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

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The Journal of the Institution of Electronic and Radio Engineers

C Eng = R Eng or R Eng(Dip) + X

DURING discussions of the Finniston Report it has become evident that many engineers have not realized that the award of the proposed designations R Eng and R Eng(Dip) will not be dependent on completion of a stipulated period of responsible experience but, unlike C Eng, will be made to engineers who have completed an approved basic 'formation package' of education and training. In this respect the recommendations, if accepted, would bring UK registration practice down to the level prevailing in most other EEC countries, where proved performance is not a prerequisite of registration.

The Finniston Report recognizes that the ideal basic 'formation package' for a professional engineer would probably be a 'thick sandwich' course (1 year in industry, followed by 3 years' academic study, followed by a year of post-graduate training). It is only because it is recognized that many smaller companies could not afford to sponsor even the 'training' part of this package that it is proposed that the initial training element of the thick sandwich pattern should be integrated with the academic studies and become the responsibility of the educational establishment. Industry would be responsible for the provision of 'accredited' post-graduate training, which would be so arranged as to constitute a lead-in to the engineer's first professional appointment. In algebraic terms:

B Eng + post-graduate training (Finniston's EA3 and EA4) = R Eng
M Eng + (possibly extended) EA3 and EA4 = R Eng(Dip)

It should not be assumed that the designations R Eng and R Eng(Dip) are immutable: the Committee had some difficulty in finding suitable designations which were not already in use and would, we believe, be happy to consider alternatives. The Report itself recognizes that, as a follow-up to R Eng and R Eng(Dip), a later award which indicates that the holder has completed a period of employment in a responsible post to the satisfaction of his peers will continue to be of value, and urges Institutions to co-operate in this respect by accepting R Eng and R Eng(Dip) as meeting their basic education and training requirements for Corporate membership.

The Institution can see no objection to this in principle, but a possible serious one in practice. If Finniston's proposed British Engineering Authority (BEA) is approved and its establishment leads to the demise of CEI in its present form, the designation C Eng could disappear, and all the costly effort which has gone into the process of securing the general recognition and high esteem which it now possesses would have been wasted. This would be a great pity, particularly since under Clause 14 of our present Royal Charter every Corporate member of the IERE is entitled by Her Majesty's Privy Council authority to 'take and use the name and title of or describe himself as a Chartered Electronic and Radio Engineer'. This fact alone would make designations such as R Eng MIERE, R Eng(Dip) MIERE, R Eng FIERE, and R Eng(Dip) FIERE, something of a nonsense since the Chartered Engineer status of the MIERE and FIERE in every case would indicate a far higher level of general professional standing than the R Eng or R Eng(Dip) which preceded them.

But we do not believe the clock need inevitably be put back to this extent. There is a widespread feeling amongst the large companies in the electronics industry at least, that the designation C Eng must be preserved, and we believe it is similarly valued in other industries and the public service. It is generally agreed that the proposed BEA will not succeed unless it can attract the goodwill of industry and the engineering Institutions: provided it does this, it would surely be possible to persuade Privy

Council to allow the Chartered Institutions which constituted CEI to award the C Eng designation, in addition to their specific designatory letters, to their Corporate members. In the IERE case, only a very minor addition to that highly relevant Clause 14 of our Royal Charter would be needed to give legal effect to this proposal.

It might be reasonably argued that if such a change were effected, adoption of the Finniston proposals had merely substituted the new BEA for CEI as the validating authority for the professional engineer's 'formation package', and such a trivial change did not justify all the effort involved in bringing it about. But to argue that way would be to miss Finniston's main point, which is that the engineering industry and profession and their importance to the national economy have been undervalued and underpublicized for so long that the establishment of a 'champion of change' in the form of an Authority backed by and with direct access to Government offers the best if not the only means of beginning the process of putting things right. Which is why the IERE is so anxious to ensure that the Report does not founder for want of sensible development of its incomplete ideas concerning the registration and full career formation of the nation's current and future resources of engineering talent.

S.M.D.

Notice to all Corporate Members of the Institution

NOMINATIONS FOR ELECTION TO THE 1980–81 COUNCIL OF THE INSTITUTION

In accordance with Bye-law 49, the Council has nominated the following members for election at the Annual General Meeting to be held in London on Thursday, 23rd October 1980:

President

For Election: J. Powell, T.D., B.Sc., M.Sc.

Vice-Presidents

Under Bye-law 46, all Vice-Presidents retire each year but may be re-elected provided they do not serve thereby for more than three years in succession.

For Re-election: H. E. Drew, C.B.; Professor J. R. James, B.Sc., Ph.D.; Brigadier R. W. A. Lonsdale, B.Sc.; P. K. Patwardhan, M.Sc., Ph.D.; S. J. H. Stevens, B.Sc.(Eng.)

Honorary Treasurer

For Re-election: S. R. Wilkins

Ordinary Members of Council

Under Bye-law 48, Ordinary Members of Council are elected for three years and may not hold that office for more than three years in succession.

MEMBER

The following must retire: A. F. Dyson, Dip.El.

For Election: J. J. Jarrett

The remaining members of Council will continue to serve with the period of office laid down in Bye-law 48.

Within twenty-eight days after the publication of the names of the persons nominated by the Council for the vacancies about to occur any ten or more Corporate Members may nominate any one other duty qualified person to fill any of these vacancies by causing to be delivered to the Secretary a nomination in writing signed by them together with the written consent of the person nominated undertaking to accept office if elected, but each nominator shall be debarred from nominating any other person for the same vacancy (Bye-law 50).

By Order of the Council
S. M. DAVIDSON
Secretary

5th June 1980

ANNOUNCEMENTS

THE INSTITUTION'S PREMIUMS

Additions and Changes Announced by Council

Each year the Council awards premiums for outstanding papers published in the past volume of the Journal and since the first awards in 1946 these have been added to in number and scope to take account of the expansion of radio and electronic engineering as seen in the greatly increased coverage of the Journal. There are therefore now 20 premiums available annually and these are listed from time to time in the Journal (see, for instance, Appendix 8 to the last Annual Report published in September 1979).

