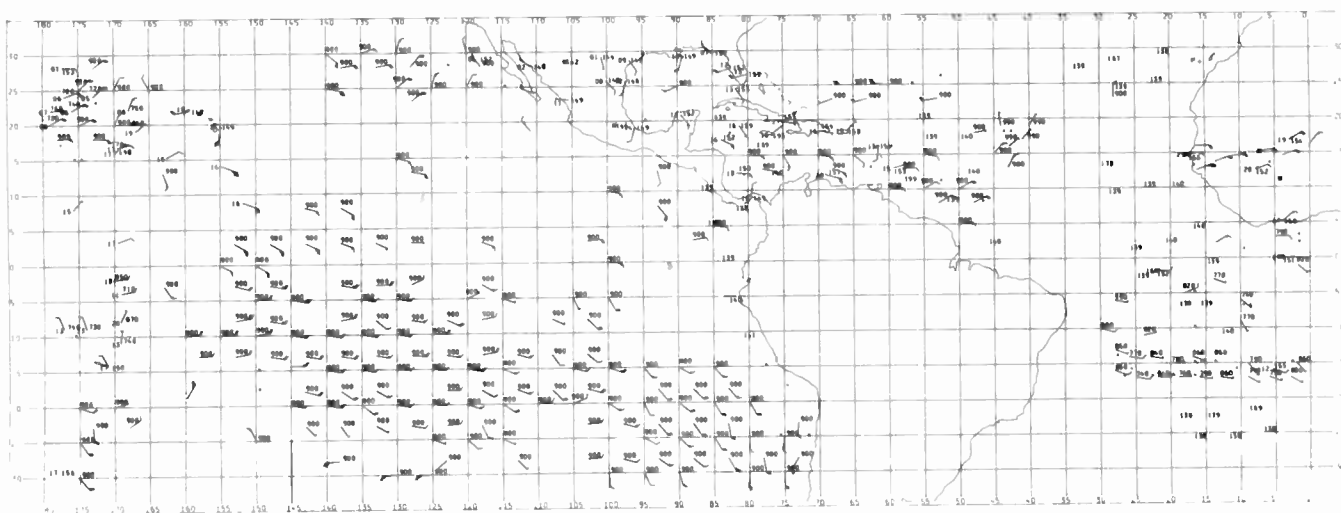


Fig. 5. The distribution of low-level wind vectors over the Pacific and Atlantic Oceans on 7th February 1979 mostly derived from cloud displacements as recorded by the SMS and *Meteosat* geostationary satellites.

infra-red and microwave sensing; and carries the first-operational system for the location of, and data collection from, a variety of observing platforms.

The first *Tiros 'N'* spacecraft was launched on 13th October 1978 in near-polar, sun-synchronous orbit with an average altitude of 870 km and a period of 102 min. The second satellite was launched on 27th June 1979 to provide two spacecraft with orbits spaced 6 hours apart to give good space and time coverage, every part of the earth being scanned four times a day.

The spacecraft, sketched in Fig. 6 is a five-sided box, 3.7 m long and 1.7 m in diameter, weighing 1410 kg. The Earth-oriented platform containing the meteorological instrumentation is stabilized for all axes within 0.2° .

The latter includes an Advanced Very High Resolution Radiometer (AVHRR) with one visible and three infra-red channels to provide images of clouds and surface features with 1 km resolution at the sub-satellite point and also measurements of sea-surface temperature. In essence it consists of a simple 20 cm aperture telescope with a scanning mirror that sweeps over a 2200 km swath of the Earth's surface and focuses the received radiation on to a silicon photo-detector and on to cooled mercury-cadmium telluride and indium antimonide detectors to produce both visible and infra-red images. The pictures are built up line by line by the mirror scanning from horizon to horizon at such a rate that the satellite advances along its orbital path just enough between each scan to provide contiguous cover of the Earth's surface. Ground stations within range of the satellite can receive direct, and in real time, pictures of median (4 km) or high (1 km) resolution and all global data are stored on board for transmission to a few major ground control centres where it is processed after some hours delay.

The satellite also carries an advanced vertical temperature sounder consisting of three separate units, all of which make global vertical temperature soundings of the atmosphere by measuring, at several different wavelengths, the infra-red or microwave radiation emitted by constant constituents such as carbon dioxide and molecular oxygen. A detailed description of these techniques may be found elsewhere^{1,2} and will not be repeated here.

The tropospheric sounding unit has 20 channels, most of which lie within the $4.3 \mu\text{m}$ and $15 \mu\text{m}$ carbon-dioxide bands and are used to determine the temperature profiles from the Earth's surface up to heights of about 30 km. There are two infra-red channels in which absorption by the atmosphere is low and permits the measurement of sea-surface temperatures. Three channels in the $6.7 \mu\text{m}$ band are used to estimate the concentration of atmospheric water vapour at three tropospheric levels. The resolution of this sounding instrument is 20 km at the sub-satellite point.

The microwave sounding unit with four channels in the 6 mm band of molecular oxygen determines the vertical temperature profile on the same principle as the infra-red instrument, but although its resolution (110 km) is poorer at these longer wavelengths, it has the great advantage of being able to 'see through' all but heavily-raining clouds.

The stratospheric sounding unit, based on the ingenious pressure modulator radiometer developed at Oxford University,^{1,2} is supplied by the UK Meteorological Office. It has three channels in the $15 \mu\text{m}$ band of carbon dioxide and provides vertical temperature profiles of the stratosphere up to levels of about 50 km. The UK Meteorological Office has undertaken to provide daily global analyses of stratospheric temperature and pressure fields based on the data from this instrument and the top three channels of the tropospheric sounding unit joined on to the fields at 100 mb (16 km) level obtained by conventional methods.

All systems on *Tiros 'N'* appear to be working well at the time of writing and to be giving images, temperature and humidity profiles of higher quality than have been obtained before. The differences between temperature profiles deduced from the *Tiros 'N'* infra-red and microwave sounders and those obtained from nearby conventional radiosonde soundings are shown in Fig. 7. During the latter part of SOP 1 an average of about 1100 such temperature soundings per day were received at Bracknell and incorporated in a global analysis. In the extratropical regions of the Southern Hemisphere these soundings were frequently the only source of upper air information and made a significant contribution to the upper-level analyses—probably more so than in the Northern Hemisphere where there are much more conventional data.

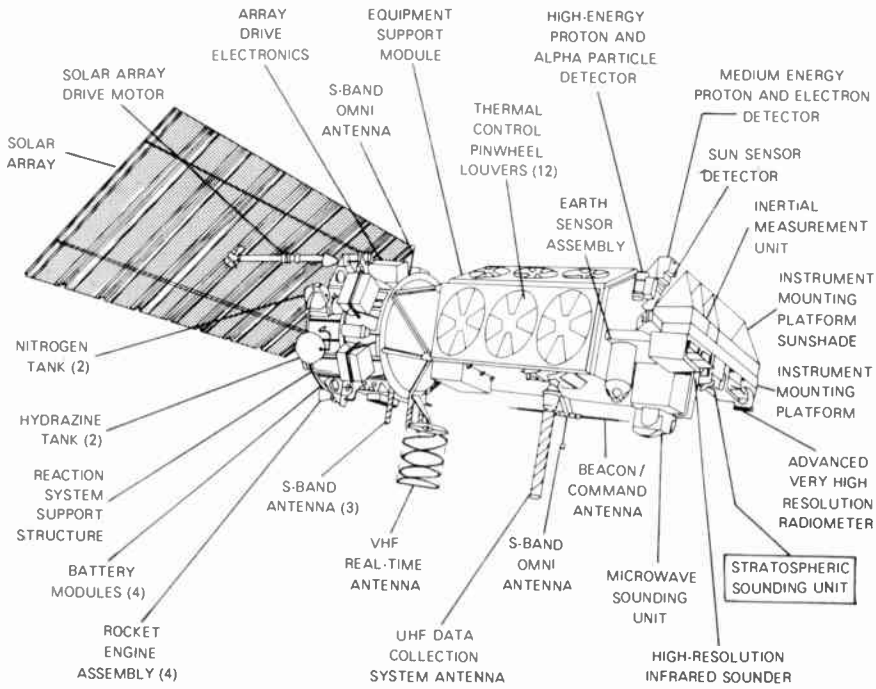


Fig. 6. A sketch of the *Tiros 'N'* satellite showing its main components.

Tiros 'N' also carries a data collection system, ARGOS, provided by France, which is able to locate, and receive data from, a variety of automatic observing platforms such as freely drifting ocean buoys and balloons, anywhere in the world including the polar regions which are inaccessible to geostationary satellites.

Other Special Observing Systems for GWE

In addition to the satellites just described, the conventional surface and upper air observing systems have been augmented for the GWE by several special systems designed to fill the most important gaps in the global surveillance of the atmosphere. These gaps are particularly serious over the oceans and in the tropical regions.

Some 45 ships have been specially equipped to make wind soundings in the tropics and have provided about 1750 soundings during the intensive period of each of the SOPs, mainly in the equatorial Atlantic, in the Indian Ocean and the SE Asian seas. Unfortunately the ship coverage was poor in the tropical Central and Eastern Pacific but here about 2500 successful soundings were made during each SOP intensive period by dropping sondes from dedicated aircraft.

The two systems together provided an average of about 135 soundings per day—about twice that achieved by the conventional land-based upper air network in the equatorial tropics.

Measurements of winds in the tropics were further augmented during the SOPs by sending up constant-volume inextensible balloons arranged to float at a constant (density) level while being carried round the globe by the winds. The balloons are continually tracked and located by the ARGOS Doppler system on the *Tiros 'N'* satellite, the winds being computed from the balloon displacements. During the two SOPs 310 balloons were released from the islands of Ascension,

Canton and Guam; unfortunately many were lost or had to be cut down on entering the Northern Hemisphere, but on average about 100 wind vectors per day were obtained. The tracks of two balloons released from Ascension Island are shown in Fig. 8.

In addition some 80-90 large commercial airliners fitted with accurate wind-finding Doppler equipment, produced about 2000 reports per day, a few aircraft being fitted with a special device which automatically transmitted the reports to the *Tiros 'N'* satellite which then relayed them rapidly to ground stations.

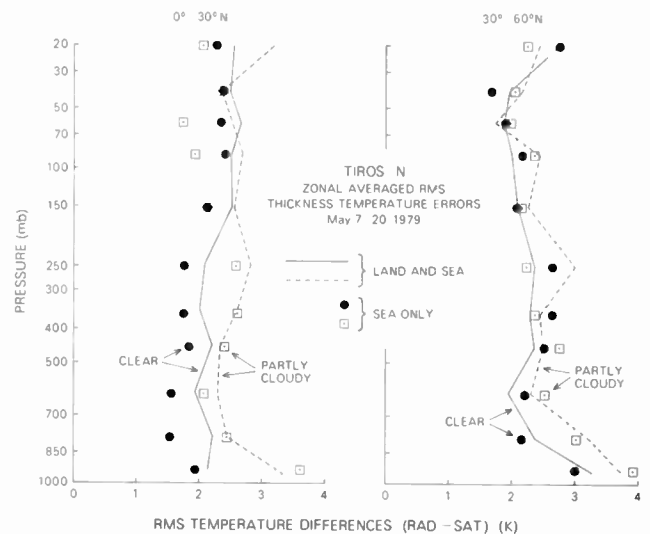


Fig. 7. Differences between vertical temperature profiles deduced from *Tiros 'N'* infra-red and microwave radiometers and those obtained by conventional radiosonde soundings.

One of the most successful innovations of the GWE has been the deployment and interrogation of specially-instrumented freely-drifting buoys in the southern oceans. Some 125 buoys were launched by the end of the first SOP and more than 200 by the end of the second period. Located and interrogated by satellites they provided measurements of surface pressure and temperature with an average spacing of < 500 km over 80% of the southern oceans south of 20° S. The distribution of buoys on 5th March 1979 is shown in Fig. 9.

Analysis and Preliminary Assessment of GWE Data

The adequacy and quality of the meteorological data obtained from the satellite and other special observing systems of the GWE can best be assessed by incorporating them (together with conventional data) into three-dimensional global analyses depicting the physical state of the atmosphere at a particular instant of time and using such an analysis as the starting point for a numerical prediction of the future evolution of the atmosphere over several days. Alternatively, the data may be assimilated into a *continuously running* model of the atmosphere such as has been specially devised by the Meteorological Office to cope with the continuous stream of satellite and aircraft observations, most of which are made outside the fixed synoptic hours of the conventional observing

system. These 'off-time' observations gathered over a 3-hour period are incorporated into an 11-level model of the atmosphere, winds and temperatures being introduced as separate analyses, balance between the wind and temperature fields being achieved through the model's equations. Weighting functions are assigned to the various types of observations according to their known observational errors, their reliability and coverage. Very little difficulty has been experienced in integrating very different types and sources of data into the model which generates consistent and realistic continuous fields even in regions where the data are relatively sparse or inhomogenous. Rarely is the model shocked into instability by the injection of new data and it may be stopped at any convenient time to produce forecasts up to about five days ahead.

However, these may be regarded as only preliminary tests of the GWE data arriving at Bracknell in real time; complete analysis and assessment of the GWE observations will have to await the arrival of the full data sets but meanwhile some first impressions may be given.

A great deal will depend on the accuracy and consistency of the temperature profiles provided by the infra-red and microwave sounders on the polar-orbiting satellites and on the winds derived from the cloud displacements mapped by the geostationary satellites. As we have seen, there are some

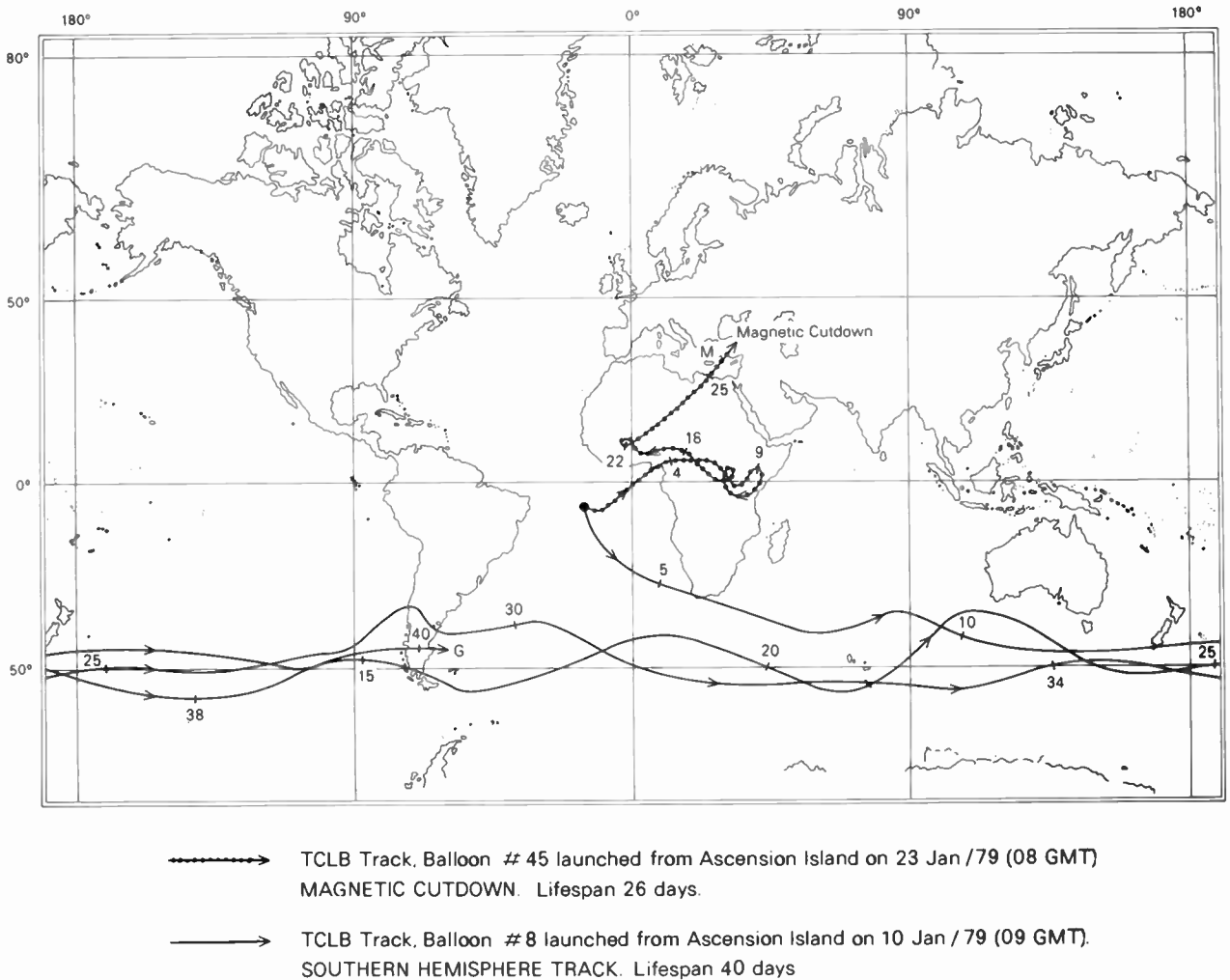


Fig. 8. The tracks of two constant-level balloons released from Ascension Island followed by the ARGOS satellite system over periods of 26 and 40 days.

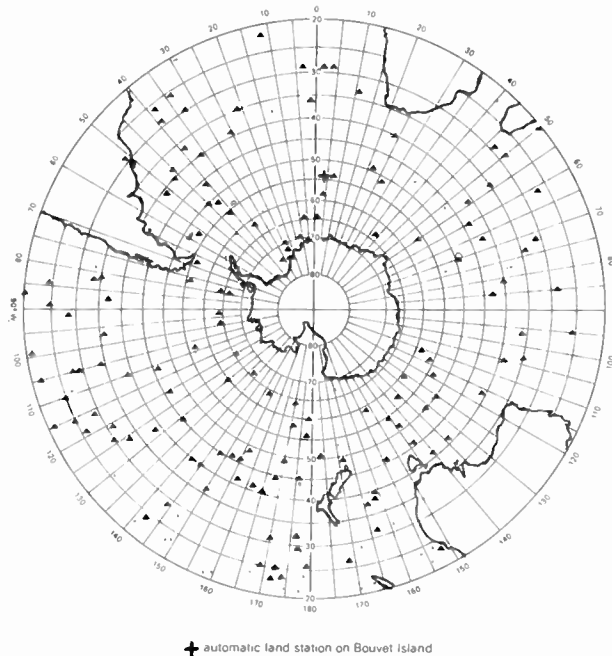


Fig. 9. The distribution of drifting buoys making automatic measurements of surface pressure in the Southern Oceans in March 1979.

problems still to be solved but the temperature retrievals from *Tiros 'N'* are a good deal better than those obtained with earlier satellites and may well prove adequate to meet the basic requirements of the GWE. If this proves to be the case then the pressure, temperature and wind fields will probably be defined with sufficient accuracy to meet the GWE objectives at least in the extra-tropical regions of the Northern Hemisphere. In the Southern Hemisphere, which has few land-based stations, there

is even greater dependence on the satellite data but the surface pressure distribution is likely to be defined with sufficient accuracy by the drifting buoys over about 80% of the southern oceans. Although there are systematic differences between sea-surface temperatures as measured by satellite and by ships, there is good reason to believe that this problem can be resolved to give these temperatures in both the northern and southern oceans with sufficient accuracy.

The problems are greatest in the tropical regions where the mass and wind fields are only weakly coupled and where the latter cannot be inferred from the former and so have to be determined independently. This puts a premium on the special wind-finding systems—the ships, aircraft and constant-level balloons deployed especially for this purpose. Preliminary analyses indicate that they will be inadequate to meet GWE objectives but if wind vectors provided by the geostationary satellites are also included, analyses such as those of Fig. 5 may be just adequate.

Although the original plan for GWE has not been fully implemented, the global atmosphere will have been observed during this year more comprehensively and accurately than ever before. This will be due, in large measure, to the crucial contribution and excellent performance of the satellite systems without which continual surveillance of the global atmosphere would be neither technically nor economically feasible.

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Television measurements through psychophysics to subjective picture quality

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SUMMARY

The most important property of a television system is the quality of the received picture as judged by the user. An explanation is offered of the way in which human beings quantify picture quality and the apparent magnitude of picture impairments. Although the explanation is speculative, it is hoped it will prove intuitively attractive, and will assist both the interpretation of the results of visual subjective studies and their application in terms of objective measurements. The psychological scale of apparent magnitude turns out to be also the scale along which co-existing impairments may be simply summed to determine their total effect. It is thus closely related to the IMP scale which has already found practical application.

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1 Introduction

In this paper it will be taken as axiomatic that the most important property of a television system is the quality of the received picture. To simplify discussion it will be further assumed that the most appropriate criterion of quality is subjective quality as judged by the user, although some of the principles to be described could be applied to more specific measures such as target identification, legibility (in the case of alpha-numeric material), and so on. Attention will be devoted primarily to the essence of the relationship between picture impairments measured in objective terms on the one hand and subjective picture quality on the other. For this reason, it is possible to deal only cursorily with the many related topics, and a blanket apology for many omissions is made at the outset.

To rate television picture quality the most widely used technique at present is category scaling. Because we are interested in a relationship over a range of magnitudes, we direct our attention to the results of 'full range' laboratory experiments under controlled conditions that can be repeated as required. For many impairments of practical interest it is readily possible to achieve a range of magnitudes sufficient to ensure that the majority of the observers utilize the whole of the rating scale.

In embarking on studies of this kind, we are entering the field of psychophysics where, if we care to look, we will find a mass of information about subjective response to stimuli of all types, and about the experimental methodology for measuring it. Without questioning *how* the human mind works, it is proposed to offer an explanation of *the way* in which we quantify perceived effects which, although speculative, may prove intuitively attractive and which can provide a sufficient understanding for practical purposes. It leads to an empirical relationship between apparent stimulus magnitude and subjective rating which does not yet seem to be widely accepted among psychophysicists. A possible reason for this state of affairs is the fact that we are dealing with a somewhat unusual class of stimuli. Rudimentary stimuli, for example light, heat, pressure, can usually be applied only in limited magnitudes because of the pain or damage they would otherwise cause. We are here concerned with more sophisticated stimuli, not subject to such constraints. For all practical purposes, impairment magnitude can be made infinite in relation to signal magnitude. Yet the resulting display can be arranged still to be comfortably viewable, even although it is useless for normal purposes.

2 Objective Limits and Tolerances

Important reasons for making measurements on television transmission systems and links are as follows:

1. Acceptance Testing. To allow the initial performance of a link to be checked against that specified.
2. Maintenance. To enable faults causing sub-standard performance to be remedied, preferably

before the performance of a complete system has been noticeably affected.

3. Assessment of Performance. To enable long-term performance to be assessed in terms of that which is desirable.
4. Planning Methods. To provide the data necessary for the improvement of planning methods, so enabling the performance of future links to be better related to requirements.

Macdiarmid² has described the procedure necessary to set performance limits for an individual link of a long chain, using the desired overall subjective picture quality as a starting point. In this connection, it is important to bear in mind that impairments arising in a link affect the received picture only when associated with the impairments arising in all the other links of the chain. Hence the subjective performance of one link alone is not usually of primary relevance, although it can often provide useful guidance.

3 Subjective Rating

It has generally been considered impracticable to obtain assessments of picture impairments in the conditions under which a television system is actually used. To obtain a stable subjective yardstick, rigorous standardization of test conditions is essential, and is most readily achieved in the laboratory. The CCIR in its Recommendation 500 has made considerable progress in formulating a list of agreed test conditions, and Ref. 3 includes an up-to-date general review. It has yet to be satisfactorily demonstrated that independent organizations can produce results based on equal absolute standards of subjective assessment, or alternatively that differences in standards are reliably constant. Also, because the laboratory test situation is inevitably artificial to some extent, the interpretation and application of results still remains a matter for discussion. Nevertheless, gradual progress is continuing.

4 Psychological Scales

The procedure for rating the quality of television pictures involves the use of the following 5-category scale recommended by the CCIR¹:

Excellent
Good
Fair
Poor
Bad

Most viewers, when asked to form their opinion, would probably agree that they envisage picture quality on some sort of psychological continuum. The continuum is mentally subdivided into the required number of intervals or grades, five in the case of the above scale, and a particular picture is rated by deciding which interval contains its particular level of picture quality. Such categorization, like any quantizing process,

inevitably loses information. By having opinions recorded directly on a continuous scale it should be possible to avoid the loss. However, up to now it has been generally thought that category scaling is most suitable for experiments employing naive (non-expert) subjects, many of whom seem to feel that the verbal cues of the grading scale make their task easier. In any case, for the range of qualities found in present-day broadcast television, the variability of opinions proves to be comparable with the category size. Thus the actual loss of information is small.

It would be possible to speculate endlessly as to the reality of the psychological continuum. The number of opinion levels which the finite human brain can register must itself be finite, and so the assumption of a related continuous scale cannot be strictly correct. Another matter is the equality of the grading intervals or otherwise, and in particular whether the terminal intervals are finite or infinite. The acceptability of the remainder of this paper depends heavily on the intuitive appeal of a finite scale as the basis of subjective judgments. The premise will therefore first be justified, at least for judgments which are made in what are essentially qualitative terms even although later subjected to quantitative interpretation.

It is in the nature of the problem that all the evidence is indirect. One approach is to note that it is possible to categorize the impairment in the picture rather than the quality of the picture. For example the CCIR recommendation¹ currently includes a 5-grade scale based on the degree of annoyance caused. Thus there exist two scales, one being the converse of the other. Surely it is unlikely that the category representing very small or zero impairment would extend to infinity. Yet results suggest that to a first approximation the scales can be directly equated, so that 'excellent' is linked to the finite, very small impairment, category. If 'excellent' is a finite category then 'bad', i.e. the zero quality end of the scale, is even more likely to be so.

Another reason for supposing that the scale is finite is that one viable technique for obtaining opinions is in fact by having them expressed in terms of their equivalent position along a simple straight-line scale. One end of the scale is taken as representing, say, perfection and the other as the opposite extreme. Apart from the end points the scale is uncalibrated. It is obvious that at least in this situation a finite scale exists which must be closely related to a subjective continuum. For the recording scale to have one or both ends at infinity is clearly not worth consideration—even if the impossible task of creating the scale could be accomplished, results marked on it could not be interpreted!

In the end it is necessary to come to a personal conclusion as to the nature of the quality grading scale. If the idea of opinion formation by subdivision of a finite psychological continuum into five roughly equal

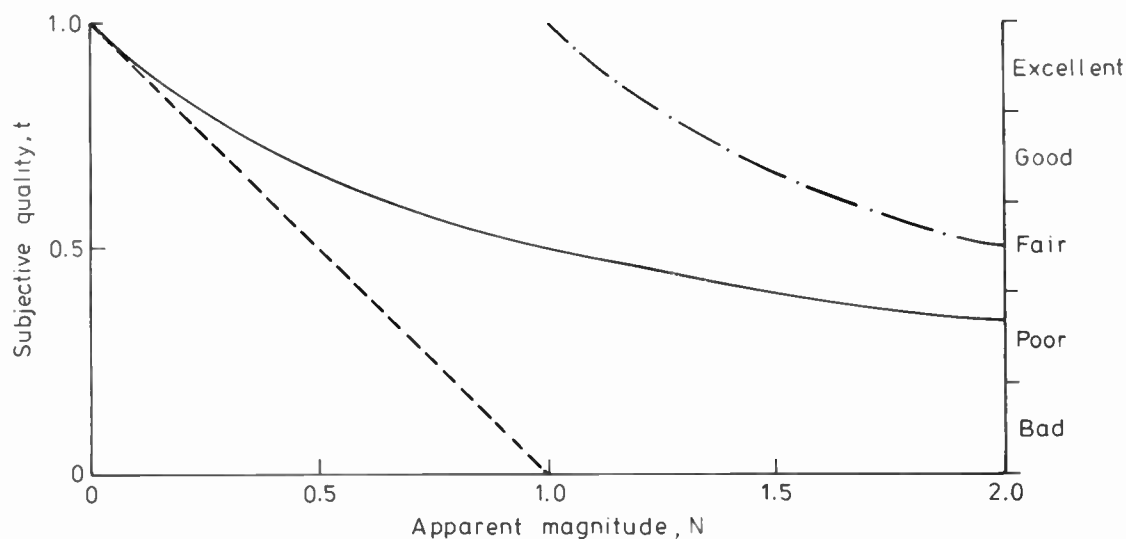


Fig. 1. Quality related to impairment magnitude.

intervals seems unlikely, then the remainder of this paper will be somewhat unconvincing. However, for some it may be sufficient to treat such a scale purely as a convenient mathematical device. For greatest simplicity, the psychological continuum t of picture quality has been taken as ranging from 0 to 1, in the sense that 1 represents perfection.

Although quality rating is best considered in terms of a finite scale, this does not mean that a subject is unable to form a mental estimate of size or magnitude as such. The physical impairments in a picture which affect its quality are virtually unrestricted as to magnitude. Thus, in contrast to the bounded quality rating scale, the magnitude estimating scale must extend to infinity at one end, even although it is defined by zero at the other. It is what the psychologists term a 'ratio scale' because the magnitudes N along it have meaning only so far as their relative values are concerned, and, in the absence of an absolute reference in addition to zero, are defined arbitrarily.

5 Hypothesis

Now we come to the central theme of this paper. The viewer is aware of the infinite possible range of apparent impairment magnitudes and yet he somehow relates magnitude N to his opinion of quality t on a finite scale. How does he do it? No one knows, and it is not intended to discuss that aspect. Let us instead apply our intuition to suggesting the way in which he might be expected to function.

There is no problem in applying a simple linear law to apparent impairment magnitudes up to any chosen finite values. Magnitudes up to unity, for example, might be simply rated according to the law

$$t = 1 - N$$

illustrated by the dashed line of Fig. 1. It will be seen that the right-hand scale suggests a division of the t continuum into the 5 categories. The intervals are made equal so as to present a neat picture, but equality is not essential to the present discussion. All the intervals must, however, be finite as discussed in the previous Section.

Clearly the simple linear law fails as soon as N departs from the range 0–1. Let us therefore consider next the other extreme where the impairment magnitude effectively reaches infinity. Towards this limit, the intended scene would have become completely unrecognizable. One way of defining infinity is as the result of dividing zero into a finite number, and probably the simplest way of describing numbers ranging to infinity is by means of their reciprocals. In Fig. 2 the chain-dotted straight line illustrates the relationship:

$$t = 1/N.$$

Because of the intrinsic nature of the problem we cannot illustrate both laws in their entirety on the same simple graph. However, a portion of the reciprocal law is shown by the chain-dotted line in Fig. 1. The curves have been replotted on a scale of $\ln N$ in Fig. 3 where they each assume an exponential shape and will be seen to bear a skew-symmetrical relationship to one another.

Evidently what is needed is a composite curve providing a smooth transition between the dashed and chain-dotted curves. In Fig. 3 the slopes are equal at $t = 0.5$, so we could slide one curve to fit the other at this level. However a discontinuity of the second derivative would remain; such an instantaneous transition from a linear law to a reciprocal law seems most unlikely.

The equation for the dashed curve may be written

$$N = 1 - t \quad (1)$$

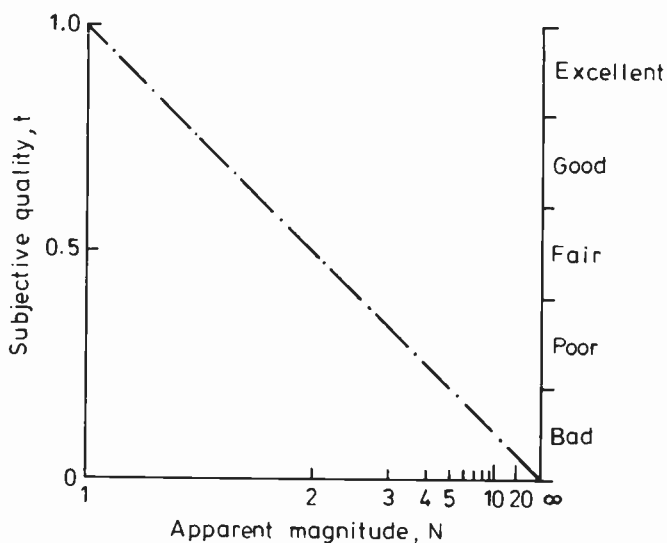


Fig. 2. Quality with large impairment magnitude.

and that for the chain-dotted curve

$$N = 1/t \tag{2}$$

It may be noted that when t is very small, and equation (2) might be assumed to apply, N in equation (1) is close to 1. Conversely when t is close to 1, and equation (1) might be assumed to apply, N in equation (2) is close to 1. It now becomes obvious that a simple way of obtaining a suitable composite equation is merely to take the product of the two functions of t , giving $N = (1 - t)/t$. This equation can be re-written

$$t = 1/(1 + N). \tag{3}$$

Equation (3) is illustrated by the full-line curves of Figs 1 and 3. The smooth transition afforded between the linear and reciprocal laws is evident, particularly in Fig. 3. The skew-symmetrical sigmoid form of the relationship there is described by the mathematical function known as the logistic. The logistic function is valuable in a number of applications and has the merit of being as simple as one could hope to find for the purpose. Despite encouragement offered by those who have tried it, the relationship is not as widely known as it deserves.

Of course, it is not suggested that a viewer consciously employs the mathematical concepts discussed here, simple as they are. The formulae are put forward merely as a possible description of the course of opinion formation, viewed from without. Although in Section 4 the N scale was introduced as a ratio scale without absolute significance, it should now be noted that N has been solidly related to t by equation (3), and thus endowed with the same absolute properties possessed by the t scale. See, however, Sections 3 and 8 as regards the standardization of subjective test arrangements.

6 Evidence

As its title implies, the previous Section is merely a speculative hypothesis. Historically however it has been arrived at comparatively recently, as a rationalization of empirical results. It is placed early in this paper in the hope that it may prove intuitively attractive at least to some readers, and will provide a useful basis for a practical understanding of what follows.

Broad support for the proposed relationship between apparent magnitude and qualitative scaling can be found in the psychometric literature concerning responses to various kinds of stimuli.⁴ Stevens⁵ has demonstrated that, for a given stimulus, apparent magnitude is usually related to physical stimulus magnitude by a simple power law. For equation (3) to apply, the rating on the t continuum must therefore be related to physical stimulus magnitude by a logistic of form

$$t = 1/(1 + |d/d_M|^G). \tag{4}$$

Inspection will show that this is just a more general form of the logistic given by equation (3), in which N is replaced by d , normalized in terms of d_M and raised to the power G .

Unfortunately, psychologists have seldom expressed their results in terms of logistic parameters so an amount of re-analysis is necessary. Also, for one reason or another, the experiments that have been made often prove to be not completely suitable. However, taking the whole of the evidence, equation (3) is found to apply quite closely provided N is raised to an appropriate power k , always within the narrow range 0.7–1.5, close to unity.

For evidence specific to the 5-grade quality scale it is necessary to turn to the results obtained with television. Here, very few results have been measured in terms of apparent magnitude, presumably because they would have had no practical applications. Hence a direct check on the value of k is not possible. However, there is a valuable alternative line of attack. In the application of results there has always been a great interest in the possibility of a law which would describe the way in which the effects of a number of different impairments add. Because there are potentially so many kinds of impairment, it is generally possible only to test each individual impairment in isolation. Certainly it is impracticable to make subjective assessments of all the possible combinations that can arise. If the quantity N is a measure of apparent magnitude, as described in the previous section, could it perhaps also be the summable quantity we require? The available evidence has recently been listed elsewhere.³ It is not as plentiful as one would wish but it is consistent in supporting the conclusion that apparent magnitude N is virtually identical to summable magnitude J .

A good illustration is provided by data obtained by the British Broadcasting Corporation using impairments

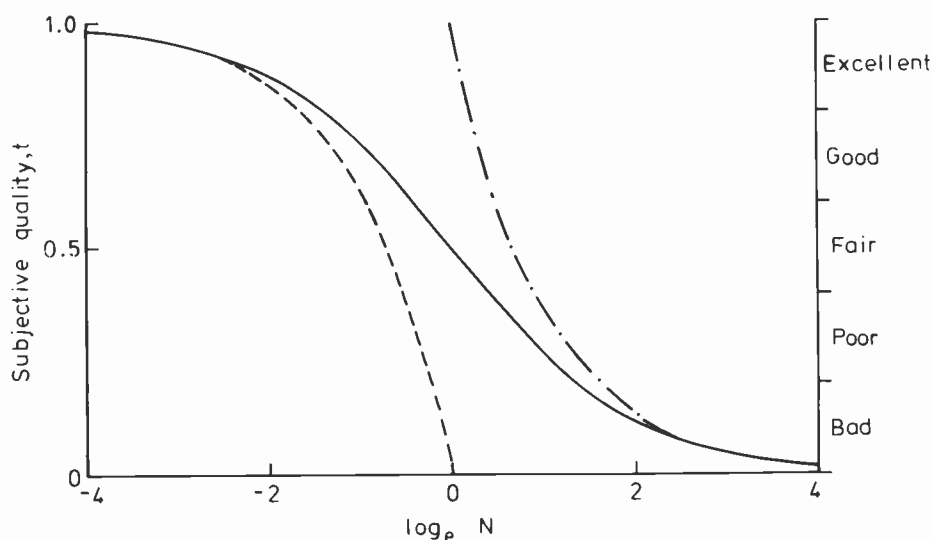


Fig. 3. Quality related to logarithmic magnitude.

comprising random noise, long-delayed echo, differential gain, and differential phase.⁶ In each of 2 experiments, the magnitude of each of the 4 different kinds of impairment was held constant, but every possible combination of up to 4 different impairments was assessed. Figure 4 shows the averaged values of apparent magnitude $N = J$ as measured, in comparison with the straight lines corresponding to simple summation. It will be seen that the fit is good for both experiments and gives no evidence that other than such a simple law is necessary.

The relationship between picture quality on the subjective continuum t , and the objective magnitude d of the impairment measured in the signal is known^{4, 7} to take the form of equation (4). d must, of course, be expressed in appropriate arithmetic units. Comparison with equation (3) indicates that $|d/d_M|^G$ is evidently the formula for the apparent magnitude N (= summable subjective magnitude J). The fact that such a simple relationship applies need not be surprising. G must be the effective power law for the transfer characteristic between the television signal and the impression given by the display; after all we have the example of luminance which is related to instantaneous signal voltage via luminance and the power law given by the gamma of the cathode-ray tube. It is, however sometimes necessary to spend a little time searching for a suitable objective measure of an impairment that can be simply related to its subjective effect.

Figure 5 is an illustration of the efficacy of the logistic psychometric function in combination with the summable property of apparent magnitude N . It shows results for a two-dimensional subjective experiment in which 6 levels of wideband random noise and 6 levels of a long-delayed echo were assessed in all possible combinations. Curves of constant noise or echo

magnitude are shown on a surface given by equation (3) in which

$$N = J = 0.177 + (d_n/0.0607)^{2.25} + (d_e/0.131)^{1.69}$$

where d_n = proportionate noise magnitude and d_e = proportionate echo magnitude.

In this expression, the first term is the apparent magnitude of the residual impairment due to imperfections in the original scene, and to those, such as line structure and flicker, caused by basic technical imperfections of the television system. The second term is the apparent magnitude of the impairment due to noise, and the third that due to echo. In plotting the surface, values of t were transformed to values of mean score \bar{u} (see Sect. 7) so that direct comparison can be made with the original data points illustrated by the small circles.

Where impairments produce effects which are sufficiently similar for their addition to be directly calculable in objective terms, e.g. the case of differential gain and differential phase (in the PAL system), it may well be that simple addition along the N scale is clearly incorrect. The exact degree of similarity within which it would be inappropriate to use N -scale addition is difficult to define. In the case of two identical sources of random noise, for example, because G in equation (4) is close to 2, N -scale addition proves sufficiently close to true power addition for practical purposes.