At its last meeting Council approved a new award, which is to be known as the **Lord Mountbatten Premium**, and will commemorate the Charter President who contributed so greatly to this Institution through 44 years of active membership. The Premium will be for the outstanding paper on the engineering applications of electronics or radio and its value will be £100.

The terms of award of the **Clerk Maxwell Premium** are being modified to complement the Mountbatten Premium by recognizing an outstanding paper on the science of electronics or radio and its value now will also be £100.

The next most meritorious papers in each of these categories will now be recognized by two other existing Premiums of long standing, respectively the **Marconi Premium** for an engineering paper, and the **Heinrich Hertz Premium** for a scientific paper. Both these premiums will have the value of £50.

The opportunity to modify certain other premiums has been taken at the same time and, at the suggestion of the Papers Committee, Council has gladly accepted changes which recognize the contributions made to the Institution by two senior members, both, happily, still living.

In 1959 a premium was endowed by Associated-Rediffusion, through its then managing director, Mr Paul Adorian, for an outstanding paper on advances in the techniques of television broadcasting. Over the years the terms of award and title of this premium have changed and in recent years the Rediffusion Television Premium has been given for outstanding papers on communication engineering. From this year, however, the directors of Rediffusion Television Ltd have generously agreed that this award should bear the name of their former colleague and Past President of the Institution, and thus be known as the **Paul Adorian Premium.** Its terms of award will continue to be the same, as will its value of £50.

The second change which takes effect from this year is the extension of the terms of reference of the **Hugh Brennan Premium.** This was endowed by Mr Brennan over twenty years ago when he was Chairman of the North Eastern Section and its objective was to recognize outstanding papers published in the Journal which had first been read before that Section. Now, however, papers read before any Local Section and subsequently published in the Journal will be eligible for consideration for this premium which has the value of £25.

Biographies of both these Fellows of the Institution whose names are borne by the premiums will be published in a future issue of the Journal.

Tenth Anniversary of SRS

The Systems Reliability Service, a unit of the United Kingdom Atomic Energy Authority, was inaugurated on 1st April 1970 to make generally available techniques and accumulated experience in the field of systems reliability assessment that the Authority's Safety and Reliability Directorate had acquired in the course of many years' work on safety studies of nuclear reactor systems and chemical plant complexes. SRS was planned to be a commercially-based unit which could make available to industry and government departments a consultancy and information service on reliability assessment and analysis.

Proposals to incorporate SRS in a National Centre of Systems Reliability (NCSR) to be operated by the Authority were approved by the Department of Industry with effect from 1st April 1974. Thus the commercial project and data bank services of SRS were complemented within a National Centre by a Reliability Technology Research Unit (RTRU) which undertakes research and development of systems reliability technology in the national interest.

Both the research and commercial arms of NCSR have evolved formal links with Universities and industrial organizations. A Reliability Technology Research Forum meets twice a year under the auspices of the RTRU; it formulates and links research activities, and organizes seminars and conferences on the subject of reliability technology. Some 80 Companies, with an interest in Reliability Engineering techniques, have joined SRS as 'Associate Members'. Membership entitles subscribers to enjoy the benefits of information dissemination; to participate in the Data Bank facilities and to obtain favourable terms for consultancy and project work which can be undertaken on a confidential basis.

NCSR sponsored the first National Conference on Reliability in September 1977 at Nottingham University. A second Conference, jointly sponsored by NCSR and the Institute of Quality Assurance, was held at the NEC Birmingham in March 1979; and the third in the series is to be held at the National Exhibition Centre, Metropole Hotel, Birmingham, on 29th April to 1st May 1981, on the theme 'Reliability—risks, resources, rewards'.

Corrections

Postal delays prevented the following corrections by the authors from being made to the paper 'Resonant modes in re-entrant cavities' which was published in *The Radio and Electronic Engineer*, vol. 50, no. 3, March 1980.

Page 114: The following sentence is to be added to the caption of Fig. 3:

(The maximum error in measuring the resonant frequency is ± 5 MHz)

Page 115: Left-hand side of equation (1) should be C_{eff}/ε_0 instead of C_{eff}.
In the right-hand side of equation (3) read μ_0 in place of k_0.

Page 116: x_1 and x_2 should be replaced by capital letters X_1 and X_2 respectively in equations (6), (7) and (8).

Applicants for Election and Transfer

THE MEMBERSHIP COMMITTEE at its meetings on 25th March and 16th April 1980 recommended to the Council the election and transfer of the following condidates. In accordance with Bye-law 23, the Council has directed that the names of the following candidates shall be published under the grade of membership to which election or transfer is proposed by the Council. Any communication from Corporate Members concerning the proposed elections must be addressed by letter to the Secretary within twenty-eight days after publication of these details.

March Meeting *(Membership Approval List No. 271)*

GREAT BRITAIN AND IRELAND

CORPORATE MEMBERS

Transfer from Member to Fellow

CLARKE, George David. *Finchampstead, Berkshire.*
EVANS, Christopher David Ian. *Portland, Dorset.*

Direct Election to Fellow

ALEXANDER, Michael Edward. *Crawley, Sussex.*

Transfer from Graduate to Member

ALDRIDGE, Peter John. *Apperley, Gloucestershire.*
COOPER, Paul Anthony. *Brentwood, Essex.*
DUCKLING, Christopher. *Lightwater, Surrey.*
GILES, John Donald. *Woodley, Reading, Berkshire.*
MATTOCKS, Malcolm Stanley. *Maidenhead, Berkshire.*
RAMNARINE, Ramnarace. *London.*
UPTON, David Maurice. *Sittingbourne, Kent.*

Transfer from Associate Member to Member

LUDLOW, Richard Thomas George. *Canterbury, Kent.*

Direct Election to Member

COSGROVE, Ian Thomas. *Beeston Rylands, Nottinghamshire.*
FINNIE, James Smith. *Redhill, Surrey.*
SAMPSON, Michael Terence. *Leigh-on-Sea, Essex.*
SMITH, Martin Christopher. *London.*
STOCK, Nicholas Timothy. *Thornbury, Bristol, Avon.*

OVERSEAS

CORPORATE MEMBERS

Transfer from Member to Fellow

IFIDON, Rowland Oke. *Benin City, Nigeria.*

Transfer from Graduate to Member

FATOYE, Ezekiel Olasunmoye. *Lagos, Nigeria.*
FOYE, Samuel Olatunji. *Jebba, Kwara State, Nigeria.*

Transfer from Associate Member to Member

LEO, Teng Yong. *Singapore.*

Transfer from Associate to Member

MUPESO, Edward Chitembwe. *Lusaka, Zambia.*

Direct Election to Member

HORLER, Neil Anthony. *Kingston, Ontario.*
OFULUE, Jacob Onyemaonyeolu. *Auchi, Nigeria.*

April Meeting *(Membership Approval List No. 272)*

GREAT BRITAIN AND IRELAND

CORPORATE MEMBERS

Direct Election to Fellow

HINDER, Richard Alan. *Croydon, Surrey.*

Transfer from Graduate to Member

FRASER, Ian Edward. *Newton-Le-Willows, Merseyside.*
HUSSAIN, Manzoor. *Watford, Hertfordshire.*
OWENS, Kenneth John. *Melksham, Wiltshire.*
PEARSON, John Edwin. *Welwyn, Hertfordshire.*
RITCHIE-O'NIELL, John. *Reading, Berkshire.*
ROBERTS, Gareth Silyn. *Horndean, Hants.*
STEWART, Peter Jackson. *Hilton, Cambridgeshire.*
STRINGER, Howard Frederich. *Chertsey, Surrey.*
WEST, Michael Graham. *West Wickham, Kent.*

Direct Election to Member

CUMMINGS, Robrt Conrad. *Pontardulais, W. Glamorgan.*
DURHAM, David Benedict L. *Malvern, Worcester.*
FONSECA GUERRA, Juan Marcial. *Birmingham.*
FRANCIS, Sidney Edward. *Godalming, Surrey.*
HENDERSON, Stanley. *Carlisle, Cumbria.*
PEMBROKE, Charles. *Wokingham, Berkshire.*
RADCLIFFE, Charles Edward. *Swansea.*

NON-CORPORATE MEMBERS

Transfer from Student to Graduate

CHU, Joseph Hung Ming. *London.*

Direct Election to Graduate

ADJEI, Christopher Mensah. *London.*
AHMAD, Shakil. *Poole, Dorset.*
LYDIARD, Peter. *Crawley, West Sussex.*
CHOI, Yat Man. *Cardiff.*

Transfer from Associate to Associate Member

GODFREY, Christopher. *Cardiff.*

Direct Election to Associate Member

BELL, Peter Andrew. *Newcastle-upon-Tyne.*
BURGE, Walter Thomas. *Warrington, Cheshire.*
TIPPIN, Leslie Gordon. *West Wickham, Kent.*

Direct Election to Associate

FOX, Barry Howard. *London.*
GAFFNEY, Leonard. *Newcastle, Galway, Ireland.*
NEWBY, Martin John. *Buckhurst Hill, Essex.*

Direct Election to Student

BARKER, David. *Skipton, N. Yorkshire.*
ESP, David Gary. *Walton-on-Thames, Surrey.*
HOOLEY, John Albert. *Burnham-on-Sea, Somerset.*
HUNTLEY-HAWKINS, Peter Anthony. *Woking, Surrey.*
ISHAQ, Syed Zahid. *Coventry.*
WORRALL, Geoffrey Peter Adrain. *Solihull, Warwickshire.*

OVERSEAS

CORPORATE MEMBERS

Transfer from Graduate to Member

ABEYSUNDARA, Bandula Wijeratne. *Moratuwa, Sri Lanka.*

Direct Election to Member

CHEUNG, Kin Yin. *Causeway Bay, Hong Kong.*

NON-CORPORATE MEMBERS

Transfer from Student to Graduate

HO, King Tak. *Kowloon, Hong Kong.*
LIU, Yan Wing. *Kowloon, Hong Kong.*

Direct Election to Graduate

YEUNG, Man Wah David. *Hong Kong.*

Transfer from Student to Associate Member

LAM, Weng Keen. *Singapore.*

Direct Election to Associate Member

SIVABALAN, Selliah. *Abu Dhabi, U.A.E.*
WONG, Sik Hung. *Hong Kong.*
YUEN, Wah Tit. *Hong Kong.*

Direct Election to Associate

ANANDA, Jeeva Singgaram. *Kuala Lumpur, Malaysia.*
HINTON, Martin Roy. *Riyadh, Saudi Arabia.*