A topic that often provokes inconclusive debate is the so-called masking of one impairment by another. The full-line curve of Fig. 1 throws light on the matter. Because of the monotonic variation in slope it is evident that the addition of an impairment of apparent magnitude $N = 0.1$, say, will lower t by 0.09 in the absence of other impairments, but only by 0.02 if an unrelated impairment with $N = 1$ is already present. This difference could well be thought of as the masking of one

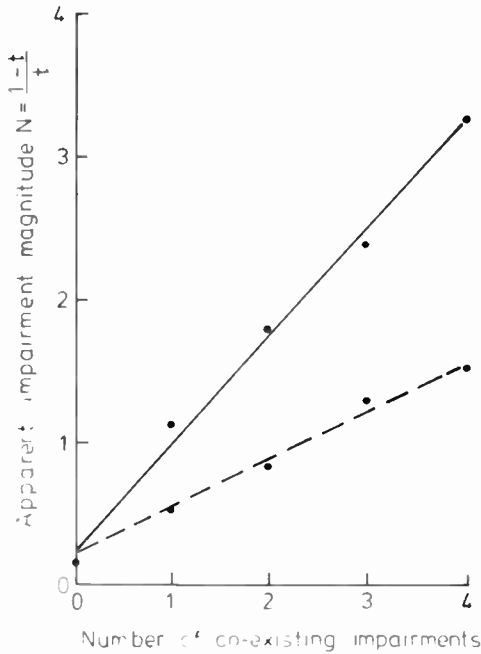


Fig. 4. Average effect of different numbers of co-existing impairments.

impairment by the other. If the two impairments comprise noise of similar spectral distributions, the masking is more obvious in that it is impossible to distinguish between the effects of the two sources of impairment. As previously noted, however, the effects on total quality are much the same. The important point is that, whether or not masking exists in a qualitative sense, the introduction of an additional impairment will almost always reduce the picture quality t in some degree.

The general relationships between the various quantities discussed in Sections 4–6 are to be found listed together with other related formulae in Appendix 12.1 of Reference 7. The logistic parameters G_i and d_{M_i} are here qualified by the suffix i to indicate that the relationships are based on opinions on the t continuum, rather than the quantized results obtained with a 5-grade rating scale discussed in the next Section.

7 Mean Score

A vital topic in the field of opinion rating is the variability of opinion standards that occurs both between and within subjects. A discussion of the matter, including descriptions of opinion distribution models, is to be found elsewhere^{8–10} and there is space here only to review the salient points.

Frequently it is necessary to define a value on the picture-quality continuum as an average measure of a population of opinions. It is convenient to make use of the median value here termed t_m . This is the quantity employed to calculate N in Fig. 4. The median has the advantage that where transformations of scale are concerned, it can be directly translated from one scale to another. Because at this stage we are dealing with a

mathematical model of a large population, use of the median does not imply any loss of accuracy. However, as noted in Section 4, opinions are usually obtained in terms of categories. It is possible to deduce the corresponding value of t_m from a given distribution of opinions in terms of categories, but the process adds a slight complication and in any case has only fairly recently become understood.¹⁰

The median would be inappropriate to describe the trend in terms of a category scale, because it would lead to a discontinuous result. It is expedient here to use a mean value. For the 5-grade quality scale, by allocating a score u of 1 to excellent, 0 to bad, and equally spaced intermediate values to the other grades, it is possible to arrange for mean score \bar{u} to equal t_m at values of 0, 0.5 and 1. The complete relationship between \bar{u} and t_m for a typical opinion distribution is shown by the full-line curve of Fig. 6. The values will be seen sometimes to diverge noticeably from the equality condition indicated by the dashed-line curve. Nevertheless, partly because of the nearly linear relationship over most of the range, it has been found that the hypothesis of Section 5 and the logistic psychophysical relationship can both be verified taking \bar{u} as an approximation to t_m . It is only by gathering evidence from several experiments that support for the hypothesis in terms of t_m , rather than \bar{u} , can be gained.

Equally, the quantity $I = 1/\bar{u} - 1$ is used in practice to provide the scale of subjective impairment magnitude, in terms of units termed imps, along which unrelated impairments may be simply summed. Impairment magnitudes on a geometric series of $\frac{1}{8}$, $\frac{1}{4}$, $\frac{1}{2}$ and 1 imp provide a series of mark points in the range of most frequent interest, at which to quote corresponding

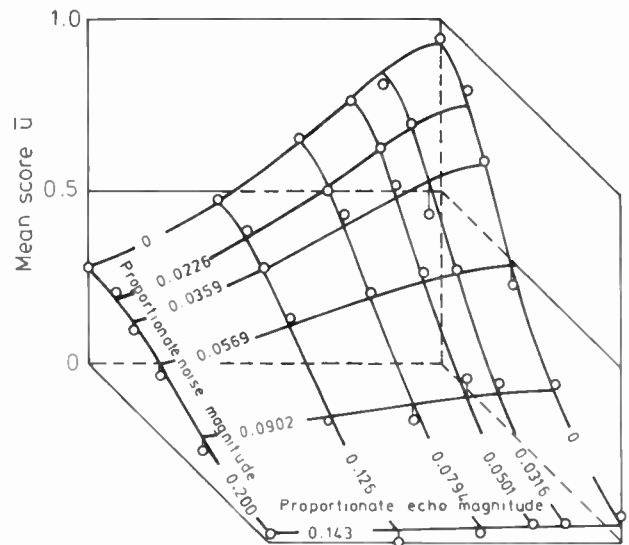


Fig. 5. Combined effect of random noise and long-delayed echo in television. Values of t calculated according to eqn. (3) and converted to mean score. — Curve of constant noise or echo magnitude on surface representing mean score. 0 Data point.

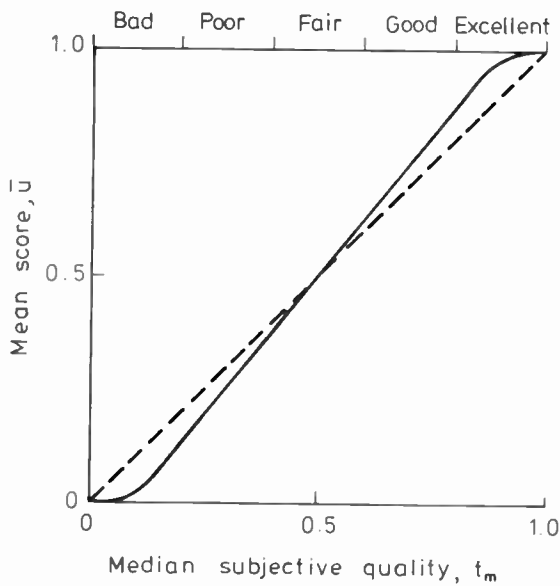


Fig. 6. Relationship between median quality and mean score.

objective magnitudes. Where the objective values are expressed in decibels the series for a given kind of impairment is, of course, equally spaced.

A full summary, based on mean score, of all the results for broadcast television so far obtained by the British Post Office is to be found in Reference 3. In view of the recent date of this publication, the results will not be repeated here. At present, it is not necessary to disturb the established method of subjective testing and analysis of results in terms of categories. Where desirable it is relatively easy to transform results into terms of the t scale, as in Reference 7. There are, however, indications that a more general move towards use of the t scale may become desirable in the foreseeable future.

8 The Future

Considerable progress has been made towards international agreement on the methods of carrying out subjective tests of television, and of analysing and expressing the results. The CCIR list of standardized test conditions is now virtually complete. Thus the time can be foreseen when strictly comparable results should begin to emerge from the different laboratories, with a resulting increase both in the volume of results available and the confidence with which they can be used in determining performance requirements.

As regards details of test procedure apart from actual display conditions, CCIR Recommendation 500¹ is as yet still imprecise. For example, although the number of different grading scales has been reduced to three, only rather obscure guidance is given on the choice among these 'according to the nature of the problem'. Opinion, as for example expressed at the recent Conference on Television Measurements (London, May 1979), is fairly unanimous that picture quality is usually the criterion of

ultimate interest. Therefore preference should be given to the scale in which the various levels of quality are categorized. There should be some clear advantage attached to use of another scale, and even then the problem of interpreting results in terms of quality will have to be faced. The main advantage of the CCIR impairment scale, which takes annoyance as the criterion, is that it can be employed by experts to rate the magnitude of a given type of impairment in the presence of other types. Thus it is well suited to operational assessments on working systems; the fact that Recommendation 500 embraces both this type of assessment and laboratory studies is a possible source of confusion.

Since the early 1960s, the British Post Office has been using the quality scale in conjunction with a rigorously-controlled laboratory test procedure designed to preserve as far as practicable the conditions of practical use of the television system.⁸ It has proved possible to demonstrate that subjective results can be repeated over intervals as long as 15 years with a shift of absolute assessment standard of about only 1 dB, expressed in terms of the objective magnitude of an impairment. In view of the fact that different teams of experiments were involved and that, in one instance, the observers were drawn from a different geographical area, the demonstration of an absolute standard with such a high degree of stability is highly encouraging. Comments on the method have been included in Section 1 of Annex IV of the latest issue of CCIR Report 405.¹¹

So far however it has rarely been possible to make satisfactory comparisons between results obtained by different laboratories. Probably the most ambitious exercise of this kind was attempted in 1977 by the European Broadcasting Union. The test procedure, which employed the impairment scale, was that referred to in Section 2 of Annex IV of CCIR Report 405.¹¹ Seven organizations made independent investigations of the effect of luminance-signal bandwidth. The results were disappointing, indicating ratios between bandwidths ranging up to 2:1 for a specified magnitude of subjective impairment. Expressed in terms of a corresponding impairment caused by random noise, the range is about 10 dB. Further investigation of the sources of this variation is highly desirable but none appears to be immediately proposed.

The subjective scale of picture quality can give us a yardstick. It is a yardstick that has to be established in the laboratory because it is possible to do so there. However, the measurements made with it have to be applied to the conditions of actual use of a system, and there the yardstick is likely to have altered. In domestic television, for instance, the interest value of the programme diverts attention from impairments in a way which it is impracticable to simulate in laboratory tests. The allowance to be made for such effects when applying

results is a matter not yet properly understood. It will be difficult to make useful progress in this matter whilst doubts remain about the comparability of standards for different laboratories, as discussed in the previous paragraph.

A number of relatively new problems have yet still to be resolved. Possibly the most important of these is concerned with impairments that are intrinsically limited in their magnitude. Hitherto methodical subjective testing has generally been directed to impairments arising in the analogue transmission of established broadcast television systems. Although much remains to be done in this area, it is evident that in the immediate future considerable attention is likely to be diverted, firstly to novel transmission methods involving digital coding, and secondly to new visual systems not necessarily using the broadcast standard. Impairments arising in basic coding systems are necessarily of small magnitude, and the mechanisms which cause them do not always lend themselves naturally to the production of large magnitudes. In such circumstances the full-range rating method cannot be directly applied. It is possible to 'embed' the test condition in a full-range test made with an artificially-generated matching impairment.¹² Although the results then obtained are based on the normal full-range standard, the technique suffers from the disadvantage that only a small fraction of the opinions cast in the test are actually used. Most of the other techniques being considered involve an element of comparison to avoid the confused results that occur when impairment magnitudes are limited to very small magnitudes.¹³ When we come to record the small differences of opinion that can be expressed if comparisons are allowed, it becomes evident that a continuous rating scale is desirable to avoid the quantizing error that is inevitable using a grading scale.

A closely related problem will arise with improved television systems of the future. The loss of information due to using a 5-grade scale in a full-range experiment with television of current broadcast standard is small in relation to the sampling error. Were this not the case, the use of such scales, would have been abandoned. However, the removal of unwanted effects due to line structure and flicker will lead, in more advanced systems, to essential results corresponding to t values close to unity. Here the fundamental sampling error will be much reduced, because of the narrowing of opinion distributions caused by the proximity of the scale boundary. By comparison, the quantizing error will be unacceptable. Like the problem of assessing codecs, this one also leads to the desirability of recording opinions on a continuous scale directly. Using such a technique, it has been found possible to preserve a relationship with 5-grade scale results by providing for each picture condition a simple scale resembling the right-hand abscissa of Figs. 1-3, to be marked by the viewer.

Among the other ideas presented in this paper, it is hoped that the concept of the t scale as the basic continuum on which opinions are formed, and then quantized into grades where necessary, will prove reasonably simple to understand and accept. Physical and mathematical representations also of the scale are essential to achieve the greatest possible accuracy in the collection, analysis and application of subjective results. However, the methods we select must not only be adequate for their purpose, they must also be sufficiently generally understandable, and lead to results which can be comprehended by the engineers who have to apply them. Results in terms of five quality grades³ have the appeal of simplicity, and are slowly becoming universally accepted. In this situation, it is probably unwise to change to the alternative method of presentation in terms of a continuous scale, except in circumstances where the accuracy provided by a grading scale is clearly inadequate.

9 Conclusion

The apparent magnitude of an impairment in a picture is also the quantity that can be simply summed to give the total apparent magnitude when a number of different impairments are simultaneously present. For a given impairment, apparent magnitude, or summable magnitude, is usually related to objective magnitude in the signal by a simple power law, provided the appropriate arithmetic measure of the impairment is chosen. By using a decibel or other logarithmic scale for objective magnitude, it follows that a direct relationship can be established to the logarithm of apparent magnitude.

However the scale of apparent magnitude is not the one on which viewers primarily judge an impairment. In this paper, a speculative explanation has been offered of the way in which we rate impairments according to apparent magnitude. It is hoped the description will assist in the interpretation of the results of visual subjective studies, and their application in terms of objective measurements.

No suggestions are offered as to the physiological mechanism of opinion formation which might contain the analogue of the t scale. No doubt one day it will become possible to make exact comparisons between the actual process and the empirical models with which we have at present still to be content.

10 Acknowledgment

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Biographies of **Professor G. D. Papadopoulos** and **Mr K. Z. Pekmestzi** were published in the November issue of *The Radio and Electronic Engineer*.

A simple method for swept frequency response testing

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SUMMARY

This paper deals with the design of simple, and relatively inexpensive, dedicated equipment which is tailored to test a 6th-order analogue pitch rate compensation unit using sweep frequencies and return signal processing to estimate gain and phase characteristics. It is shown that the test time required for an optimum swept frequency test compares very favourably with the total test time needed using either automatic test equipment or manual steady state testing at a few frequencies. The test technique provides a 'hard-copy' signature of the system under test in a form which has found particular favour for maintenance purposes in such diversified areas as electronic circuits, aircraft structures, and air breathing systems. The technique should be regarded as complementary to more flexible frequency domain test methods in forming the complete inventory of equipment needed during system life-cycle testing.

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Nomenclature and Abbreviations

a	peak value of a sinewave
f	frequency (hertz)
f_0	resonant frequency (hertz)
$\theta_1(t)$	input signal
$\theta_0(t)$	output signal
S	frequency sweep speed
f_H	maximum frequency to be swept
f_L	minimum frequency to be swept
K	constant
T	fundamental period
Δf	half power bandwidth
τ	delay
R	logarithmic sweep rate
v.f.c.	voltage to frequency converter
p.r.c.u.	pitch rate compensation unit
F.r.a.	Fourier response analyser
s.u.t.	system under test

1 Introduction

It is necessary over the lifetime of a system periodically to check whether its response falls within tolerance bands stipulated by the manufacturer. Frequency domain testing has enjoyed vast popularity for this purpose with a wide variety of reported applications ranging from jet engines, servo actuators, chemical processes, and compensation networks, to the alignment of intermediate frequency circuits.¹ Reasons for the popularity of frequency domain testing include;

- (i) Test results are simple to interpret
- (ii) Test results are in a form suitable for easy cross-reference with design data
- (iii) Sub-assembly test results can be readily combined together to estimate the overall response of a linear system
- (iv) Sinusoidal stimuli are readily used with correlation techniques to minimize the effect of unwanted noise on measurements²
- (v) If needed, time domain performance data can be extracted from frequency domain test measurements.³

A list of modern methods of determining frequency response is given by Towill and Payne.⁴ Available techniques can be broadly grouped together in three categories:

- (a) steady state measurements
- (b) sweep frequency measurements
- (c) Fourier transformation techniques.

If a system is linear and time invariant, its measured frequency response, in the absence of noise, is independent of the testing technique used. The test time, however, will be dependent on the testing technique. A popular instrument employed for measurements in

categories (a) and (b) above is the Fourier response analyser (F.r.a.).⁵ This instrument is based on the cross-correlation principle⁶ and is available in many versions ranging from a simple analogue device costing a few thousand pounds to a sophisticated computer-controlled test set some ten times more expensive.⁷ Special instruments are also available for frequency response estimation via Fourier transformation techniques,⁸ and it is also becoming common to use a general-purpose digital computer, with appropriate interfaces, as part of an automatic test set incorporating fast Fourier transform software.⁹

2 Contribution of the Present Paper

For workshop, depot, and on-site maintenance testing, the slow sweep frequency technique remains popular,¹⁰ especially where a hard-copy signature is required. Depending on the maintenance logistics adopted, a F.r.a. may be used, or a simple dedicated tester developed to suit a range of systems. The latter strategy may well prove the more economic policy if the system user wishes to provide many cheap distributed test facilities rather than a few centralized automatic test sets.

It is important to emphasize that dedicated swept frequency testers should be seen as satisfying a particular need within an overall dynamic test policy, integrating, rather than competing with, more sophisticated instrumentation which might be used during the manufacture, overhaul, and refit of the system.¹¹

In this paper dedicated test equipment, based on the slow sweep frequency test concept, is tailored to estimate accurately and (relatively) rapidly the frequency response, in amplitude ratio and phase angle form, of an analogue device known as a pitch rate compensation unit (p.r.c.u.) which is a module commonly found in aerospace stability augmentation systems. The technique is not, of course, restricted in application to this device, but the satisfactory testing of such a system in an industrial environment is thought to be a reasonable target for evaluation of the equipment. There is some

originality in the circuit design and development, but the real contribution of the paper is the engineering of the tester and subsequent matching to the p.r.c.u. Satisfactory gain and phase estimates are obtained even in the presence of modest amounts of measurement noise.

3 The Slow Sweep Frequency Technique

A swept frequency measurement is one where the return signal from the s.u.t. is monitored while, at the same time, the frequency of the sinusoidal forcing function is continuously changed until the frequency band of interest has been covered.^{12,13} In the slow sweep frequency test the sweep speed is sufficiently slow so that the response is approximately the same as that which would be measured under steady state conditions. With such a stimulus the frequency response can be evaluated and displayed whilst the test is in progress. Hence a continuum of data is provided rather than a discrete measurement set, which is obtained in a steady state spot frequency test. In the simplest form the data can be a pen record of the output of the system under test. An example of an existing slow sweep frequency test applied to the p.r.c.u. is shown in Fig. 1.

A more sophisticated test will not only define the frequency range to be swept but will return the frequency to its initial value at the end of the sweep. In analogue and some hybrid instruments the nominal frequency is controlled by a d.c. voltage (via a voltage to frequency converter), and this voltage can also be used to control the horizontal deflection of the output display unit. The vertical signal corresponds to the parameter being measured (magnitude ratio or phase). Usually the display is an oscilloscope or X-Y plotter. A block diagram illustrating the basic technique is shown in Fig. 2.

4 Estimating Sweep Times

The speed of sweep is important because if it is too fast the measured maximum value of the response will be significantly less than the corresponding true maximum

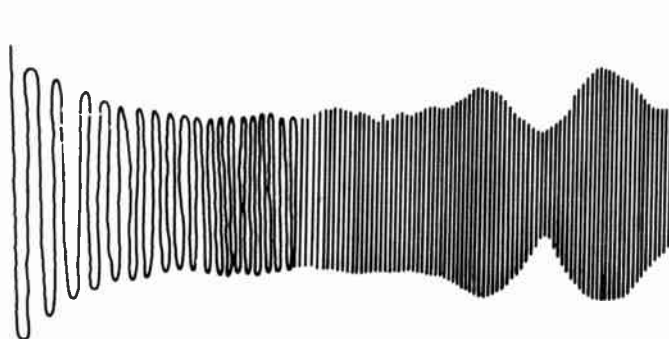


Fig. 1. Pen record of s.u.t. sweep frequency response as stored for 'life history' performance profiles.