Direct Election to Student

CHAK, Ping Yuen. *Tsuen Wan, Hong Kong.*
CHAN, Hak Chun. *Kowloon, Hong Kong.*
CHAN, Siu-Wai. *Kowloon, Hong Kong.*
CHAN, Kai Chong. *Kowloon, Hong Kong.*
CHAN, Sung Fai. *Hong Kong.*
CHAN, Wai-Keung. *Kowloon, Hong Kong.*
CHAN, Wing-Kwong. *Kowloon, Hong Kong.*
CHAN, Yau Ting. *Quarry Bay, Hong Kong.*
CHENG, Wah-Chi. *Hong Kong.*
CHEUNG, Ming Fat. *Aberdeen, Hong Kong.*
CHEW, Wai Kwok. *Hong Kong.*
CHIU, Karm Ting. *Hong Kong.*
CHOI, Chun-Wai. *Kowloon, Hong Kong.*
FAN, Che Fung. *Shaukiwan, Hong Kong.*
HO, Kum Chuen. *Singapore.*
JAP, Jone Young. *Singapore.*
LAM, Chi Pang. *Hong Kong.*
LAM, Ip Shing. *Sai Ying Poon, Hong Kong.*
LAM, Sing-Ho Michael. *Hong Kong.*
LEUNG, Kwok Leung Colin. *Kowloon, Hong Kong.*
LEUNG, Kei Yuk. *Kowloon, Hong Kong.*
LING, Kwok Leung. *Kowloon, Hong Kong.*
LUI, Tack. *Aberdeen, Hong Kong.*
MA, Mor-Hoi. *Shek Tong Tsui, Hong Kong.*
MUI, Heung On. *Hong Kong.*
OTUN, Abiodun. *Lagos, Nigeria.*
POON, Hung Leung. *Shaukiwan, Hong Kong.*
SHU, Cheong Clement Johnny. *Shaukiwan, Hong Kong.*
TSANG, Hen-Loon Raymond. *Tsuen Wan, Hong Kong.*
TSUI, Chi Wah. *Shantin, Hong Kong.*
TUNG, Tsang Kow. *Kowloon, Hong Kong.*
WAN, Ping Wing. *Shaukiwan, Hong Kong.*
WONG, Din Tat. *Hong Kong.*
WONG, Kai Chi. *Aberdeen, Hong Kong.*
WONG, Kwok Ting. *Kowloon, Hong Kong.*
WUT, Kai Cheung. *Cheung Chau, Hong Kong.*
YEUNG, Kam Yuen. *Kowloon, Hong Kong.*
YIP, Kai Shung. *Kennedy Town, Hong Kong.*
YU, Yung Shing. *Causeway Bay, Hong Kong.*
YUI, Tat Man. *North Point, Hong Kong.*

Members' Appointments

CORPORATE MEMBERS

Captain K. A. W. Pilgrim, O.B.E., RN(Ret.) (Fellow 1957, Member 1949, Graduate 1946) has been appointed Executive Director, Sales, for the Hatfield Division of British Aerospace Dynamics Group. Captain Pilgrim served in the Royal Navy from 1941 until 1973, his last appointment being on the staff of Flag Officer Sea Training as Chief Staff Officer (Technical) and Captain of the Portland Naval Base. He was appointed OBE in 1962. From 1973 to 1975 he was with Hawker Siddeley Dynamics, initially as Project Manager in the Under-water-to-Surface Guided Weapon Division, subsequently becoming Divisional Manager.

He has since held appointments as Divisional Manager of the Guided Weapon Support Division and of the Air Strike Weapons Division. Captain Pilgrim served on the Institution's Technical Committee from 1956 to 1965 and on the Membership Committee from 1965 to 1968.

Lt-Col. D. A. Wilcox, M.Sc., RA(Ret.) (Fellow 1977, Member 1970, Graduate 1963) who joined British Aerospace Dynamics Group, Filton, in 1978 as a Senior Systems Engineer, has been appointed Head of Systems Studies and Integration Engineering Department. Prior to his retirement from the Army, Lt-Col Wilcox was on the staff of the Royal Military College of Science.

T. Boucher (Member 1972, Graduate 1967) has been appointed Head of Advanced Projects (Avionics) of Base Ten Systems. Mr Boucher has been with the company since 1973, latterly as Chief Engineer. Prior to joining Base Ten, he was with British Aerospace at Filton for eleven years.

K. W. Chan, M.Sc. (Member 1979, Graduate 1971) is now on the scientific staff of Bell Northern in Ottawa. Before going to Canada he was Senior Project Engineer with Transmitton, at Ashby de la Zouch, Leicestershire and he previously held appointments with Plessey Telecommunications, London Transport Executive and ST&C.

Gp Capt. J. P. Downes, B.Sc., RAF (Member 1966, Graduate 1961) has taken up an appointment as Command Aircraft Electrical Engineer, HQ RAF Strike Command. He was Station Commander RAF Halton from 1977 to 1979.

B. J. Goldsmith (Member 1973, Graduate 1967) who has been with International Aeradio since 1973, has been appointed Systems Manager, Data Communications Division.

B. G. Hole (Member 1959, Graduate 1956) has been appointed Project Manager Fire Control, Gun and Seacat System Post Design at the Admiralty Surface Weapons Establishment, Portsdown. He has been with the Scientific Civil Service since 1951 and before taking up his present post was Project Manager, Cymbeline Army Weapons Locating Radar, in the Ministry of Defence Procurement Executive.

P. M. Holker (Member 1969, Graduate 1966) who has been with Marconi Communication Systems since 1966, holding positions as Sales Manager in South Africa, the Middle East and, for the past five years, at Chelmsford, is now at Hatfield Instruments as Sales Manager.

M. E. Jones, B.A. (Member 1972, Graduate 1969) who was Chief Engineer with British Aerospace in Abu Dhabi, has returned to the United Kingdom and taken up an appointment as Project Manager—Tracked Rapier, with the company in Stevenage.

D. C. McLean (Member 1973, Graduate 1969) is now an Electronics Specialist in the Chemistry Department at the University of Edinburgh. He was previously for ten years Technical Officer, Electronics, at the University of Stirling.

Lt Cdr H. C. Parker, RN (Member 1974, Graduate 1969) has taken up the post of Weapon Project Officer of the Naval Air Technical Evaluation Centre at HMS *Daedalus* following his promotion.

Lt Cdr R. M. Prynn, RN (Member 1973, Graduate 1969) has been appointed Captain Weapons Trials as team leader of a SONARC trials team, following a two-year posting as Weapons Engineer Officer in HMS *Ariadne*.

C. E. Ramsbottom (Member 1959) who has been on the staff of the Department of Electrical and Electronic Engineering at Wolverhampton Polytechnic since 1963, has been presented with a Premium Award by the Midland Centre of the Royal Television Society, for giving two outstanding lectures on the historical background to television and associated technology. Mr Ramsbottom has served for a number of years on the Committee of the West Midland Section, offices held including that of Honorary Secretary.

R. K. Robertson (Member 1963, Graduate 1959) who joined the Marconi Company as an Installation Engineer in 1959 following service in the Royal Air Force, has recently been appointed Manager of the Radio and Line Division of Marconi Communication Systems.

K. K. Saddi (Member 1970, Graduate 1966) is now a Senior Development Engineer with the Zambia Broadcasting Services, Lusaka: he has been with the organization since 1967.