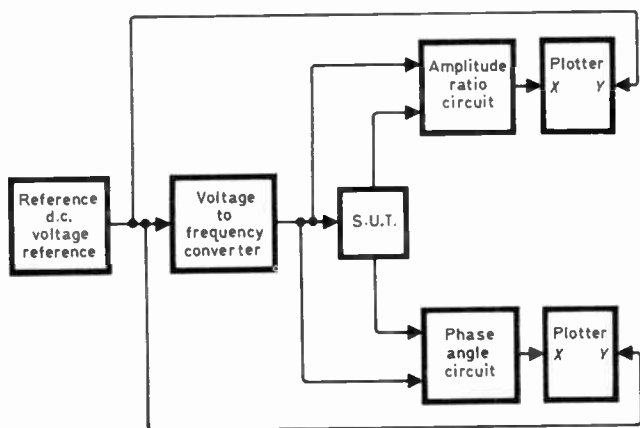


Fig. 2. Block diagram of the proposed sweep frequency test.

measured under steady state conditions and the resonant frequency will be shifted in the direction in which the frequency is changing,¹⁴ as illustrated in Fig. 3. An estimate of the required sweep time can be derived using the following equations:¹⁵

Let f_0 = each resonant frequency in turn, then

$$Q = \frac{f_0}{\Delta f}$$

where Δf is the half-power bandwidth of each resonance in turn.

If $Q < 10$ then the equations of Appendix 1¹⁵ are amended to:

(i) for a linear sweep function

$$C_v = \frac{Q}{f_0} \tag{1}$$

and sweep time

$$T = 10C_v \frac{f_H}{f_L} \tag{2}$$

(ii) for a logarithmic sweep function

$$C_x = \frac{Q}{\sqrt{f_0}} \tag{3}$$

and sweep time

$$T = \frac{10C_x}{f_L} \ln \left(\frac{f_H}{f_L} \right) \tag{4}$$

where

f_H = maximum frequency

f_L = minimum frequency.

As will be seen later, these equations give reasonable guidelines for a target test time even though the resonances of the p.r.c.u. are more heavily damped than those considered in the reference. Following the

application of equation (2) and (4) the test time is further refined during the development phase of the s.u.t.

The approach used in reference 15 is to concentrate attention on the s.u.t. resonances, so as to highlight critical regions in the determination of suitable sweep times. As the stimulus varies with time we define

$$\theta_i(t) = X_0 \sin(f(t) \cdot t) \tag{5}$$

where X_0 is a constant and $f(t)$ is a variable.

For a linear sweep,

$$f(t) = (a + bt) \tag{6}$$

and for a logarithmic sweep,

$$\frac{df(t)}{dt} = K \cdot f(t) \tag{7}$$

where K is a constant. It can be seen that for the latter sweep function the rate of change of excitation frequency is proportional to the instantaneous frequency. Which of these patterns is best in a given situation is determined by using equations (1) and (3) for each resonance in turn in order to ascertain whether a particular sweep function could be adopted over the whole frequency spectrum. For the adoption of a single linear sweep function all the values of C_v would, ideally, be equal. Similarly if all the values of C_x were equal then a single logarithmic sweep function could be employed.

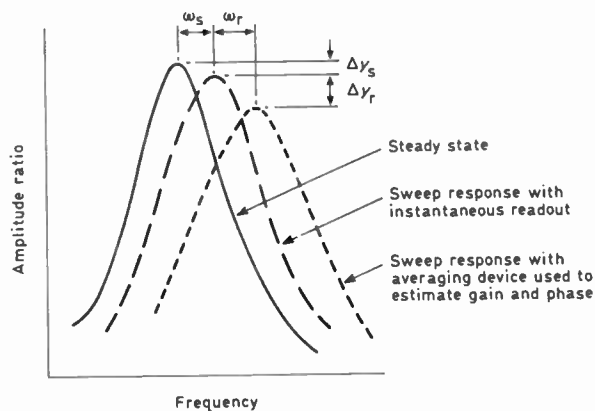


Fig. 3. Effect of the sweep frequency measurement technique on resonance estimates.

The practical maximum allowable frequency sweep speed is not only a function of the physical characteristics of the s.u.t. as implied in equations (2) and (4), but is also dependent on the dynamics of the measuring system.

Two types of detector circuits are generally considered:

- (i) circuits which measure the instantaneous peak signal value
- (ii) measurements involving an averaging process.

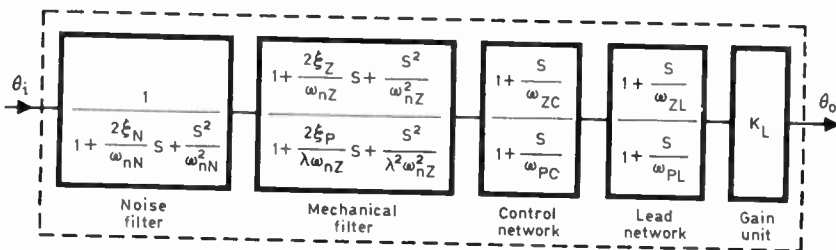


Fig. 4. Details of a typical hardware s.u.t.

Block diagram of p.r.c.u.

Nominal parameter values for the p.r.c.u.

Parameter	ζ_N	ω_{nN}	ζ_z	ω_{nz}	ζ_p	λ
Nominal value	0.78	125.5	0.25	30	0.57	0.883
Parameter	ω_{zc}	ω_{pc}	ω_{zL}	ω_{pL}	K_L	
Nominal value	4.0	1.0	6.5	24	1.0	

(All frequencies measured in rad s^{-1})

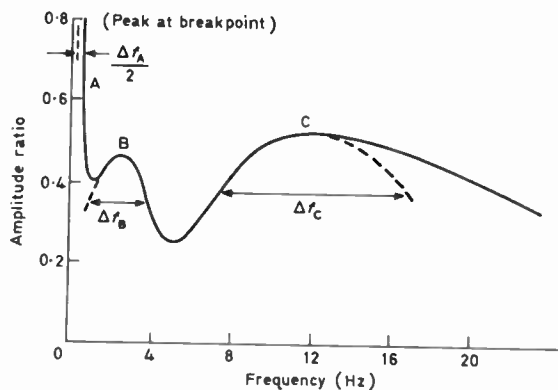


Fig. 5. Approximating the p.r.c.u. frequency response by resonances A, B and C.

Measurement of instantaneous value is the faster form of signal level detection and does not impose extra restrictions upon the maximum allowable sweep speed. It is, however, expected to be inaccurate in high noise environment unless a tracking filter is used. As field trials have shown that the hardware p.r.c.u. is linear and relatively noise-free the detector circuits used herein have been based on the measurement of the instantaneous signals. However as will be shown later, the detector circuits cope surprisingly well with modest noise levels.

The p.r.c.u. frequency response characteristic can thus be divided into two separate sections:

- (i) Resonance A, frequency range 0.16–0.7 Hz
- (ii) Resonances B and C, frequency range 0.7–28 Hz.

5 The Slow Sweep Frequency Technique Applied to the P.R.C.U.

A block diagram of the p.r.c.u. and nominal parameter values are shown in Fig. 4. To facilitate evaluation of the sweep frequency tester an analogue model utilizing 41 discrete components was designed and manufactured.¹⁶ Spot frequency testing confirmed that the analogue model met the p.r.c.u. performance specifications.

Figure 5 illustrates how the magnitude ratio frequency response characteristics can be approximated to three resonant peaks A, B and C. To implement equations (2) and (4) the bandwidth and resonant frequency of each approximated resonance is measured and values of Q , C_v and C_x are calculated. Table 1 summarizes the results and indicates that, for resonances B and C, the C_x values can be considered to be roughly equal. Hence a single logarithmic sweep function can be adopted to measure these two resonances. Using equations (2) and (4) approximate sweep times can be calculated. The results are illustrated in Table 2 and indicate that a linear sweep function is the optimum sweep mode for resonance A as a lower test time is indicated.

Table 1
Estimates of Q , C_v and C_x from the p.r.c.u. amplitude frequency response

Resonance	Resonant frequency Hz	Bandwidth Δf Hz	Q	C_v (linear)	C_x (logarithmic)
A	0.16	0.5	0.33	2.0	0.83
B	2.4	3.3	0.73	0.3	0.47
C	12.0	10.4	1.15	0.1	0.33

Table 2
Predicted sweep time for estimating the p.r.c.u. amplitude ratio frequency response

Sweep technique	Frequency range	Frequency sweep time (seconds)
logarithmic	0.16–0.7 Hz	assuming $C_x = 0.83$ $T = 76$ s
linear	0.16–0.7 Hz	assuming $C_v = 2$ $T = 67$ s
logarithmic	0.7–28 Hz	assuming $C_x = 0.47$ $T = 25$ s

Consequently a two-stage frequency generator has been designed, the first stage generating a linear sweep from 0.16 to 0.7 Hz and the second stage generating a logarithmic sweep from 0.7 to 28 Hz. As can be seen from Table 2, the provision of a linear-logarithmic sweep pattern will result in a total test time of the order of 92 seconds.

6 Sweep Frequency Instrumentation

In order to generate a sweep frequency the voltage to frequency converter (v.f.c.) must be stimulated by a voltage that varies with respect to time in correspondence with the desired function. This is achieved by using an integrator with a constant direct input voltage to generate the linear sweep and an integrator with positive feedback to generate the logarithmic sweep function,¹⁶ that is, the integrator is driven with a proportion of its instantaneous output. The sweep time is then controlled by the percentage 'positive' feedback and a d.c. offset sets the frequency at the start of the sweep. A comparator with an adjustable d.c. reference controls the reset frequency. An additional comparator with an adjustable reference provides a variable changeover from linear to

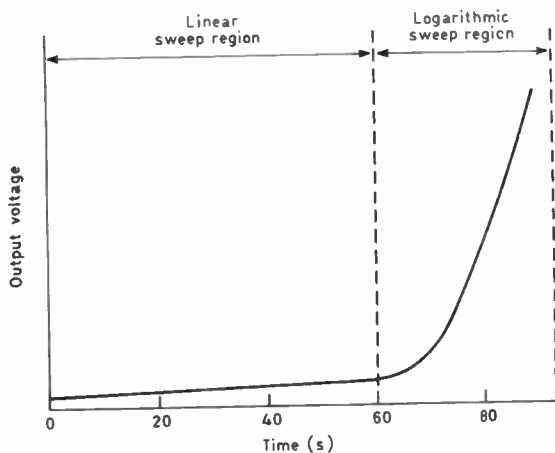


Fig. 6. Linear logarithmic voltage ramp produced by the signal generator.

logarithmic modes. The front panel of the function generator module provides RESET (frequency return to initial value), RUN and HOLD functions. Indicators give a visual display of the current operating mode.

Figure 6 illustrates the combined linear-logarithmic voltage function, which constitutes the v.f.c. forcing function. The sweep frequency output of the v.f.c. is used to drive the p.r.c.u.

Magnitude ratio and phase estimation circuits are based on the measurement of instantaneous values. The magnitude ratio circuit measures the peak voltage every cycle. When sampling is complete the measuring circuit is reset. Thus an ostensibly continuous plot is obtained which is automatically updated once every cycle. Phase estimation is based on the production of a constant voltage pulse the width of which is proportional to the phase difference. Integration of the voltage pulse produces a voltage magnitude proportional to phase. A voltage pulse, the width of which is proportional to a phase shift of 180° is also produced and a voltage magnitude proportional to this phase shift, at the instantaneous frequency, is obtained by the same integration process. Division of the two analogue voltages produces an estimate of phase which is independent of frequency. The block diagrams for the magnitude ratio and phase measuring circuits are illustrated in Figs. 7 and 8. Further details on design and calibration of these circuits can be found in reference 16.

7 Results

The integrator time-constant, the initial condition reference, the linear ramp reference, and the positive feedback control were adjusted to produce the sweep times calculated in Section 5. Then the equipment was 'fine tuned' experimentally to produce an optimum sweep time which in this case is somewhat less than that predicted in the Appendix; this is apparently due to the relatively low *Q* values of the p.r.c.u. resonances. Examples of typical responses are illustrated in Figs. 9 and 10. A response of the type shown in Fig. 9 is obtained when the sweep characteristic is not matched to the actual response of the s.u.t. whereas Fig. 10 shows

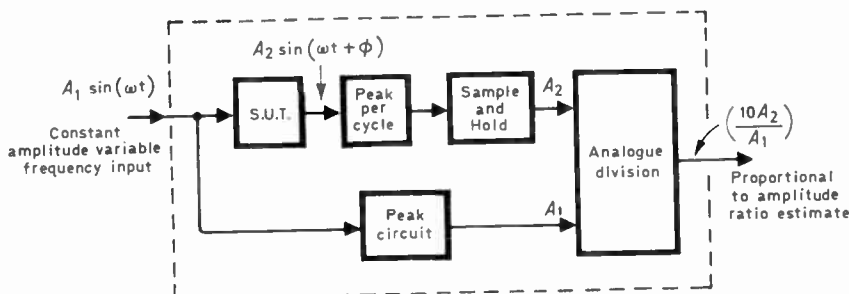


Fig. 7. Block diagram of the magnitude ratio measuring circuit.

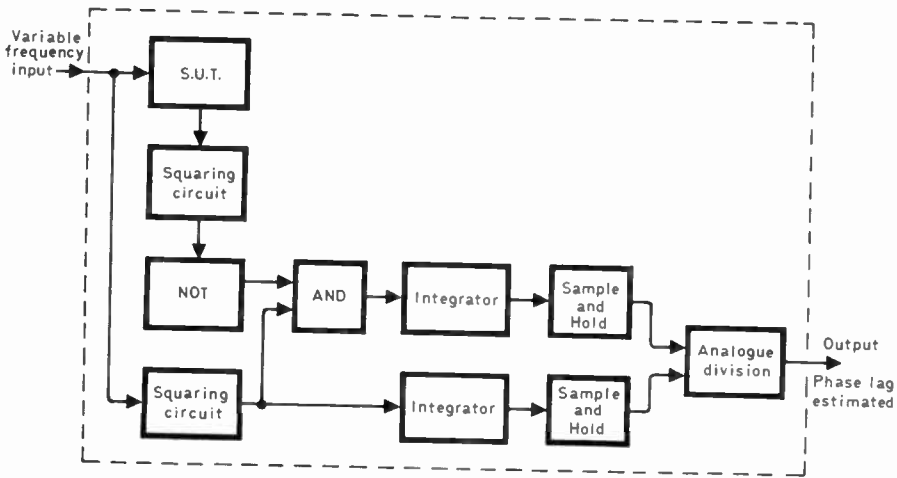


Fig. 8. Block diagram of the phase measuring circuit.

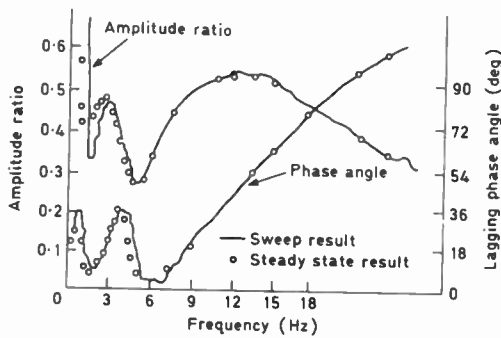


Fig. 9. Experimental results for linear sweep rate.

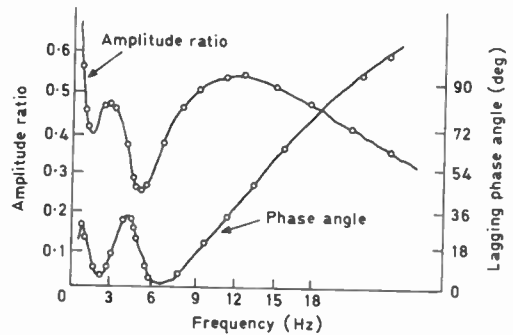


Fig. 10. Experimental results for linear logarithmic sweep rate.

the frequency response for the optimum sweep condition. The 'signature' thus obtained is perfectly adequate for p.r.c.u. maintenance purposes.

Although the detector circuits are based on the measurement of instantaneous values, the system will cope surprisingly well with reasonable noise levels. This

is particularly true of the phase measurement circuit. Figure 11 illustrates the magnitude and phase frequency responses obtained with a signal to noise ratio of 10 : 1. It can be seen from the waveforms that the noise produced an error in the form of an offset for the magnitude ratio result, but the estimates are within 7% of the nominal trace, whilst for phase estimates the results are within 1% of the noise free response. Whether or not such a 7% error is permissible in a given application

Table 3

Summary of p.r.c.u. testing via various frequency domain techniques

Test technique	Test time	Comments
Special sweep frequency tester	76 s	Linear-logarithmic sweep mode
Spot frequencies via manual mode F.r.a.	90 s	Using an optimum test set of five frequencies and integrating over one cycle
Automatic test equipment containing computer controlled F.r.a.	46 s	Recording and plotting 5 points per decade

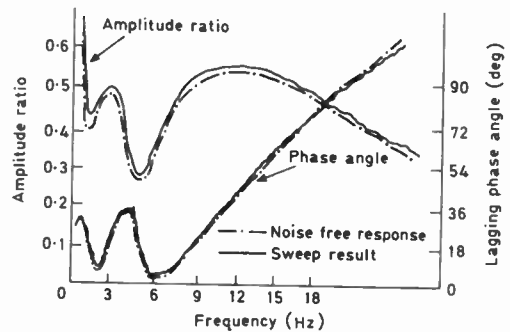


Fig. 11. Effect of noise on the sweep frequency response.

depends on the tolerance band set around the nominal response to take account of component variability. Since the p.r.c.u. user will make a judgment on the general pattern of the response, rather than an individual feature, this magnitude of error would be acceptable, although as previously pointed out, noise levels of this order are not anticipated in field testing of the device under analysis.

8 Conclusions

The test time for the optimum sweep frequency test using dedicated equipment is of the order of 76 seconds. It can be seen from Table 3 that this test time compares very favourably with the total test time required by the steady state spot frequency method and by automatic equipment. The advantages of the sweep frequency technique are:

- (i) The equipment is relatively inexpensive. For quantity production in a batch size typically ordered for maintenance depots, it might be as little as one third of the cost of a manual mode F.r.a.
- (ii) The test program and the recording of the results is automatic, requiring very little operator involvement, so is ideally suited to the maintenance environment.

On the other hand, the simple sweep frequency technique naturally does not possess the wide versatility of F.r.a. instrumentation which will continue to be necessary for specific systems. These include situations where a high noise level is encountered,² or where special facilities such as estimation of the higher harmonic content of the return signal is needed.¹⁷

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10 Appendix 1: Sweep Test Time Estimates

The following derivations follow those developed by J. T. Broch.¹⁵

Assuming that the frequency sweep function is linear in the region around the resonant peak then the input signal within the range may be written in the form:

$$\theta_i(t) = X_0 \sin(at^2 + bt) \tag{8}$$

where

$$2at + b = \omega t, \quad \frac{d\omega}{dt} = 2a$$

and

$$b = \omega_L = \text{the low frequency limit.}$$

For resonance with high Q values the unit impulse response is given by

$$h(t) = X_{\text{peak}} \exp\left(-\frac{\omega_0}{2Q} t\right) \sin(\omega_0 t + \phi) \tag{9}$$

where

$$\omega_0 = 2\pi f_0, \quad f_0 = \text{resonant frequency}$$

$$Q = \frac{f_0}{\Delta f},$$

Δf is the half power bandwidth of resonance.

Assuming $Q > 5$ the convolution can be mathematically formulated as:

$$\theta_0(t) = \int_{t_1}^t X_0 \sin(at^2 + bt) X_{\text{peak}} \times \exp\left[-\frac{\omega_0}{2Q}(t-\tau)\right] \sin(\phi + \omega_0(t-\tau)) d\tau \tag{10}$$

t_1 is the instant where

$$f_L = \frac{\omega_L}{2\pi}.$$

The equation can be transformed into an 'energy-envelope' relationship of the kind

$$\theta_0^2(t) = C \int_0^t \frac{\exp\left[-\frac{\omega_0}{Q}(t-\tau)\right]}{1+(\tau/\tau_0)^2} d(\tau/\tau_0) \quad (11)$$

where C is a normalizing constant and

$$\tau_0 = \left(\frac{\Delta f}{2S}\right).$$

This results in the error in maximum resonance

$$= F_1 \left(\frac{S}{\Delta f^2}\right).$$

To keep the 'error' in the order of a few tenths of a decibel

$$\frac{S}{(\Delta f)^2} \leq 1. \quad (12)$$

Now

$$Q = \frac{f_0}{\Delta f}$$

hence

$$\Delta f^2 = \frac{f_0^2}{Q^2}.$$

Therefore

$$S = \frac{df}{dt} \leq \frac{f_0^2}{Q^2}. \quad (13)$$

Three cases need to be considered depending on the assumption made concerning the behaviour of Q .