P. Sethi, B.Sc. (Member 1973, Graduate 1968) has joined Siemens, Munich, as a Senior Project Engineer in the M.O.S. Development Department. He was previously with Valvo, Hamburg, for eleven years.

J. J. Trodden (Member 1974, Graduate 1971) has left the Royal Navy and joined Cossor Electronics, Harlow, as Trials Co-ordinator.

Sqn Ldr T. Winchcombe, M.B.E., RAF(Ret.) (Member 1969, Graduate 1956) has taken up an appointment with the Ministry of Defence Procurement Executive. Prior to retirement he was on the staff of the Senior Radio Servicing Engineer at RAF Support Command Signals Headquarters.

NON-CORPORATE MEMBERS

Flt Lt C. G. Brown, RAF (Graduate 1972) has been appointed Electrical Engineer 1B3, RAF Headquarters, Germany: he was previously Officer Commanding Ground Radio Flight at RAF Kinloss.

M. H. Thurlow (Graduate 1970) is Head of Radar Engineering and Launcher Integration with the Guided Weapons Division of British Aerospace at Bristol. Before joining the company in 1971, he was an electronics planning engineer at the Division's Stevenage location. He was for 10 years in the Royal Air Force retiring as a Sergeant concerned with ground radar equipment.

R. Idowu Adu, B.Sc., Dip.El., M.Sc. (Graduate 1979) has returned to Nigeria to resume duty with the Ministry of Communications, Posts and Telecommunications Department as a Transmission Planning Engineer after successfully completing an M.Sc. degree in telecommunications systems at the University of Essex. He has been on the staff of Post and Telecommunications Department since 1966.

S. A. Idowu, B.Sc. (Graduate 1979) is a Plant Officer with the Ministry of Communications, P&T Department, Lagos. He obtained a CNAA degree in 1978 following study at the Polytechnic of North London.

It is regretted that the career details of Mr R. Idowu Adu were attributed to Mr S. A. Idowu in the note in the January/February 1980 issue and apologies have been extended to both members for any embarrassment this may have caused.

P. A. Brown, B.A. (Associate Member 1973, Associate 1972) who is with the North Western Electricity Board as First Engineer (Telecommunications) for the Lakeland Area, has been granted a B.A. degree by the Open University.

Capt. R. R. Holmes, B.Sc., R Signals (Graduate 1979) has been appointed Second-in-Command, Task Force H Signal Squadron. He has recently completed a course on design of management information systems at the Royal Military College of Science, Shrivenham.

E. F. Lever, M.Eng. (Graduate 1973) who was a Principal Engineer with Racal Mobilcal, is now Assistant Chief Engineer.

Sqn Ldr R. A. Scott, RAF (Associate Member 1975) has been posted to RAF Chivenor as Officer Commanding Electrical Engineering Squadron, following a staff appointment at HQ Strike Command RAF.

J. F. Lynch (Associate Member 1973, Associate 1972) has taken up an appointment as Electronic Technician (Grade 5) in the Department of Electronic Science and Telecommunications at the University of Strathclyde.

Lt P. A. Trott, RN (Graduate 1979) is now Deputy Weapons Engineer Officer at Fraser Gunnery Range in Portsmouth. His previous appointment was Deputy Weapon Engineer Officer, HMS *Nubian*.

Sub Lt K. P. White, B.Eng., RN (Graduate 1979) who joined the Royal Navy a year ago on completion of his degree course at the University of Wales Institute of Science and Technology, has now entered the Royal Naval Engineering College, Manadon.

R. Williamson (Associate 1955) has taken up an appointment as Senior Project Administrator with Sperry Gyroscope in Bracknell. He was previously a Senior Scientific Officer with the Natural Environment Research Council.

Profile of European Semiconductor Manufacturers

It is a sign of the times that half (10 out of 20) of the leading European Semiconductor manufacturers, as identified by market share in a new Mackintosh Report, are 'off-shore' operations owned by American or Japanese multinationals. The market leader, Philips, with 19% of the total European semiconductor market, is a European (Dutch) concern, although ironically their market share is boosted by deliveries from their American-based production facilities (Signetics).

In the report 'Profile of European Semiconductor Manufacturers' it is shown that of a total European market for i.c.s and discretes estimated to be worth $3648M by 1982 (up from $2644M in 1979 and $2031M in 1977) indigenous European manufacturers still remain some 52%.

The reasons for this encouraging situation are several. For instance, 'Profile' shows that indigenous companies have increased their production facilities and consequently the total market value of their devices has increased. To counter, 'overseas' manufacturers are no longer treating Europe as just another off-shore facility, but are also increasing their manufacturing plants, such that in some cases they are larger than the parent company's home plant.

The main feature of the 'Profile of European Semiconductor Manufacturers' is a directory of the major European semiconductor manufacturing plants, giving the location, plant size, number of employees, and estimated European Semiconductor Sales for over 50 facilities. The directory also indicates the process capabilities of these plants and itemizes the technology which makes custom design economical and lists the design houses. A section dealing with custom designed i.c.s discusses a number of ways a prospective user might embark on such a venture.

'Profile of European Semiconductor Manufacturers' is a Mackintosh Publications Report of 98 A4 pp, price £150 (UK) or $375 (Overseas). Further information from Mackintosh Publications Limited, Mackintosh House, Napier Road, Luton, England, LU1 1RG. Tel: (0582) 417438.