(a) When the resonances have Q values which are constant the maximum allowable sweep speed is:

$$\begin{aligned} \text{assuming } \frac{df}{dt} &= \frac{f_0^2}{Q^2} \\ \frac{df}{dt} &= \frac{1}{Q^2} f^2 \end{aligned} \quad (14)$$

Hence the period necessary for one sweep from frequency f_L to frequency f_H is:

$$\begin{aligned} T &= \int_{f_L}^{f_H} Q^2 \frac{1}{f^2} df \\ &= Q^2 \left[\frac{1}{f_L} - \frac{1}{f_H} \right] = \frac{Q^2}{f_L} \left[1 - \frac{f_L}{f_H} \right] \end{aligned} \quad (15)$$

It can be seen that as a result of this assumption the required sweep function is hyperbolic.

(b) When the resonances have Q values which increase with the square root of frequency ($Q = C_x \sqrt{f}$)

$$\begin{aligned} \frac{df}{dt} &= \frac{f^2}{Q^2} = \frac{1}{C_x^2} f \\ T &= \int_{f_L}^{f_H} C_x^2 \frac{df}{f} = C_x^2 L_n \left[\frac{f_H}{f_L} \right] \end{aligned} \quad (16)$$

As a result of this assumption the required sweep function is logarithmic.

(c) When the resonances have Q values which increase linearly with frequency ($Q = C_v f$)

$$\frac{df}{dt} = \frac{f^2}{Q^2} = \frac{1}{C_v^2}$$

Therefore

$$T = \int_{f_L}^{f_H} C_v^2 df = C_v^2 (f_H - f_L) \quad (17)$$

So on this assumption the required sweep function is linear.

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Digital transmission of video and audio signals over an optical-fibre system

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Based on a paper presented at the 7th International Broadcasting Convention (IBC 78) held in London on 25th to 29th September 1978.

SUMMARY

During 1977, BBC Research Department was invited to participate in field trials of an experimental optical-fibre digital transmission system installed between the Post Office telephone exchanges at Hitchin and Stevenage. BBC equipment was specially adapted for these trials to provide multiplexed digital video and audio signals at 140 Mbit/s and to decode these signals after transmission.

High-quality audio and video signals were sent between Hitchin and Stevenage, a total distance of some 18 km for the return path, over a period of one month during which time a number of tests were conducted. Brief descriptions of the equipment and tests are contained in this paper. No serious difficulties arose in the trials.

1 Introduction

Equipment for coding and multiplexing high-quality video and audio signals to form a 120 Mbit/s bit stream (comprising two bit-interleaved 60 Mbit/s bit streams) was developed at BBC Research Department¹ for use in field trials on a British Post Office experimental high-speed digital line system in 1975. The results of these trials² showed that this system was capable of carrying the video and audio signals with imperceptible levels of impairment, but that the timing jitter and error rate of the system were influenced by the analogue video waveform. Pattern-dependent errors and timing jitter, which arise when a transmission system is carrying the highly-structured bit-stream patterns related to the television scanning format, are not readily detectable in tests using fixed patterns and pseudo-random sequences. There are, therefore, valuable tests which can be conducted on certain digital transmission systems using 'real' digital video and audio multiplexed signals.

The experimental multiplexing equipment was then used to provide a 60 Mbit/s bit-stream for further tests in conjunction with the Post Office in 1976.³ On this occasion the transmission system was a 6 GHz radio link from the Post Office Satellite Earth Station at Goonhilly Downs to an *Intelsat IV* communications satellite in a geo-stationary orbit over the Indian Ocean. A 4 GHz radio link from the satellite back to Goonhilly completed the circuit. The performance of the digital satellite system was, under certain conditions, signal-dependent thus emphasizing again the value of tests using real video and audio digitally-coded signals.

In 1977, Standard Telephones and Cables installed an experimental 140 Mbit/s† optical-fibre system between the telephone exchanges at Hitchin and Stevenage, in collaboration with the Post Office. Following an invitation from STC, the opportunity was taken to carry out field trials on the optical-fibre system so building up further BBC experience on various types of digital transmission systems.

This paper gives a brief outline of the STC optical system and of the BBC coding and multiplexing equipment before describing the transmission tests.

2 The Optical System

A simplified block diagram of the complete optical-fibre transmission system⁴ is shown in Fig. 1. The optical fibres are installed in underground cable ducts between the two exchanges, which are 9 km apart, with repeaters every 3 km. For the tests involving multiplexed video and audio signals, it was convenient to locate the programme encoding and multiplexing equipment in the telephone exchange at Hitchin and to loop the 'forward' and 'return' paths of the system at Stevenage.

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† The precise bit rate is 139.264 Mbit/s corresponding to the standard fourth-order multiplex rate agreed for the European digital hierarchy.

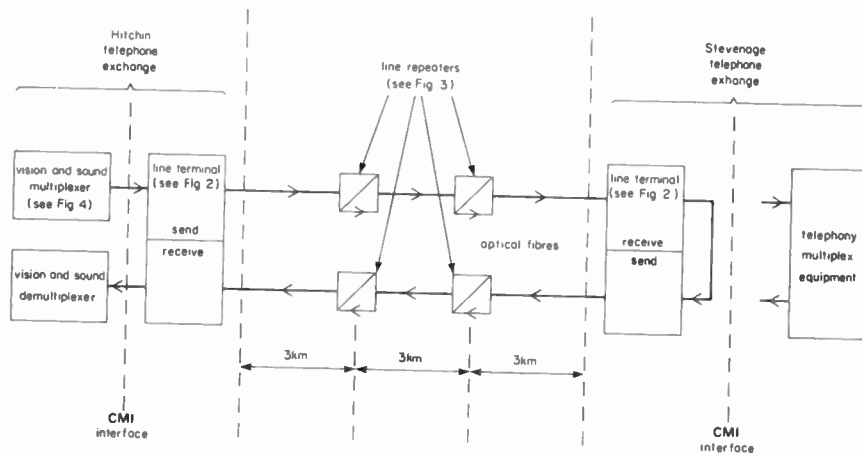


Fig. 1. The complete system.

Interfacing with the optical-fibre system follows standard practice for the bit-rate of 140 Mbit/s and is achieved via a coded mark inversion (CMI) interface code.⁵ The appropriate CMI-to-binary plus clock code converters are provided in the terminal equipment. A block diagram of a line send/receive terminal is shown in Fig. 2.

The requirements that the transmitted signal should have no d.c. component (so that a.c. coupling could be employed in the terminal equipment), that the signal should contain sufficient timing information, and that in-traffic error rate monitoring facilities would be possible led to the use of a special line code in the STC equipment. Binary signals to be transmitted over the optical system are first scrambled, and then a parity bit is inserted after every 17 data bits. The presence of the parity bit enables a check to be made on the error rate of the system, and at the same time the use of 'odd' parity ensures that at least one transition between '0' and '1' (or vice versa) occurs during each parity group of 18 bits (this means that no more than 34 consecutive '1's' or '0's' can occur in the bit stream). The bit-rate actually transmitted through the fibre is accordingly $139.264 \times 18/17 = 147.5$ Mbit/s. The wavelength of the

gallium-arsenide striped laser light source is $0.85 \mu\text{m}$ at which the fibre attenuation is about 4.7 dB/km. Pulse dispersion in the fibre is about 1.6 ns/km.

The repeaters which are installed in the system comprise an opto-electric detector, an equalizer amplifier with automatic gain control, a data regenerator and an electro-optical modulator. A block diagram is shown in simplified form in Fig. 3. Repeaters are used principally to remove noise and jitter from the signal by regenerating the data signal at intervals throughout the system and remodulating. By employing regeneration, the system error rate can be kept very low. Equalizers are included in the repeaters prior to the regenerators to correct for pulse dispersion introduced by the optical fibre.

Insulated copper conductors are included in the cable sheath with the optical fibres to provide engineering telephone facilities, alarm signalling and power supplies for the repeaters.^{4,6} The cable used between Hitchin and Stevenage comprises four optical fibres and four insulated copper wires surrounding a central steel strength member. A polyethylene jacket provides the outer cover and the overall diameter of the cable is 7 mm.

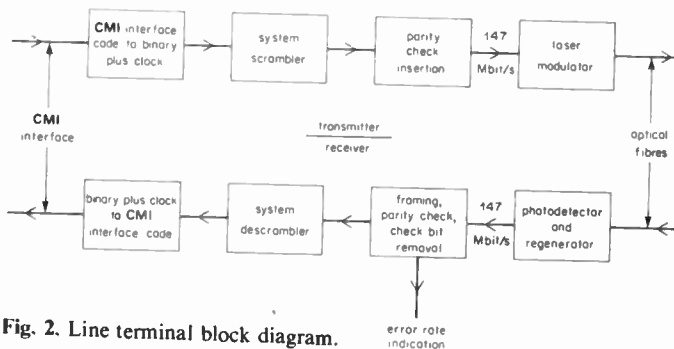


Fig. 2. Line terminal block diagram.

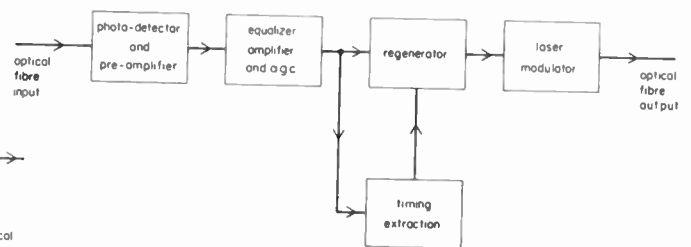


Fig. 3. Line repeater block diagram.

3 The BBC Experimental 140 Mbit/s Coding and Multiplexing Equipment

The 140 Mbit/s multiplexing equipment is basically that developed for use at 120 Mbit/s for the field trials of a Post Office digital line system or at 60 Mbit/s for tests via a satellite link.¹ Modifications, involving principally an increase in the clock rate and the provision of 'dummy' extra pulses, have been made so that the equipment operates at the higher bit rate of 139.264 Mbit/s. One 'dummy' extra pulse is provided for every 6-bit video sample word, but no additional information is conveyed by these extra pulses.

Channels are provided for two high-quality System I PAL colour television signals and twelve high-quality sound signals. A block diagram of the sending equipment is shown in Fig. 4; the processes shown are reversed in the receiving equipment. For the tests on the optical-fibre system, video channel A could be provided with off-air television signals or locally-generated television test patterns, whilst channel B carried a black level, syncs and colour burst signal which was generated in digital form. Only one set of six sound-programme channels was equipped.

3.1 Video Coding, Scrambling and Error Protection

The System I PAL colour television signal to be sent in channel A of the multiplex is coded into an 8-bit parallel p.c.m. (pulse-code modulation) signal sampled at the sub-Nyquist rate of exactly twice the colour subcarrier frequency.⁷ Code conversion facilities are provided so that the p.c.m. signal can be converted into 6-bit d.p.c.m. (differential pulse-code modulation) or 6-bit h.d.p.c.m. (hybrid differential pulse-code modulation)† giving

substantially the same picture-quality as the 8-bit p.c.m. signal, for transmission in the multiplex 'package'. As an alternative, the six most significant bits of the 8-bit p.c.m. signal can be used as a 6-bit p.c.m. signal, although this does not give such good picture quality as 6-bit d.p.c.m. or h.d.p.c.m. To minimize pattern-dependent effects, each video bit-stream is provided with an optional video scrambler to modify the patterns resulting from the television scanning format. The video scrambling facilities provided by the coding and multiplexing equipment are additional to any scrambling provided by the optical system.

Each video bit stream passes through a Wyner-Ash (16, 15) error-correction coder unit.^{1,8} Six error-correction coders are employed, together with six decoders in the receiving terminal, operating independently on the six bits of the video sample word. This arrangement provides effective interlacing of the error correctors and, because of the bit interleaving of channels A and B and the presence of a dummy bit in each video channel, gives protection against bursts of up to 14 consecutive errors in the final multiplexed bit stream.

The scramblers, which are of the self-synchronizing feedback type³ precede the error-correction coders and, in the receiving equipment, the descramblers follow the error correctors. By so arranging the equipment, operation of the error correctors is not upset by the inevitable error extension produced by the descramblers.

3.2 Sound Coding

The sound signals are coded using BBC 'NICAM-1'‡ equipment, which employs a digital companding (i.e.

† With h.d.p.c.m. coding, either p.c.m. or d.p.c.m. code-words are transmitted, the choice depending upon the dynamic characteristics of the video input signal.

‡ 'NICAM' stands for 'near-instantaneous companding audio multiplex'.

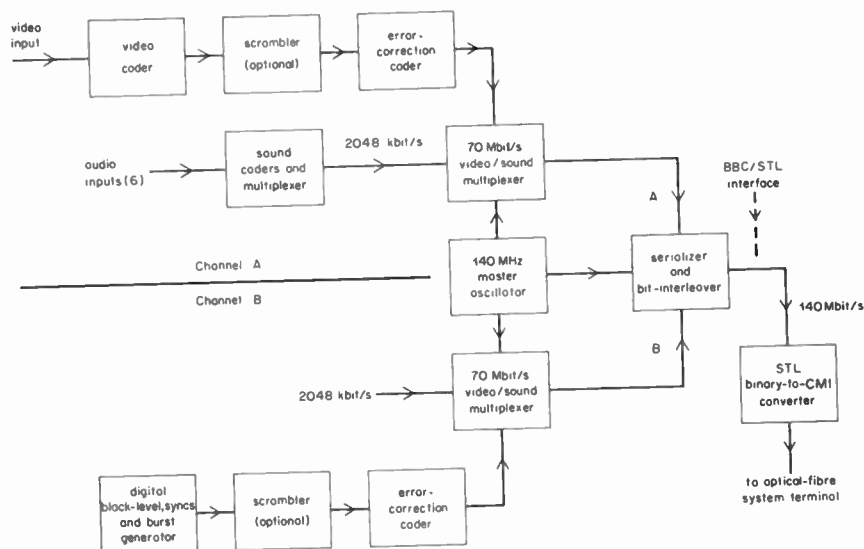


Fig. 4. The 140 Mbit/s coding and multiplexing equipment.

compressing and expanding) technique. In this equipment, six sound-programme signals are individually sampled at 32 kHz and initially quantized to 13-bit accuracy; they are then described using 10 bits per sample word.⁹ A resolution of 13, 12, 11 or 10 bits is chosen for each successive block of 32 samples in each of the six programme channels according to the peak amplitude occurring in that block. A 2-bit scale-factor word is transmitted with each block of 32 samples to indicate to the decoder the resolution of the 10-bit samples. Finally, the 10-bit sample words from all six channels are multiplexed together, with the scale-factor words, parity bits for use with an error-concealment technique (protecting sample words and scale-factor words) and framing information, to form a 2048 kbit/s bit stream.

3.3 Multiplexing

A 70 Mbit/s multiplexer combines the video bit stream and the bit stream from the 2048 kbit/s sound coder and multiplexer to make a composite parallel multiplex bit stream, adds one 'dummy' extra bit for each 6-bit video sample word, and frame alignment words and increases the bit rate to 70 Mbit/s (or rather, more precisely, 69.632 Mbit/s) using positive justification.

The two synchronous 70 Mbit/s signals are fed, in parallel form, to a serializer and bit interleaver. The resultant 140 Mbit/s serial bit stream is then converted to the CMI interface code⁵ for connection to the optical system.†

3.4 Error Monitoring

An error-monitoring unit in the receiving and demultiplexing equipment receives information on the number of error corrections taking place (from the video error correctors) and on the number of incorrect frame-alignment words received (from the multiplex frame detector). A digital counter in this unit can be used to count the error corrections or the incorrect frame-alignment words. It can also be set to display the bit-error rate in the received bit stream, derived either from the activity of the error correctors or from the incidence of incorrect frame-alignment words. Failure of parity framing (when the error correctors are unable to cope with the errors which are being received) or loss of multiplex frame synchronism are indicated by alarm lights and an optional audible alarm.

4 Preliminary Tests

Prior to the tests on the optical-fibre system between Hitchin and Stevenage, preliminary tests were conducted at STL in Harlow in order to check the operation of specially-constructed interface equipment (supplied by

STL) and to identify any other problems which might arise when the BBC equipment was connected to the optical system. Of particular interest was the performance of the Wyner-Ash error-correction codecs, which form part of the BBC multiplexing and demultiplexing equipment, in a transmission system which included the optical terminal scramblers and descramblers.

Certain configurations of scrambler can, when used after the error-correction coder, affect the performance of the error correctors. Table 1 illustrates the error extension produced by two possible configurations of the system 15-stage scrambler/descrambler upon 16 consecutive bits in the multiplexed bit stream when a single error is present. In the first example, an error in bit 0 creates errors in the first and fifteenth following bits as the error propagates through the descrambler shift register and is introduced into the descrambled bit stream when it reaches taps 1 and 15. It will be seen from the illustration that the error corrector designated 1B receives two errors in succession, which it is unable to correct, with this configuration. In the second example, the same degree of error multiplication is present but the errors are distributed to separate correctors (designated 1A, 4B and 1B in the figure) which are now able to cope with the situation. Tests in the laboratory confirmed that operation of the Wyner-Ash error correctors was upset by the presence of an optical system scrambler and descrambler with the 0, 1, 15 configuration.

It was possible to send the multiplexed video and audio signals through an optical modem without the optical system scrambler and descrambler fitted, provided that the video signals were scrambled in the BBC equipment. Under some circumstances a video signal could be sent unscrambled but this depended upon the picture content or test waveform and the type of coding (p.c.m. or d.p.c.m.). Errors in non-scrambled signals appeared to be most frequent in the region of the field sync pulses, and measurements of timing jitter revealed both line and field frequency components in the jitter waveform.

5 Main Tests

The BBC coding and multiplexing equipment was installed adjacent to the optical system terminal in the telephone exchange at Hitchin. The area in the exchange set aside for the experiments is shown in Fig. 5; the sending equipment is on the left-hand side of the picture and the receiving equipment on the right, with the optical-fibre terminal at the centre. A u.h.f. log-periodic aerial was mounted on the exchange roof to receive Band IV transmissions from the transmitter at Sandy Heath, and to provide, via an off-air receiver, 'live' video and audio signals for use in the tests. Tape-playing equipment provided two further audio signals, and video test equipment provided colour bars, colour 'chequer-board' (a pattern of coloured squares derived from

† The binary-to-CMI converter (and the CMI-to-binary converter for the demultiplexer equipment) were supplied by Standard Telecommunication Laboratories Ltd. (STL).

Table 1

The effect of two scrambler/descrambler configurations on the Wyner-Ash error correctors

EXAMPLE I

Scrambler with taps at 0, 1, 15

0	1	2	3	4	5	6	7	8	9	10	11	12	13	14	15
X	X														X
1A	1B	2A	2B	3A	3B	4A	4B	5A	5B	6A	6B			1A	1B

The result of using a (0, 1, 15) scrambler is that error corrector 1B receives two errors in succession, and therefore fails.

← Position of bits in shift register of descrambler.

← Position of extended errors resulting from a single error.

← Error correctors operating on bits from descrambler. (Error corrector 1A is 1st error corrector in channel A, etc. In this diagram, bits 12 and 13 are 'dummy' extra bits.)

EXAMPLE II

Scrambler with taps at 0, 7, 15

0	1	2	3	4	5	6	7	8	9	10	11	12	13	14	15
X							X								X
1A	1B	2A	2B	3A	3B	4A	4B	5A	5B	6A	6B			1A	1B

The results of using a (0, 7, 15) scrambler is that error correctors 1A, 4B and 1B each receive one error, which can be corrected.

← Position of bits in shift register of descrambler.

← Position of extended errors resulting from a single error.

← Error correctors operating on bits from descrambler.

vertical and horizontal colour bars), horizontal grey scale waveform, red screen and a sync, burst and black-level waveform. Another sync, burst and black-level waveform was generated digitally at the sub-Nyquist sampling rate of twice colour subcarrier frequency ($2f_{sc}$) and was available as a p.c.m. signal at the output of the generator. This p.c.m. signal was used as the digital video input for channel B throughout the tests.

The optical system was operated with a system scrambler in circuit at all times, as the use of a scrambler is an intrinsic feature of the design. For most tests, the optical system was looped at Stevenage at the terminal repeater level. This involved connecting the output of the regenerator to the laser modulator (see Fig. 2) so that the Stevenage equipment operated just as a repeater. For a short period of tests, however, the system was looped at the CMI level at Stevenage.

5.1 Error Rate Measurements

As already described, both the optical system and the demultiplexing equipment contained error detectors, so that the error rate could be monitored at two points simultaneously whilst the system was in traffic. The error monitor unit in the demultiplexer was normally used to indicate the number of error corrections taking place in the Wyner-Ash error corrector in channel A, although the error rate in channel B was checked with the monitor unit from time to time.

Error rates were recorded for a wide range of different conditions. The digital video signal in channel B was, for all tests, the digitally-generated sync, black level and colour burst p.c.m. signal; that in channel A was selected from the various test patterns available (colour bars, chequerboard, horizontal grey scale, red screen or sync, burst and black level) or was 'live' transmission material

from the u.h.f. receiver, and coded as p.c.m., d.p.c.m. or h.d.p.c.m. Two different configurations of system scrambler were used in the optical terminal (corresponding to the examples shown in Table 1), and the video scramblers† in the BBC equipment were used to give different combinations of scrambled and non-scrambled digital video signals.

The error rate of the optical system was checked from time to time during each day with the 100% chequer-board signal in the video A channel. A variation in the system error rate from 4.5 in 10^9 to 1.6 in 10^8 (a

† The video scramblers were of the self-synchronizing feedback type (as mentioned in Sect. 3.1) with six stages and tapped to give the (0, 1, 6) configuration.



Fig. 5. The experimental equipment installed in Hitchin telephone exchange.

range of about 3.5 : 1) was observed, and this appeared to be related to the level of activity of the telephone switchgear in the exchange. This aspect was not investigated further, however.

5.2 Measurements of Timing Jitter

Clock signals recovered in digital repeaters and receiving terminal equipment inevitably contain some residual timing disturbances (known as 'jitter') which will be imparted to the digital signals whenever they are retimed. Jitter can impair system performance by increasing the incidence of errors and by phase modulation of analogue signal samples in digital-to-analogue converters.

Purpose-built measuring equipment was used to monitor timing jitter on the received signal at the CMI interface between the optical terminal and the video and audio demultiplexing equipment. Jitter was recorded with the same conditions as for the error tests, but a different combination of video signals was selected for transmission in channel A, to include signals which had been found to give the highest levels of jitter. Tests were also conducted with a pseudo-random sequence of digits in place of the A channel video signal.

5.3 Tests on System Looped at CMI Interface Level in Stevenage Exchange

A brief series of tests was conducted with the connection between the 'outward' and 'return' paths at Stevenage made at the CMI interface level instead of the terminal repeater level. This is best illustrated by referring to the schematic diagram of Fig. 2. When the system operates with a loop at the terminal repeater level, the output of the regenerator is connected directly to the input of the laser modulator (i.e. only the two blocks on the right-hand side of the diagram are involved in the loop connection). Looping at CMI interface level involves the entire line terminal equipment shown in the diagram; in other words, the connection between outward and return paths is made at the extreme left-hand side of the diagram.

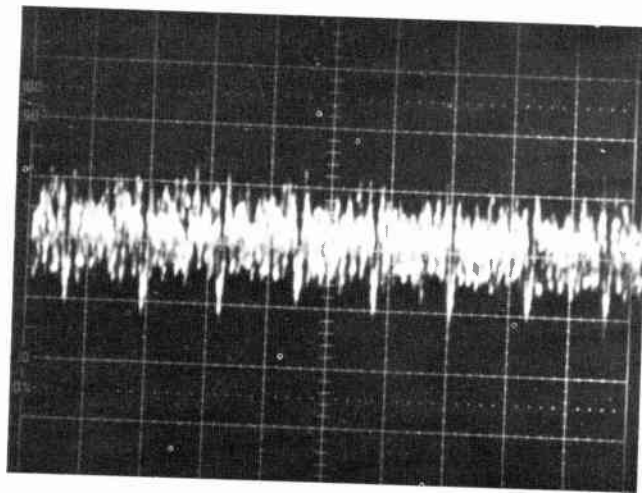
5.4 Subjective Tests

Throughout the tests, informal subjective assessments of received picture and sound quality were made. As might be expected, consistently good picture and sound quality, as determined by the source coding parameters, was obtainable from the system during the entire period of the tests.

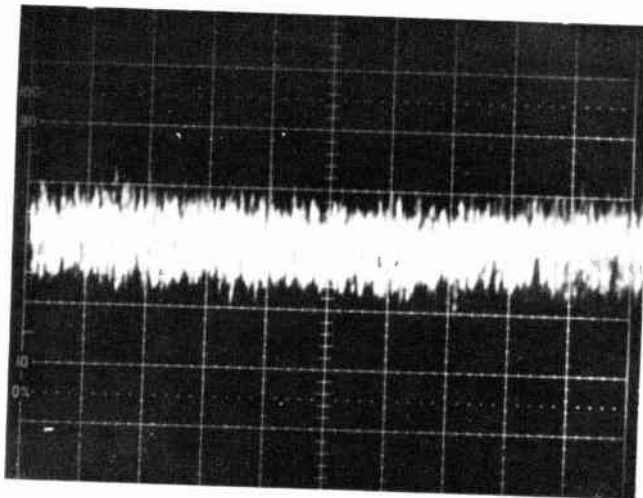
6 Results of the Tests

The optical-fibre system was able to carry multiplexed digital audio and video signals between Hitchin and Stevenage with negligible impairment to picture or sound quality during the course of the tests.

The bit-error rate which was present in the optical system when it was carrying the telephony multiplex traffic was about $2 \text{ in } 10^{10}$. This rose to an average value of about $8 \text{ in } 10^9$ when digital video signals were being sent over the system. The bit-error rate in the bit stream higher than this, owing to the error multiplication produced in the optical system descrambler. On average, the bit-error rate registered by the video error-monitoring facility in the demultiplexing equipment was about $3 \text{ in } 10^8$ (i.e. about four times the error rate in the optical system). The errors appeared to be randomly distributed and well within the correction capability of the Wyner-Ash error correctors in the BBC demultiplexing equipment.



(a) Without video scrambling.



(b) With both video signals scrambled.

Fig. 6. Jitter waveforms from STL jitter-measuring equipment connected to the output of the optical system.
Vertical scale: 25° per major division.
Horizontal scale: $50 \mu\text{s}$ per major division.

When the system was looped at the CMI level, parity information from the incoming signal at Stevenage (the 'outward' path from Hitchin) was removed by the parity checking circuits, so that the system error-monitoring circuits at Hitchin detected only the errors introduced on the 'return' path from Stevenage. This did not affect the operation of the Wyner-Ash error correctors or the error monitoring circuits in the BBC demultiplexing equipment. The error rates observed in the video signal when the system was looped at the CMI level was very similar to those recorded when the system was looped at the terminal repeater level, but the error rate indicated by the error detector in the optical system terminal equipment was reduced by a factor of 10 in this condition. This was a greater reduction than had been anticipated (one would have expected the error rate indication to be halved), and it was deduced therefore that the outward path, from Hitchin to Stevenage, was contributing substantially more errors than the return path.

The error performance of the optical system, although acceptable, differed from that found from tests on electrical cable systems. The previous tests on 120 Mbit/s cable systems² showed that the cable-system error rate was normally virtually zero and this was for a considerably longer circuit (200 km with 106 regenerative repeaters) than the optical system (18 km with four repeaters). The precise cause of the errors in the experimental optical system was not clear but it might possibly have been non-optimal functioning of the timing-extraction circuits in some repeaters particularly in the circuit from Hitchin to Stevenage.

The measurements of timing jitter revealed that, when video and audio signals were being sent over the system, the lowest levels of peak-to-peak jitter were obtained when both digital video signals were scrambled. No significant difference in jitter performance could be detected when the optical system scrambler was changed from the (0, 7, 15) scrambler to the (0, 1, 15) scrambler. In the tests with live television transmissions providing the video signal for channel A, and no video scramblers in use, a slow variation could be observed in the r.m.s. jitter. This was found to be a result of the slight difference between the field frequency of the television transmission and that of the test equipment which was providing the video signal for channel B. When the video scramblers were used, the variation disappeared. No conditions were found where the jitter amplitude was sufficiently high to cause bit errors.

Timing jitter was rather higher when the system was looped at the CMI level. It is possible that this increase in jitter could have been caused by the introduction of an additional phase-locked loop into the system at Stevenage, but this was not investigated.

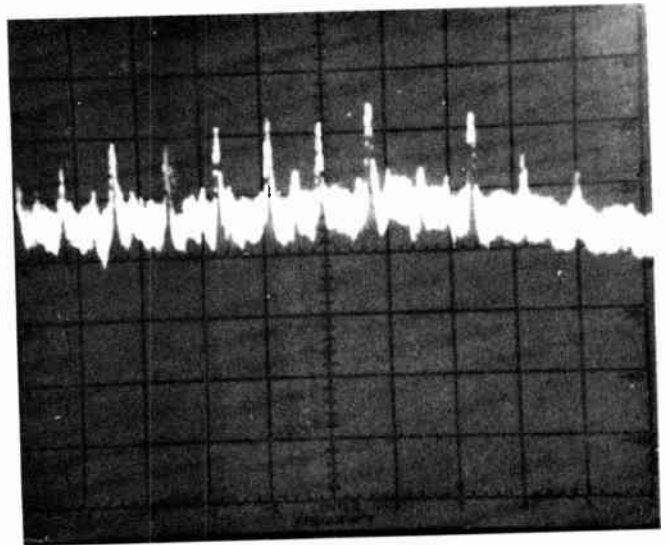
Inspection of the jitter waveform in the jitter-

measuring equipment revealed that without video scrambling, components at the television line rate of 15.625 kHz were present, but were removed when scrambling was introduced. The jitter waveforms are shown in Fig. 6 and the corresponding spectra are shown in Fig. 7 (picture content was found to have very little effect).

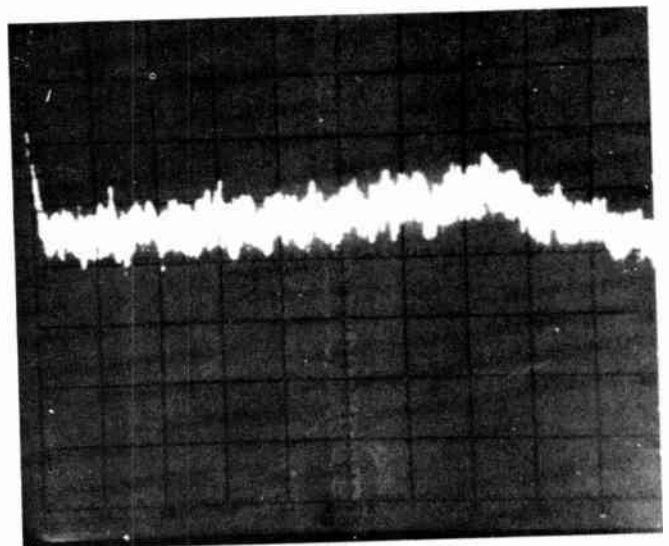
Detailed tables of the results obtained during the tests described in this paper may be found in Ref. 10.

7 Conclusions

Tests conducted on an experimental 140 Mbit/s optical-fibre transmission system have shown that no



(a) Without video scrambling.



(b) With both video signals scrambled.

Fig. 7. Spectrum of jitter signal from STL jitter-measuring equipment connected to the output of the optical system.

Vertical scale: 10 dB per major division.
Horizontal scale: 20 kHz per major division.

serious difficulties arose when transmitting digital sound and television signals. Although some pattern dependency was detected, the consequential effects on timing jitter and error performance were well within acceptable limits and no impairment of the audio or video signals could be perceived by listeners and viewers during informal subjective tests. The difference in bit-error rate between the Hitchin–Stevenage and Stevenage–Hitchin sections of the system seems to imply that some of the repeater timing-extraction circuits were not functioning optimally during these tests, and may have been contributing to the slight pattern dependency which was observed.

The results obtained indicate that the optical transmission system is substantially transparent, owing to the incorporation of the self-synchronizing system scrambler. Nevertheless, they also show that additional scramblers may be beneficial when highly-structured signals, such as digital television signals, form part of the traffic. Furthermore, the tests have emphasized that care must be taken to ensure that scramblers which operate on signals protected by error-correction codes do not interfere with the functioning of the error corrector.

8 Acknowledgments

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of Standard Telecommunication Laboratories Limited and by colleagues in BBC Research Department.

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Using microprocessors to count microwave signals

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SUMMARY

The harmonic heterodyne frequency measuring technique described here provides wide f.m. tolerance, high sensitivity and automatic amplitude discrimination. It requires only one microwave component—a sampler—and uses a microprocessor to perform the required computations. This minimizes the hardware content and facilitates data manipulation.

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1 Introduction

Direct-counting frequency meters are restricted by the speed of today's logic circuitry, which tails off around 500 MHz. Consequently automatic microwave counters must employ some form of down-conversion to extend frequency measurement into the gigahertz range. Simple prescaling is feasible for input frequencies of up to 1.5 GHz, but beyond this something more than straight division is needed.

One classical method involves heterodyne conversion. Here incoming microwave signals as high as 20 GHz are mixed with a high-stability local oscillator signal of known frequency, resulting in a difference frequency within the range of a conventional counter (see Fig. 1). The high-stability local oscillator signal is generated by first digitally multiplying the frequency of the instrument's time-base to a convenient fundamental frequency, f_n , typically in the range 100 MHz to 500 MHz. Then this is put through a harmonic generator to produce a comb line of frequencies necessary to encompass the full frequency range of the counter. One line of this comb, kf_n , is then selected by a microwave filter and directed to the mixer. Emerging from the mixer is an intermediate frequency equal to $f_x - kf_n$. This is amplified and sent to the counter. The display contains the sum of the intermediate frequency and kf_n . In practice the system begins with $k = 1$ and steps the microwave filter through the comb line until a detector indicates that an intermediate frequency in the desired range is present. The microwave filter may be a YIG filter or an array of thin-film filters that are selected by p-i-n diode switches.

Another method involves the use of a transfer oscillator (see Fig. 2). Used in a number of counters working up to around 23 GHz, this technique is based on phase-locking a low-frequency voltage-controlled oscillator (v.c.o.) to the microwave input signal. The v.c.o. frequency f_1 can then be measured in a conventional counter and all that remains is to determine the harmonic relationship between the frequency and the input. A second v.c.o. is often used to help determine the harmonic number n . The counter then measures f_1 , multiplies the n factor (usually by extending its gate time) and displays the result. Comparing these two techniques the heterodyne method is faster, more tolerant to f.m. on the input signal and less sensitive.

2 Harmonic Heterodyne Conversion

The advent of microprocessors has enabled a new technique called harmonic heterodyne conversion to be applied. Providing many of the advantages of both traditional techniques while significantly reducing cost, this method requires only one microwave component, a sampler, and relies on a microprocessor to perform the required computations, thereby eliminating a good deal of digital hardware and its associated expense. The

microprocessor also adds to the counter's power and versatility by making it possible to manipulate the measured data.

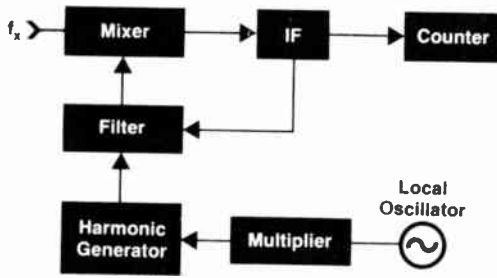


Fig. 1. Basic heterodyne converter configuration, where unknown microwave frequencies are mixed with a high-stability local oscillator signal of known frequency to derive a difference frequency within the range of a conventional counter.

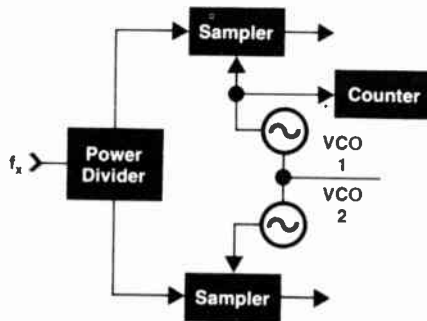


Fig. 2. Typical transfer oscillator schematic. This relies on phase locking a low-frequency voltage-controlled oscillator to the microwave input signal. A second v.c.o. is then used to establish the harmonic relationship between the two frequencies. This is usually achieved by varying the gate time in multiples of n .

A simplified block diagram of a typical harmonic heterodyne converter is shown in Fig. 3. All the harmonics of the internal oscillator are simultaneously mixed with the unknown signals by a sampler, the output of which comprises the sum and difference frequencies produced by each harmonic of the internal oscillator, a programmable frequency synthesizer locked to the counter's time-base is incremented in frequency until one of the outputs of the sampler is in the counting range of the low-frequency counter. The i.f. is then counted and the unknown frequency is then determined from the relation:

$$f_x = Nf_1 \pm f_{i.f.1}$$

where N is the harmonic number of the frequency synthesizer, f_1 is the programmed frequency of the synthesizer and $f_{i.f.1}$ is the product of Nf_1 and f_x .

The frequency f_1 of the programmable synthesizer is known, since it is known where the indexing of the synthesizer is stopped by a signal from the i.f. detector.

And $f_{i.f.1}$ is known since it is checked by the low-frequency counter. Still to be determined are N and the sign (\pm) of the i.f. This is done by making one more i.f. measurement with the synthesizer frequency offset from its previous value by a known amount ΔF . The resulting second intermediate frequency is $f_{i.f.2}$, which is also checked by the low-frequency counter. The offset ΔF is much smaller than the i.f., so if f_x is less than Nf_1 , $f_{i.f.1}$ (produced by mixing Nf_1 with f_x) will be greater than $f_{i.f.2}$. Conversely if f_x is greater than Nf_1 , $f_{i.f.1}$ will be less than $f_{i.f.2}$; N is then determined as follows:

$$f_{i.f.1} = Nf_1 - f_x \quad (Nf_1 > f_x)$$

$$f_{i.f.2} = Nf_2 - f_x \quad (Nf_2 > f_x)$$

$$N = \frac{f_{i.f.1} - f_{i.f.2}}{f_1 - f_2}$$

or if f_x is greater than Nf_1 , then

$$f_{i.f.1} = f_x - Nf_1 \quad (Nf_1 < f_x)$$

$$f_{i.f.2} = f_x - Nf_2 \quad (Nf_2 < f_x)$$

$$N = \frac{f_{i.f.2} - f_{i.f.1}}{f_1 - f_2}$$

The unknown frequency is then computed as follows:

$$f_x = Nf_1 - f_{i.f.1} \quad (f_{i.f.2} < f_{i.f.1})$$

$$f_x = Nf_1 + f_{i.f.1} \quad (f_{i.f.2} > f_{i.f.1})$$

Since the frequency of the synthesizer is known to the accuracy of the counter's time-base and the i.f. is referenced back to the same source, the accuracy of the microwave measurement is determined by the time-base error and the ± 1 count error inherent in any counter.

3 Frequency Modulation Problems

The discussion up to now has dealt with the ideal case in which the counter is measuring input signals with little or no f.m. However, many signals in the microwave

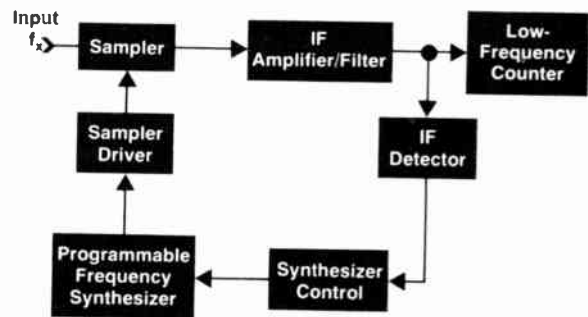


Fig. 3. Simplified schematic of the harmonic heterodyne frequency conversion technique which involves mixing all the harmonics of a frequency synthesizer with the unknown in a sampler. The synthesizer frequency is incremented until one of the sampler outputs falls within the range of the associated low-frequency counter. A microprocessor controls the synthesizer and performs the necessary computations to give a direct-reading output.