Standard Frequency Transmissions

(Communication from the National Physical Laboratory)

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

MARCH 1980	MSF 60 kHz	GBR 16 kHz	Droitwich 200 kHz
1	−2·9	12·2	23·7
2	−2·6	13·0	23·4
3	−2·8	13·0	23·2
4	−2·8	15·2	23·0
5	−2·8	15·5	22·9
6	−2·8	15·0	22·8
7	−2·8	14·5	22·7
8	−2·7	*	22·6
9	−2·5	15·3	22·4
10	−2·4	16·0	22·3
11	−2·3	16·6	22·1
12	−2·5	16·0	22·0
13	−2·5	16·8	21·9
14	−2·3	16·5	21·7
15	−2·2	16·6	21·5
16	−2·1	16·5	21·4
17	−2·0	16·2	21·3
18	−2·0	18·5	21·3
19	−1·9	18·7	21·2
20	−1·7	19·0	21·1
21	−1·5	19·0	21·0
22	−1·2	18·8	20·9
23	−1·3	19·7	20·9
24	−1·2	18·7	20·9
25	−1·1	18·3	20·9
26	−1·0	18·0	20·9
27	−1·2	18·0	21·0
28	−1·2	18·1	21·0
29	−1·0	18·9	21·0
30	−1·0	18·4	20·9
31	−1·0	18·7	20·9

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 10^{11} per day.

Colloquium Report

Analogue Filters

Organized by the IERE Components and Circuits Group and held in London on 6th November 1979.

The colloquium on Analogue Filters contained nine papers covering six different types of filter: passive lumped, active lumped, modulation, microwave, crystal and s.a.w.; a contribution on a seventh type (c.c.d. filters) had to be cancelled at very short notice and was replaced by a general discussion. Owing to the large number of topics considered, most of the talks were partly tutorial for the benefit of the majority of those present who were expected to be non-specialists in any given subject area. Before the start, the chairman, Dr J. K. Stevenson, drew attention to the educational value of colloquia, which should not be treated simply as small-scale conferences, and he encouraged the audience to ask elementary questions on unfamiliar topics, in addition to the usual technical questions on more familiar subjects. An informal atmosphere was easy to establish in surroundings as comfortable and relaxed as those in the lecture theatre at the Royal Institution.

The morning session opened with a paper by Dr L. F. Lind (*University of Essex*) entitled 'Cascade Synthesis of Finite Transmission Zero Networks'. The cascade synthesis method involves multiplying the transmission matrix for a complete network by the inverse matrix for a small section which is to be removed. After performing the multiplication, the value of the element (or elements) in the small section is found by the requirement that the resulting matrix should be of lower degree. The method is applied repeatedly, with elements extracted from either end, until all the elements are found. This paper extends a recent paper by the author which was restricted to all-pole ladder structures. The most important advantage of the given method over earlier methods is the considerable improvement in element accuracy.

The second paper by Dr J. N. Torry (*Portsmouth Polytechnic*) was entitled 'Some Sensitivity Results from a Ladder Analysis Program'. The chosen analysis method consisted of finding a transfer function as a rational function in s, substituting values of frequency, and determining the overall transmission matrix. The sensitivity of a network relates the change in the response to the change in the element values. Ladder network sensitivity was determined in various ways. The expressions considered were difference sensitivity, relative difference sensitivity, slope normalized sensitivity (as defined recently by Fidler and Nightingale), and summed sensitivity (a combination of the remainder). The next paper, given at short notice by Dr J. K. Stevenson (*Polytechnic of the South Bank*) was entitled 'The Use of Nullators and Norators in the Design and Analysis of Active Circuits'. The inputs and outputs of infinite-gain controlled sources, which are approximated using operational amplifiers or transistors, are totally independent and may be treated as separate elements, termed nullators and norators. The voltages and currents in a circuit are defined completely by the nullators and the 'conventional' elements, i.e. the norators are sources which only supply what is requested elsewhere. This property can be used to simplify an analysis of active circuits and to help design new circuits. Examples were given.

The following paper by Dr D. G. Haigh (*Imperial College*) and Dr J. K. Stevenson, entitled 'Generalization of some Published Transformations for Active and Passive Networks', was delivered by Dr Haigh. Some recent tranformations by Fliege, Hilberman, and Palomero-Garcia can be applied to networks to produce new networks having the same natural modes, i.e. the same denominator roots of the voltage transfer ratio. The existing transformations were described and shown to be special cases of a general transformation; some novel transformations were then presented as alternative special cases of the general transformation. All of these transformations were illustrated with examples.

The final paper in the morning session by the late Dr W. Saraga—an appreciation follows this report—and Mr M. Zyoute (*Imperial College*) was entitled 'New Active-RC Bode-type Variable Equalizers'. The talk was delivered by Dr Saraga who began by introducing some general concepts relevant to the performance and design of variable loss-frequency equalizers, and in particular the concept of Bode-type variable equalizers. This was followed by a comparison of some recently published approaches by Brglez, Saraga and Zyoute to the active RC realization of Bode-type variable equalizers. The most recent results by the authors, both theoretical and practical, in the design of such equalizers were then presented.

The afternoon session began with a paper by Dr B. G. Pain (*Polytechnic of the South Bank*) entitled 'A Three-Dimensional Review of Modulation Filters'. Modulation filters are lumped linear filters containing multiplier-modulators, and they enable programmable high-Q inductorless networks to be realized which are suitable for microelectronic realization. Modulation filters were reviewed with the aid of a graphical description which was introduced by considering the Weaver modulator. By replacing the multipliers by on-off switches, a sampled-data filter is obtained. The graphical approach indicates additional responses and aliasing considerations without the need for precise mathematical modelling. It also helps to explain the recovery of a periodic signal from noise. A few anomalies which arise from different switching methods were discussed.

The second session ended with a paper by Dr M. I. Sobhy (*University of Kent*) entitled 'Computer-Aided Design of Microwave Filters'. State-space techniques were used to describe a network, and topological methods were used to formulate the state and output equations. There were no restrictions on the network topology and element interconnections. The circuits could contain lumped elements, transmission lines with and without coupling, and any number of sources. Since frequency domain equations may be easily derived from time domain equations, both time and frequency domain analysis synthesis could be carried out. Network synthesis was achieved by iterative minimization of the error between the desired and initial responses.