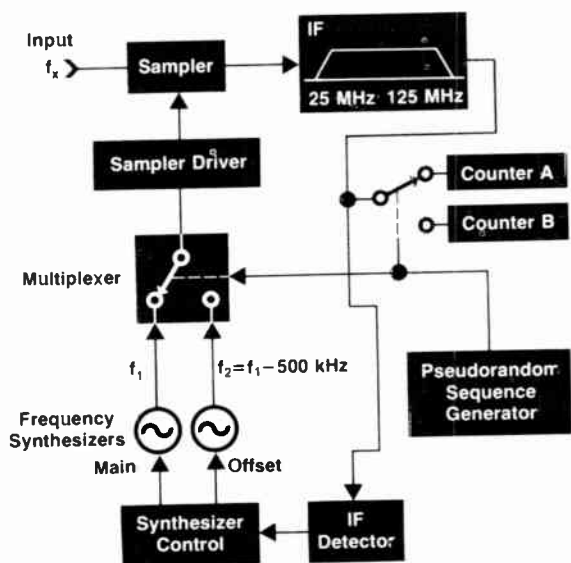


Fig. 4. A pair of synthesizers and counters are used in Hewlett-Packard's first counter to exploit the harmonic heterodyne technique. This is to prevent frequency modulation on an unknown signal locking up the synthesizers at the wrong frequency. A pseudo-random sequence generator switches between the two synthesizers and associated counters to minimize risk of any coherence between the switching rate and the f.m. level on the unknown signal.

region, particularly those from microwave radios, have significant f.m. on them. To prevent this modulation from causing incorrect computation of the harmonic number n , harmonic heterodyne techniques are implemented as shown in Fig. 4. There are two synthesizers offset by precisely 500 kHz, two counters and a pseudo-random sequence generator that controls a multiplexer and two counters synchronously.

The operating algorithm is as follows: with the multiplexer having selected the main oscillator output, the main oscillator frequency f_1 is swept from 350 MHz to 300 MHz in 100 kHz steps. The offset oscillator frequency f_2 is maintained at $f_1 - 500$ kHz by a phase-lock loop. When the i.f. detector indicates the presence of an i.f. signal in the range of 50 MHz to 100 MHz, the synthesizer stops its sweep and the counter starts its determination of the harmonic number n . The pseudo-random sequence output switches between the main oscillator and the offset oscillator and between counters A and B. The harmonic number and the sign of one i.f. are computed as previously described, since counter A accumulates $f_{i.f.1}$ and counter B accumulates $f_{i.f.2}$. The pseudo-random sequence is then disabled, the main oscillator is selected, and the frequency $f_{i.f.1}$ is measured by counter A to the selected resolution. The pseudo-random switching prevents any coherence between the switching rate of the multiplexer and the modulation rate of the f.m. that might produce an incorrect computation of n . The pseudo-random effect is analysed as follows:

Let $g(t)$ represent a period of one sequence as shown in Fig. 5.

When $g = +1$ the local oscillator is set to f_1 and $f_{i.f.1}$ is gated into counter A. When $g = -1$ the local oscillator frequency is changed to f_2 and $f_{i.f.2}$ is gated in counter B. The number $M(t)$ accumulated by a separate counter is given by the integral of the frequency during the time the counter is gated on. Let $M_1(t)$ and $M_2(t)$ represent the numbers accumulated by counters 1 and 2 respectively.

$$M_1(t) = \int_0^{t_1} f_{i.f.1}(t) dt + \int_{t_2}^{t_3} f_{i.f.1}(t) dt + \dots$$

$$= \int_0^T f_{i.f.1}(t) dt \quad \text{for } g = +1$$

$$M_2(t) = \int_{t_1}^{t_2} f_{i.f.2}(t) dt + \int_{t_3}^{t_4} f_{i.f.2}(t) dt + \dots$$

$$= \int_0^T f_{i.f.2}(t) dt \quad \text{for } g = -1.$$

Let us now consider the upper sideband case. The lower sideband is identical except for a change in sign.

$$f_{i.f.1}(t) = f_x(t) - Nf_1$$

$$f_{i.f.2}(t) = f_x(t) - Nf_2$$

$$M_1(t) = \int_0^T f_x(t) dt - Nf_1 \int_0^T dt \quad \text{for } g = +1$$

$$M_2(t) = \int_0^T f_x(t) dt - Nf_2 \int_0^T dt \quad \text{for } g = -1$$

$$M_1(t) - M_2(t) = \int_0^T g(t)f_x(t) dt - N(f_1t_1 - f_2t_2)$$

where t_1 is the length of time for which $g = +1$ and t_2 is the length of time for which $g = -1$. Then:

$$N = \frac{M_2 - M_1}{t_1f_1 - t_2f_2} + \frac{\int_0^T g(t)f_x(t) dt}{t_1f_1 - t_2f_2} = N_c + \epsilon.$$

The value of one integral is not available so the second term appears as an error in the calculated harmonic number N_c . For a pseudo-random sequence $t_1 - t_2 = t_c$ where t_c is the sequence clock period. This difference can be made zero by a small modification of the sequence circuit or it can be disregarded if the sequence is long enough. Thus:

$$t_1 = t_2 = t/2$$

$$f_1 - f_2 = \Delta f$$

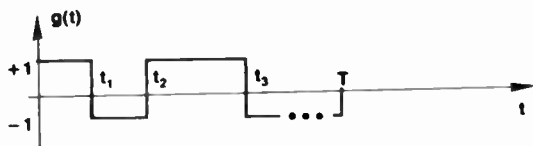


Fig. 5. Output of the pseudo-random sequence generator used to gate the counters. (See text.)

$$N = \frac{2(M_2 - M_1)}{T\Delta f} - \frac{2}{T\Delta f} \int_0^T g(t)fx(t) dt.$$

Considering just the error term as function of time and taking the Fourier transform and the shifting relationship, the equation becomes:

$$E(j\omega) = \frac{2}{T} \left[\frac{1 - \exp(-j\omega t_1)}{j\omega} - \frac{\exp(-j\omega t_1) - \exp(-j\omega t_2)}{j\omega} + \dots \right] \frac{F_x(j\omega)}{\Delta f}$$

The term in brackets can be recognized as the Fourier transform of the pseudo-random sequence. The term $(F_x(j\omega)/\Delta f)$ can be considered as the input to a linear system and $E(j\omega)$ as the output. The transfer function of the system is:

$$H(j\omega) = \frac{2}{T} \left[\frac{1 - \exp(-j\omega t_1)}{j\omega} - \frac{\exp(-j\omega t_1) - \exp(-j\omega t_2)}{j\omega} + \dots \right]$$

To determine the counter's sensitivity to sine wave modulation present in $fx(t)$ it is necessary to know the $|H(j\omega)|$. This can be evaluated at frequencies that are multiples of $f_0 = 1/T$ by taking the square root of the power spectrum for the repetitive sequence, thus:

$$|H(jn2\pi f_0)| = 2 \frac{\sqrt{p+1}}{p} \left| \frac{\sin \pi/p}{n\pi/p} \right| \quad n > 0$$

where p is the sequence length in clock periods

$$p = \frac{T}{T_c}$$

To take care of the actual value of $|H(f)|$ for frequencies not multiples of f_0 a safety factor of 2 is used. The worst case occurs for frequencies of the order of f_0 . For large values of p :

$$\epsilon_{\max} \approx \frac{4}{\sqrt{p}} \quad \frac{\Delta f_x}{\Delta f} \leq 1/2 \quad p \geq 64 \left(\frac{\Delta f_x}{\Delta f} \right)^2$$

where Δf_x is the peak deviation of the unknown signal.

4 Setting F.M. Thresholds

Thus to allow the counter to tolerate 10 MHz peak f.m. on the input signal and $f = 500$ kHz, then the p value should exceed 25,600. Since $p = 2^m - 1$, where m is the number of shift register stages in the pseudo-random sequence generator, a 15-stage shift register would be needed to generate this sequence. It is obvious from the expression for p that the more f.m. one wants to tolerate the longer the sequence has to be, which in turn affects the counter's measurement time. This has to take into account sweep time, gate time and time to determine n . Typically sweep time is 150 ms or less and the gate time

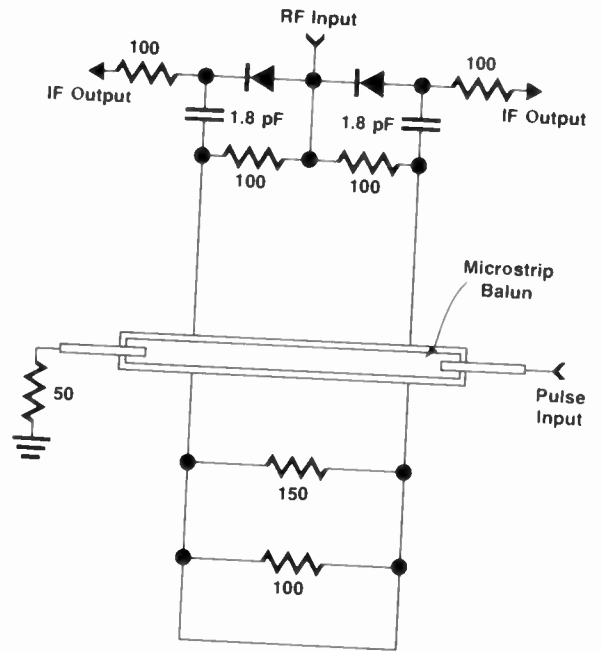


Fig. 6. Operation of this thin-film sampler is similar to that of samplers used in earlier H-P microwave counters. The sampling pulse couples to the slotted line through a balun that generates two opposite polarity pulses to drive Schottky diodes.

for 1 Hz resolution is one second. In normal operation the pseudo-random sequence length is 360 ms, thus making the total measurement time about 1.5 s. Under these conditions, the counter can tolerate 20 MHz peak-to-peak frequency deviation on the unknown. This corresponds to the bandwidth of most telecommunications channels. The maximum f.m. that can be tolerated is obviously determined by the total bandwidth of the i.f.

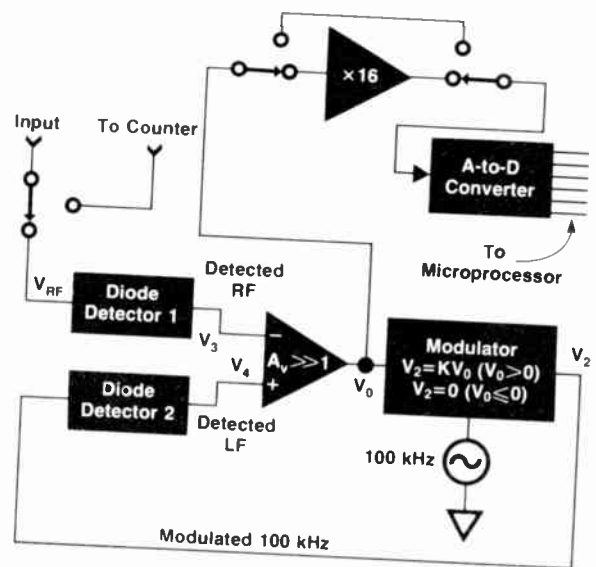
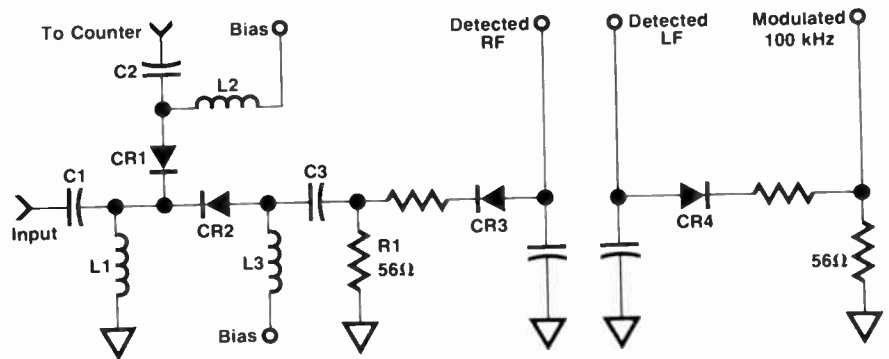


Fig. 7. A pair of Schottky diode detectors are used to compensate for any temperature variations. The d.c. output voltage remains proportional to the r.f. sinewave voltage so long as the temperature of the matched detectors remains the same.

Fig. 8. Amplitude measurement system. This is fabricated using thin-film technology and features planar low-barrier Schottky diodes manufactured in-house.



5 The Thin-film Sampler

The only microwave component required to make the frequency measurements in the new design (HP5342A) is a sampler (see Fig. 6). This comprises a thin film hybrid mounted in an aluminium package. On the hybrid are two Schottky diodes placed across a slotted line. The sampling pulse couples to the slotted line through a stripline balun that generates two opposite-polarity pulses to drive the diodes. The two down-converted signals are taken from two isolated resistors to the preamplifier: resistors across the slot are used to absorb secondary reflections introduced by the sampling pulse.

Measurement, control and system co-ordination in this technique are implemented using a 6800 microprocessor. This handles interfacing to the analogue circuits, power-up routine, display control, keyboard manipulation and control, frequency and amplitude measurements, arithmetic calculations and diagnostics. Memory support includes 128 bytes of r.a.m. and 2 k bytes of r.o.m. that store the m.p.u. programs. Decoder drivers for peripheral circuit interfaces and diagnostic switches to facilitate signature analysis for trouble shooting are also located on the processor board. In addition use of the microprocessor has enabled the accurate measurement of amplitude simultaneously with frequency. A schematic of this section is shown in Fig. 7.

The actual switching circuitry is shown in Fig. 8. Diodes 1 and 2 are planar Schottky devices that behave like point-contact diodes and have the stability and ruggedness of ordinary Schottky diodes. They do not respond to r.m.s. above levels of -20 dBm, thus the circuit actually measures voltage. The instrument is made to read out in units of power and remains accurate so long as the signal does not contain much amplitude modulation.

The d.c. output voltage V_0 is proportional to the r.f. sine wave voltage regardless of the transfer function of the detectors so long as the detectors are matched. Microwave components for the amplitude measurement are integrated on two sapphire substrates contained in a common housing.

One substrate is used for the microwave detector CR3

and the other for the low frequency detector CR4; CR1 and CR2 are p-i-n diodes used to route the input signal either to the counter or to the detector. Inductors L1, L2 and L3 act as r.f. chokes at high frequencies and as part of the 500 MHz high-pass filter with capacitors C1 to C3 at lower frequencies. This method of injecting bias into a broadband, planar circuit minimizes losses. Low s.w.r. is important for amplitude measurements. Accordingly, careful attention was paid to parasitic reactances and impedance matching: Fig. 9 shows a typical s.w.r.

The amplitude measurement circuit is made to serve as an attenuator by biasing p-i-n diode CR2 with high current and p-i-n diode CR1 with only about 0.4 mA. With this technique the amplitude measuring hybrid can handle high levels ($+20$ dBm) without overloading the input to the sampler.

6 Data Manipulation

To display amplitude in dBm, a logarithm of the measured voltage is needed. Utilizing a 13-bit integrating analogue-to-digital converter allows this function to be accomplished by the microprocessor. Dynamic range is further increased by a switchable d.c. amplifier ahead of the a.d.c. This combination allows better than 0.03 dB resolution at all levels and avoids the drift problem of a logarithmic amplifier. Once the processor has logged and scaled the result it uses the frequency information obtained in a previous measurement to correct the results. Since the microprocessor knows the frequency

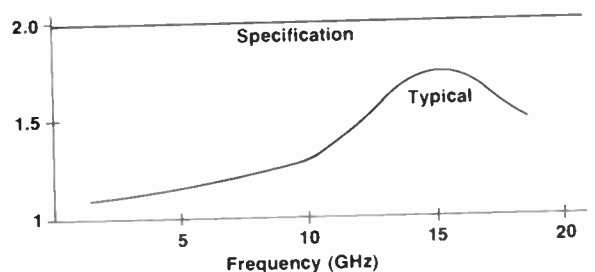


Fig. 9. Typical standing wave ratio plot for the 5342A counter in the amplitude measurement mode. This plot includes the front panel connector.

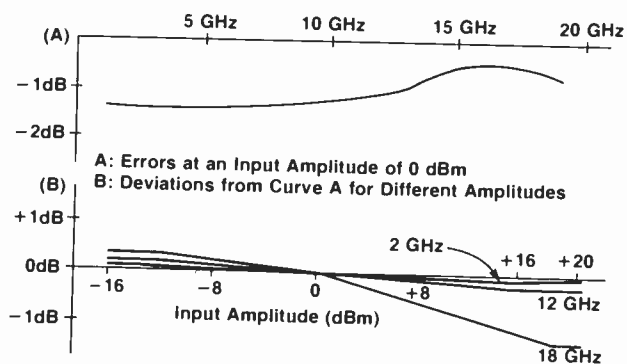


Fig. 10. To minimize errors due to standing waves, systematic errors are stored in a p.r.o.m. so that the microprocessor can correct each measurement before display. The look-up table is created during final test.

and measures amplitude of the input signal it is possible to correct not only for the frequency response of the detector and the insertion loss of the input connector and switch, but also for amplitude related errors such as mistracking between detectors. Typical errors to be corrected are shown in Fig. 10. Measurements such as these are made on every hybrid unit before installation and the data are stored in a p.r.o.m. Curve A is stored to eight-bit resolution in 250 MHz increments and curve B also to eight-bit resolution in 4 dB increments and every 500 MHz. The processor uses the frequency and uncorrected amplitude of the input signal to look up the values of correction in each table and to interpret for intermediate values. Since the table for curve B is normalized to the level used in the table for curve A, the correction values are simply added together and applied to the result. The factors affecting the overall amplitude accuracy of the system are as follows:

- Traceability of standards to NBS and their application to individual instruments in production; 0.4 dB is allowed.
- Digital processors can contribute up to 0.1 dB.
- Analogue errors such as op-amp offset, modulator linearity, noise a-d conversion accuracy can add a maximum of 0.3 dB.

(d) Drift with temperature from 0°C to 50°C is typically 0.2 dB.

Worst case uncertainty for all these variables together is thus ± 1.2 dB. These errors although not random are independent and will rarely add to this value. The root sum of the squares uncertainty, a more realistic value of expected error, is less than 0.6 dB. To this value must be added mismatch uncertainty which depends upon the s.w.r. of both the source and load.

7 Conclusions

The microprocessor in the new instrument is not only used intrinsically in making the measurement itself, but also adds to the instrument's power and versatility by making it possible to manipulate the measured data. Keyboard entry permits the operator to define frequency and amplitude offsets and also make deviation measurements. The HP5342A measures frequency from 10 Hz to 18 GHz with a resolution of 1 Hz. It has an 11-digit display and with an amplitude measurement option, the user can see input level displayed in dBm with 0.1 dB resolution and ± 1.5 dB accuracy. An option extends the frequency range up to 24 GHz without derating the 1 Hz resolution.

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Cellular pipeline digital filters

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SUMMARY

This paper describes an efficient method for the implementation of digital filters with special-purpose hardware. The second-order section is realized by applying the canonic form algorithm in conjunction with two's complement serial arithmetic. The pipeline implementation was used for high-speed performance. The objectives were to obtain a cascaded cellular structure with small circuit complexity and well suited for practical l.s.i. realization. The derived cellular filter is a functionally complete second-order section, that is, it provides the arithmetic operations, the delay requirements and the rounding using only one control signal. Also, it is expandable and permits flexible filter design. Higher-order digital filters can be obtained with repeated use of the proposed second-order section.

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1 Introduction

Digital filtering has been implemented mainly by software, due to the large number of computations involved. However, in many applications, as in communication systems, there is need for real-time digital signal processing. With the recent advances in large-scale integration, an inexpensive hardware implementation of digital filters is now possible. Some interesting work has been done in this direction. Jackson *et al.*¹ proposed a filter structure well suited for l.s.i. by introducing the serial-pipeline multiplier.² In Peled and Liu's realization³ the partial results of the filter algorithm are precomputed and stored in a r.o.m. Thus, no multiplier is required. Another interesting implementation is that proposed by De Mori⁴ which yields a cellular structure. In this work all the arithmetic operations are based on the carry-save addition technique.

The complete operation, however, of a digital filter requires, besides the arithmetic operation, other operations too. These operations are the data delay, the rounding and mainly the synchronization of the whole system with the appropriate control. Such a design has not been considered up to now. The addition of the above features to the existing realizations destroy their modular form, making them inefficient for l.s.i. implementation. Also, a digital filter design must be flexible so as to allow an increase in the accuracy. The above mentioned implementations, however, are designed to handle a fixed number of coefficient digits.