The final session, which was devoted to electro-acoustic filters, started with a paper by R. C. Peach (*GEC Hirst Research Centre*) entitled 'Surface Acoustic Wave Filters—Their Present Capabilities and Future Prospects'. Surface acoustic wave (s.a.w.) filters are gaining in popularity due to the design flexibility, small size, and the minimal amount of alignment required. The amplitude and phase characteristics can be tailored independently. At present, centre frequencies of between 10 MHz and 1·5 GHz are practicable, with bandwidths from 200 kHz to 40% of the centre frequency. Conventional s.a.w. structures suffer from a spurious signal called the triple transit response, which is usually suppressed by increasing the insertion loss to above 20dB. Recent work by the author and others has shown that this response can now be suppressed by using a more complex transducer structure.

The concluding paper, entitled 'Crystal Filter Realization' was by Mr R. J. Beattie (*Cathodeon Crystals*). The simple equivalent circuit for a quartz crystal is a lossy series resonant circuit in parallel with a capacitor. Component limitations suggest particular topologies, especially the lattice which is usually realized in semi-lattice form. Different realizations of narrow-band 8-crystal filters with attenuation poles at finite frequencies were considered. An N8 design with four cascaded lattices has a good stopband but a large range of inductance values and a high impedance level. Two identical N4 single lattices in cascade produce a poorer theoretical characteristic and often require central matching pads. An N8 design with two cascaded lattices is recommended.

It is hoped that another colloquium on Analogue Filters will be held after a further two years, possibly in November 1981. With this in mind, the chairman invited the audience to suggest possible improvements. The resulting comments, which will be made known to future authors, indicated that a large proportion of the participants would like the talks to contain a more elementary introduction and an increased practical content. Although improvements will continue to be sought, the wide range of questions at different technical levels throughout the colloquium suggested that the present event had been very worthwhile and had provided benefit to both specialists and beginners.

J. K. STEVENSON

Wolja Saraga—An Appreciation

It is with much regret that I have to report the passing of Dr Wolja Saraga of Imperial College on 15th February 1980, after a short illness.

For those who knew him, Wolja will never die—his influence was profound and will live on, especially in his colleagues and pupils, and through them future generations of electrical engineers. Wolja was an authority on circuit theory, particularly active filters. For many decades, he has been a prolific author of papers to journals, conferences and colloquia, including the two IERE colloquia on Analogue Filters; in 1952, when he was working in industry, he presented a paper at an Institution meeting on 'An aerial analogue computer' which subsequently gained him and his co-authors the Clerk Maxwell Premium. As a researcher, Wolja sought simple solutions and was concerned with clarity of expression. As a teacher, he displayed unlimited patience and consideration. To all, he was a kindly friend with a wonderful sense of humour. Wolja will be missed very much. J.K.S.

The Status of the METEOSAT System

Meteosat 1 continued to function very well in 1979, its second year in orbit. The satellite on-line availability for the two years was 99·5% of its theoretical maximum. In addition, the satellite operated faultlessly in its data dissemination and data collection roles.

Spurious switchings of on-board units continued to occur and required the reconfiguration of the satellite. These phenomena are attributed to electrostatic discharges between isolated surfaces. since the end of November 1979, the spacecraft has had difficulties with its main power supply control unit which has interupted the Earth imaging and dissemination missions. The data collection mission is still functioning.

The three missions supported by *Meteosat* are:

The Earth Imaging Mission. The satellite has transmitted more than 63,000 images of excellent quality, taken in three spectral bands: the thermal infra-red, the visible, and the water vapour absorption band.

The Dissemination Mission. The performances of the Darmstadt Processing Center have been continuously improved during 1978 and 1979 and the following monthly status has been achieved:

dissemination of more than 13,000 image formats to more than 100 user's stations;

calculation of 30,000 wind vectors and 56,000 sea surface temperatures;

archiving of 7,000 products on magnetic tapes, including 3,000 images with a 92% coverage, and archiving of 2,000 image photographic negatives.

The Data Collection Mission. This mission handles the data coming from 40 data collection platforms (DCPs) and disseminates them back to their users.

The *Meteosat* system is used by a large variety of users, meteorological, hydrological, oceanological or seismological services, as well as radio amateurs receiving the products disseminated towards a dozen PDUSs and more than 100 SDUSs. A large number of users have met several times to exchange experience on meteorological satellites and, in particular, on *Meteosat* utilization.

The preparation of *Meteosat 2* which is due to be launched in September 1980 on the third experimental launch of the European *Ariane*, is being actively undertaken.

Work has started on the preparation of a *Meteosat* operational programme. The launch of a first satellite is foreseen in early 1984 and discussions are taking place with the meteorological services of the participating member states on the general requirements for a system lasting for 10 years starting with the above launch.

This brief report sets the scene for the paper 'A signal demultiplexer for primary data user stations' by R. J. H. Brush of the University of Dundee, which is printed on pages 297–306 of this issue. The information has been obtained from the *Meteosat Journal* published by the European Space Agency. Other references to *Meteosat* include the paper by Sir John Mason in the December 1979 issue of *The Radio and Electronic Engineer* and an article also published in the Journal in April 1978.

New and Revised British Standards

Copies of British Standards may be obtained from BSI Sales Department, 101 Pentonville Road, London N1 9ND. Non-members should send remittances with orders. Subscribing members will be invoiced and receive 40% discount.

SAFETY OF DOMESTIC ELECTRONIC EQUIPMENT

Successive editions of the British Standards Institution's specification for safety requirements of household electronic equipment have added to the scope of this standard—originally for domestic radios.

Now entitled BS 415 **Safety requirements for mains-operated electronic and related apparatus for household and similar general use** (£10.70), the new edition specifies safety requirements for a whole range of mains electronic equipment now available for use in the home. It includes monochrome and colour television receivers, radio receivers, clock radios; stereo amplifiers, tuners and turntables; record players, music centres, tape recorders, video cassette recorders, and electronic musical instruments. Auxiliary equipment provided for use with this apparatus is also covered, e.g. microphones, loudspeakers, cable connected remote control devices and battery eliminators.