In this paper a digital filter implementation is introduced which overcomes the above disadvantages. The proposed implementation is well suited for practical l.s.i. realization because it provides the following features:

- Functional completeness
- Full cellularity
- Small circuit complexity
- One control signal
- High-speed operation

The above features were achieved by designing a special pipeline circuit with the following properties:

- The data delay units inherent with the pipeline operation play also the role of the digital filter delays. Thus, no external memory is required.
- The rounding operation is performed within the pipeline circuits automatically.
- All the operations are performed in two's complement serial arithmetic and no conversions are required.

The basic idea used in the proposed circuit is to realize the recursive and the non-recursive parts of the second-order section by treating the operations in each part (two additions and two multiplications) as a single operation. That is, the partial terms of the two multiplications and the two additions are all added simultaneously in the

same circuit using two's complement serial arithmetic.

The above implementation yields a stand-alone, completely cellular second-order section. In this implementation each cell can be considered as a one-bit second-order section and consequently this structure is very flexible to the extent that more stages could be added to increase the accuracy of the operation. Finally, the structure of the second-order section is such that it permits the construction of a higher-order filter with the same speed as the second-order section.

Details on the operation of the second-order section are given in Section 2. In Section 3 the design of a higher-order digital filter is presented that uses as a building block the proposed pipeline second-order section. Finally, in Section 4 the operation speed, the circuit complexity and the power consumption are considered. Furthermore, comparisons are made with previously proposed realizations.

2 The Proposed Pipeline Second-order Section

The input-output relation of the second-order digital filter is described by the following equation,

$$y_n = x_n + a_1 x_{n-1} + a_2 x_{n-2} - b_1 y_{n-1} - b_2 y_{n-2} \quad (1)$$

where $\{x_n\}$, $\{y_n\}$ are the input and the output sequences, respectively. The constants a_1 , a_2 , b_1 , b_2 are the coefficients of the filter. The direct implementation of eqn. (1) requires four delays, while the implementation with the canonic form requires only two. This minimum-memory canonic form algorithm is described by the following equations,

$$z_n = x_n + b_1 z_{n-1} + b_2 z_{n-2} \quad (2a)$$

$$y_n = z_n + a_1 z_{n-1} + a_2 z_{n-2} \quad (2b)$$

where $\{z_n\}$ is the sequence of the internal state variable.

Equations (2a) and (2b) are of the same form and each one involves two additions and two multiplications. Therefore, only the implementation of eqn. (2a) needs to be examined. For digital representation of the variables and the coefficients, their range must be restricted. The variables' magnitude is scaled to be less than one. The coefficients' magnitude is assumed to be smaller than two, which holds for most digital filter applications. In accordance with the above restrictions, the variables and the coefficients are represented in fixed-point two's complement form as follows:

$$x_n = -x_n^0 + \sum_{i=1}^{N-1} x_n^i 2^{-i} \quad (3a)$$

$$z_n = -z_n^0 + \sum_{i=1}^{N-1} z_n^i 2^{-i} \quad (3b)$$

$$b_{1,2} = -b_{1,2}^0 2^1 + \sum_{j=1}^{M-1} b_{1,2}^j 2^{-j+1}. \quad (3c)$$

Substitution of eqn. (3c) in eqn. (2a) yields,

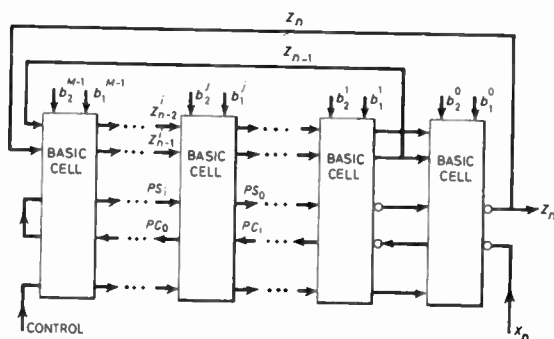


Fig. 1. The implementation of the recursive part of a second-order digital filter.

$$z_n = -(z_{n-1} b_1^0 + z_{n-2} b_2^0) 2^1 + x_n + \sum_{j=1}^{M-1} (z_{n-1} b_1^j + z_{n-2} b_2^j) 2^{-j+1}. \quad (4)$$

This equation leads to a cellular realization that treats the partial terms in a single operation. The complete second-order section is shown in Fig. 1. The circuit of the cell used to realize the individual terms of eqn. (4) is shown in Fig. 2. The proposed circuit incorporates the pipeline technique in a similar way as in the serial pipeline multiplier, namely, it accepts the data serially and the coefficients in parallel form, both in two's complement notation. The result that is obtained is also in serial two's complement form.

The cell of Fig. 2 consists of two cascaded full adders. These full adders in the j th-order cell compute serially the partial term,

$$A_j = (z_{n-1} b_1^j + z_{n-2} b_2^j) 2^{-j+1}. \quad (5)$$

This term is added to the inputs PS_i , PC_i to produce the outputs PS_o , PC_o . By cascading such cells the partial terms are added to realize eqn. (4).

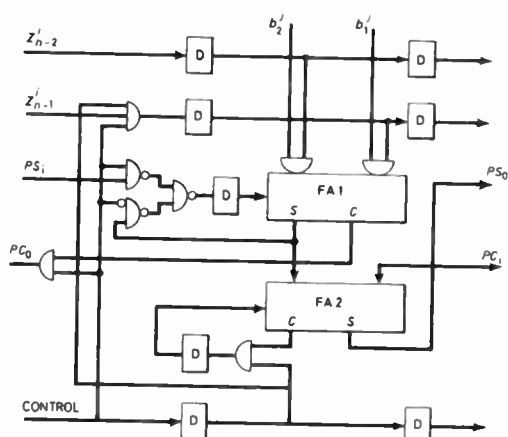


Fig. 2. The basic cell of the proposed implementation.

Next the operation of the cell is analysed. Suppose that the j th cell forms the quantity,

$$(z_{n-1}^i b_1^j + z_{n-2}^i b_2^j) 2^{-i-j+1}.$$

Simultaneously, the next cell computes the quantity below, considering that a two-bit delay is inserted in the path of the internal variables (z_{n-1}, z_{n-2}) ,

$$(z_{n-1}^{i+2} b_1^{j-1} + z_{n-2}^{i+2} b_2^{j-1}) 2^{-i-j}.$$

To synchronize the output PS_o of the j th cell with the quantities added in the $j-1$ cell, a delay unit is inserted after the PS_i input in each cell. The PC_o output of the $j-1$ cell, which is a carry, has the same weight as the product of the previous cell and is connected directly to PC_i of the j th cell. This reverse connection does not imply backward propagation, because the carry PC_o of the $j-1$ stage that is carried back is stopped in the second full adder of the j th cell and does not continue to propagate.

After the arrival of the most significant bit at a cell, special types of operations are needed. First the digit S_1 of FA1 is a sign digit and must be saved for sign extension. This operation is required for correct addition of two's complement numbers with different lengths. This sign digit is added to the carry C_2 , which is derived during the m.s.b., as well as to the carry PC_i of the next cell. It is clear that a space interval is required between consecutive words, so that a pair of zero bits appear at the cell when the addition of the above digits takes place. This is achieved by inserting a zero bit right after the m.s.b. of each internal variable. The implementation of this step is accomplished by means of the AND gate that appears in the path of the z_{n-1} internal variable. This gate is required only at the first cell but is used in all cells for modularity purposes.

The above mentioned functions are controlled by the appropriate control bit. This bit becomes zero right after the control bit that corresponds to the m.s.b. Thus, while the m.s.b. is at the cell, this bit controls the switch SW1 for the desired sign repetition, while at the same time it prevents carry C_1 from appearing at the input PC_i of the previous cell. In addition it forces the AND gate to give a zero output, thus generating the zero data bit after the m.s.b. of the internal variable. When the m.s.b. moves out of the cell, this zero control bit moves in and truncates carry C_2 . The truncation of C_1 and C_2 is possible because we have assumed that the variables have at least one sign-bit repetition. Otherwise overflow could occur.

At the same time with the sign extension operation, the zero control bit truncates the partial sum of the next pair of variables, by preventing the PS_o output of the $j+1$ cell to go through the switch SW1. This truncation is necessary to avoid mixing of the results from two distinct operations. At the j th-order cell this truncation is of weight 2^{-N-j+1} . Therefore, at the last cell this weight

becomes 2^{-N+1} which means that the least significant bit of the result is truncated. To prevent this from happening, a second zero bit is inserted before the l.s.b. of the next internal variable. This zero bit is generated at the same AND gate that generates the zero bit after the m.s.b., as is clearly shown in Fig. 2. It is not difficult to see that this zero bit is now the l.s.b. and the variables are now considered to have a word length of $N+1$ bits. Consequently, the truncated quantity of the j th cell has a weight of 2^{-N-j} and the last cell truncates the digit of weight 2^{-N} , as is required.

The last cell handles the term,

$$-(z_{n-1} b_1^0 + z_{n-2} b_2^0) 2^1 + x_n.$$

The input x_n is applied to the PC_i input of the last cell. The negative partial term is formed by modifying the last two cells as shown in Fig. 3. It is seen that inverters are inserted at the PS_o output and PC_i input, so that the proper signed-digit additions are performed. From this figure it is seen that three combinations of signed digit addition can occur.

The final result z_n must be rounded instead of being truncated. The truncation of a number is converted to up rounding by adding a quantity of weight 2^{-N} to the number just before the truncation. In the proposed circuit this rounding is performed automatically during the zero bit adjacent to the l.s.b. at the cell before the last. Specifically, the PC_o of the last cell is zero and through the inverter of Fig. 3(a) the PC_i input of the cell before the last becomes the rounding quantity 2^{-N} .

The truncations, the sign-extension and the insertion of the two zero bits are controlled in each cell by a control signal equally delayed as the internal variables z_{n-1} and z_{n-2} . This control signal is always unity and becomes zero at the same time with the zero bit that follows the m.s.b. of z_{n-1} and z_{n-2} .

The overall delay from the time of arrival of the l.s.b. of z_{n-1} and z_{n-2} until the formation of the l.s.b. of the result z_n is equal to $2M-2$ unit delays, where M is the number of digits of the coefficients. Thus, for synchronization purposes this delay must be equal with the data length plus the two zero bits, that is,

$$2M - 2 = N + 2. \tag{6}$$

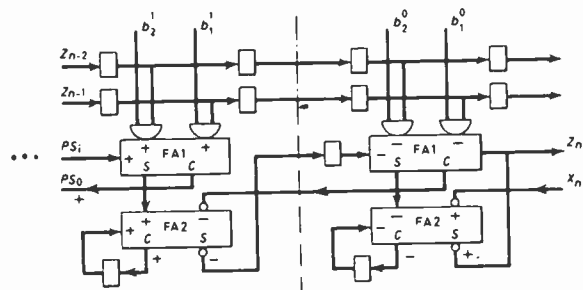


Fig. 3. Simplified diagram showing the signed-digit additions in the last two cells.

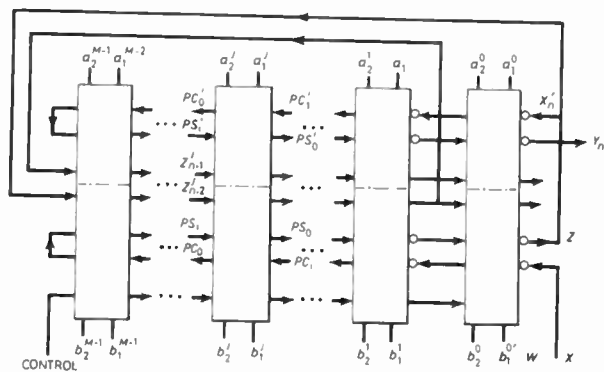


Fig. 4. The complete cellular implementation of the second-order digital filter.

If the delay relationship shown in eqn. (6) is satisfied, the pair z_{n-1} and z_{n-2} can be replaced with the next pair, that is, with z_n and z_{n-1} respectively, as is shown in Fig. 1. The required delay is provided by the cellular circuit itself. Equation (6) does not limit the usefulness of the proposed filter because in most cases the number of bits of the variables is much greater than that of the coefficient bits for overflow protection.

The circuit shown in Fig. 1 without the feedback of the output can be applied for the implementation of the non-recursive part of the filter described by eqn. (2b). In this case the additive input accepts z_n instead of x_n . The control delay line and the delay lines of the internal variables z_{n-1} and z_{n-2} can be common for both circuits. Consequently, significant circuit economy is achieved if these two circuits are combined in one, as shown in Fig. 4, where each stage includes two cells like the one in Fig. 2; one is in the top compartment and the other in the bottom in a mirror image arrangement, with common delays for the variables and the control signal. The cellular circuit of this figure implements the entire second-order digital filter and requires only one control signal.

3 Higher-order Pipeline Digital Filters

In this Section the design of higher-order digital filters is examined using the pipeline second-order section that has been proposed. A higher-order digital filter can be implemented by combining second-order sections. Equations (7a) and (7b) below show the cascade and the parallel form representation of a digital filter using second-order sections. It has been shown that these two forms present the most desirable features in comparison to other possible implementations.

$$H(z) = a_0 \prod_{j=1}^K \frac{a_{2j}z^{-2} + a_{1j}z^{-1} + 1}{b_{2j}z^{-2} + b_{1j}z^{-1} + 1} \quad (7a)$$

$$H(z) = \gamma_0 + \sum_{j=1}^K \frac{\gamma_{1j}z^{-1} + \gamma_{0j}}{b_{2j}z^{-2} + b_{1j}z^{-1} + 1} \quad (7b)$$

In Fig. 5 the implementation of the cascade form is shown. The synchronization delays between the second-order sections permit pipeline operation of the whole filter, which operates at the same speed as each second-order section. In this implementation the output is normalized and a_0 is considered to be equal to unity.

A scaling factor of one-half between successive sections is often used in order to reduce the overflow possibility. This scaling in the proposed filter can be easily achieved if the synchronization delays of the control signal are omitted. Thus, all the cells operate synchronously but each output is applied to the next section with one unit delay. This delay is equivalent with the required multiplication by one-half.

In the case of parallel form realization, each section accepts the sequence as the data input and the output is connected to the additive input of the non-recursive part of the next section, as shown in Fig. 6. As in the previous case, synchronization delays speed up the process and, with appropriate normalization, the constant γ_0 can also be taken to be equal to unity.

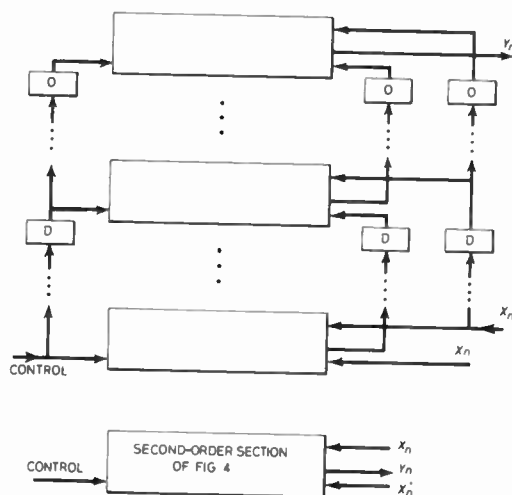


Fig. 5. Cascade implementation of a higher-order digital filter.

Non-recursive digital filters can also be implemented with this technique. The basic building block is again the filter shown in Fig. 1 without the output feedback. The cascade form realization of a non-recursive digital filter is shown in Fig. 7.

4 Performance and Comparisons

The proposed cellular pipeline digital filters have very good performance characteristics in terms of speed of operation and power consumption. The pipeline technique permits high-speed operation which is limited by the propagation delay of one cell. As can be seen from the circuits of Fig. 2 and 4, the maximum propagation delay path is $8\tau_g$, where τ_g is the gate delay. It has been

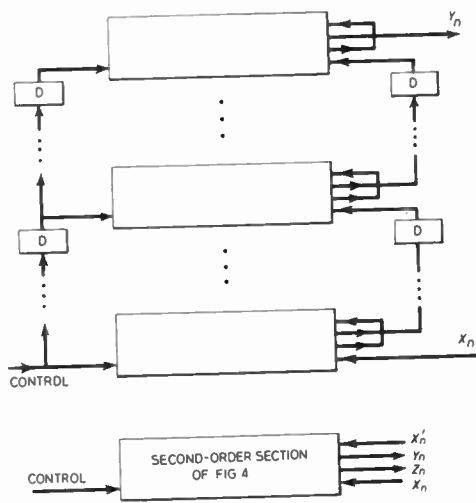


Fig. 6. Parallel implementation of a higher-order digital filter.

assumed that each full adder has been constructed with two level logic. Assuming that the delay flip-flop introduces $2\tau_g$ delay, the maximum data bit rate that can be achieved becomes

$$f_{max} = \frac{1}{10\tau_g} \tag{8}$$

The corresponding maximum bit rate for the digital filters based on the existing pipeline multipliers and De Mori's cellular method are equal to $1/7\tau_g$ and $1/9\tau_g$, respectively.

The circuit complexity of the proposed second-order section is $110M$ gates, where M is the number of digits in the coefficient. The implementation based on the existing pipeline multipliers requires $200M$ gates and De Mori's cellular implementation needs $120M$ gates. In the above estimations it was assumed that each full adder has nine gates and each flip-flop has six gates. In De Mori's case, the sign bit of the variables must be repeated M times for the correct performance of the arithmetic operations.

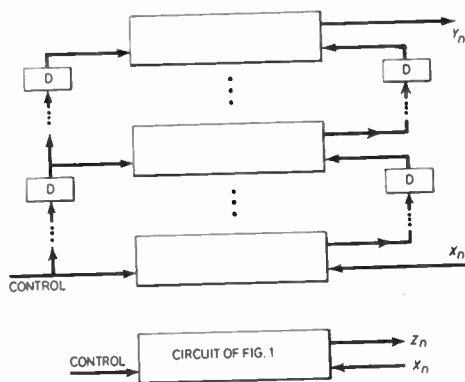


Fig. 7. Cascade implementation of a non-recursive digital filter.

This repetition reduces the efficiency of this filter by a factor approximately equal to $2/3$.

The r.o.m. filters have the same operation speed as the proposed realization. For example, using TTL Schottky, both implementations have a 50 ns delay. The r.o.m. filters, however, have a little more circuit complexity than the proposed cellular pipeline filters. The circuit of a second-order r.o.m. filter consists of a $32M$ r.o.m., M full adders, $2M$ flip-flops, 5 carry look ahead units and $2M$ switches for addition-subtraction control and parallel to serial conversion. The total number of gates required in the above circuit is approximately $130M$.

From the above comparisons, it is clear that the proposed cellular pipeline realization is, for the same speed of operation, better than the existing realizations in terms of complexity and power consumption. The primary advantages, however, are the cascaded cellular structure and the functional completeness of each cell. These features permit the design of filters by using only one component as the building block. In addition, the filter is expandable so that any required accuracy can be achieved, in sharp contrast to the previously proposed realizations that cannot be separated in operationally equivalent sections and, thus, any practical realization has a fixed accuracy. Finally, the proposed second-order section can be used for the construction of higher-order digital filters which process data at the same rate as each sub-section.

Emitter function logic (EFL)⁵ has been proved very efficient for practical realization of the serial pipeline multiplier.^{6,7} Consequently, this logic family can be used for a practical l.s.i. realization of the cellular pipeline filter by building a module consisting of four cells. This chip will need 28 pins and will permit data processing rates of 20 Mbit/s with 300 mW power dissipation.

5 Conclusions

A new approach for the realization of digital filters using the pipeline technique has been presented. The proposed realization yields a simple cascaded cellular structure which provides functional completeness and requires only one control signal. It was shown that this implementation has several advantages over the existing ones. Specifically, it results in a circuit with smaller circuit complexity and lower power consumption. In addition, its cellular structure offers great flexibility in filter design.

The integrated circuit realization of a module with four cells is of major interest, since it permits the construction of a one-component digital filter which is flexible and can be easily adapted to the customer need. We believe that such a component will become a powerful tool, simplifying the digital filter design. The EFL realization of the proposed component, as well as the extension of the serial-pipeline technique to other digital signal processing algorithms, need further study.

42

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