Requirements are given to ensure that the apparatus is designed and constructed so as to present no danger either in normal use or under fault conditions and, in particular, provides for personal protection against electric shock, excessive temperature, ionizing radiation, implosion, mechanical instability, moving parts and against fire. The standard may also be used for professional electronic apparatus likely to be used by the layman where there is no other appropriate standard. An appendix gives supplementary requirements for splash proof electronic apparatus.

This revision brings the British Standard closer in line with the corresponding international standard, IEC Publication 65, which has been approved as a harmonization document (HD 195.S3) by the European Committee for Electrotechnical Standardization (CENELEC). However, BS 415 still contains some deviations from the harmonized standard necessary either for compliance with UK legislation, or differences in technical requirements considered indispensable for the time being.

BS 415 is used by the British Electrotechnical Approvals Board (BEAB) as the requirement for granting its certification mark.

REVISED BASIC ENVIRONMENTAL TESTING PROCEDURES

BS 2011 **Basic environmental testing procedures** Part 1.1 **General** (£5.50) has been revised. For the first time, 'controlled recovery conditions' are provided. These allow for further recovery conditions with wider tolerances on temperature and humidity to cater for specimens whose electrical parameters do not vary rapidly due to absorbed humidity or surface conditions.

The object of this standard is to provide uniform and reproducible environmental (climatic and mechanical robustness) testing procedures for those preparing specifications for components and equipment. Based upon international engineering experience and judgement, these testing procedures are designed to provide information on the following properties of specimens, i.e.

ability to operate within specified limits of temperature, pressure, humidity, mechanical stress or other environmental conditions, and certain combinations of these conditions;

ability to withstand storage and transport.

Tests given in BS 2011 (= IEC 68) permit the performance of sample components or equipment to be compared. To assess the overall quality or useful life expectancy of a given production lot, it is recommended that test procedures should be applied in accordance with a suitable sampling plan and they may, if necessary, be supplemented by appropriate additional tests.

To provide tests appropriate to the different intensities of an environmental condition, some test procedures have a number of degrees of severity. These can be obtained by varying the time, temperature, air pressure or some other determining factor separately, or in combination.

SOUND AND VIDEO TERMINOLOGY

A further British Standard glossary is issued under BS 4727 **Glossary of electrotechnical, power, telecommunications, electronics, lighting and colour terms.** BS 4727 is published in four main parts, each with a series of groups of glossaries under it. The glossary now issued is under Part 3 **Terms particular to telecommunications and electronics**; Group 10 **Recording and reproduction of sound and video terminology** (£6.40). It is identical with IEC 50 Chapter 806.

An effort is made in this standard to arrange terms in a logical order proceeding from the general to the specific and from the whole to the part, and to place allied terms in proximity. In Part 3 Group 10 the terms are classified into five main categories—general terms common to audio and video followed by general terms relating to audio alone; then two sections on sound recording and reproduction on disk (one for terms relating to disk, the other for recording and reproducing equipment), a section on magnetic recording and reproduction and, finally, two sections on magnetic recording and reproduction of sound and of video.

CODE FOR ELECTRICAL APPARATUS USED IN EXPLOSIVE ATMOSPHERES

Recent rapid progress in the development of techniques for protection of electrical apparatus for use in areas where there may be a danger from the presence of a flammable gas–air mixture has indicated the need to revise the British Standards Institution's Code of Practice CP 1003.

A new twelve-part code, BS 5345 **Code of practice for the selection, installation and maintenance of electrical apparatus for use in potentially explosive atmospheres (other than mining application or explosive processing and manufacture)** is being prepared. The latest part to be published is Part 3 **Installation and maintenance requirements for electrical apparatus with type of protection 'd' flameproof enclosure** (£3.80).

BS 5345 is intended to apply to all new installations but, where applicable, the sections in the various parts covering maintenance should be applied to all installations irrespective of the date of installation. The code is therefore of special interest to all organizations responsible for the safe operation of electrical apparatus in situations where there is a potential explosion hazard due to the presence of flammable atmospheres.

Three parts of BS 5345 have been published already. Part 1 covers the basic requirements for all the other parts of this code. Part 4 deals with the use of intrinsically safe type 'i' apparatus and Part 6 provides guidance on the use of increased safety type 'e' apparatus.

With the type 'd' protection, dealt with in Part 3, all parts of the electrical apparatus which could ignite an explosive atmosphere are placed inside an enclosure which can withstand pressure developed during an internal explosion of an explosive mixture and which prevents the transmission of the explosion to the explosive atmosphere surrounding the enclosure.

FLUXES FOR SOFT SOLDERING

Guidance on various fluxes commonly used for soft soldering metals is now available in the form of a new British Standard BS 5625 **Purchasing requirements and methods of test for fluxes for soft soldering** (£7.50). This specifies purchasing requirements and appropriate test methods for different types of solid, liquid and paste fluxes that are suitable for soft soldering or 'tinning' operations involving solders listed in BS 219. Fluxes are divided into seven categories and are classified according to the nature of their active fluxing ingredients and the metals and applications for which they are recommended. Methods of testing their fluxing efficiency and residual properties are also described: some residues may promote corrosion of the workpiece and must therefore be removed after the soldering process is completed.

ELECTROMECHANICAL COMPONENTS

A new basic specification from the British Standards Institution's Electronic Components Standards Committee is BS 5772

Electromechanical components for electronic equipment; basic testing Part 1 **General** (£3.80). This forms part of a nine-part standard defining basic testing procedures and measuring methods for the following families or sub-families of electromechanical components.

Solderless connections
Connectors for frequencies below 3 MHz
Sockets for electronic tubes
Sockets for other plug-in devices
Lever switches
Push-button switches
Rotary switches
Sensitive switches
Thermal time-delay switches
Thermostatic switches

BS 5772 Part 1 contains fundamental information on test methods and procedures and is intended to be used in those cases where a generic or detail specification for a certain component has been prepared, so as to achieve uniformity and reproducibility in the testing procedures. Requirements for the performance of the components are not covered by the Standard. The relevant specification for the item under test defines the permissible performance limits. The standard is identical with IEC 512–1.

NEW STANDARD FOR MACHINE/CONTROLS INTERFACES

BS 5782 **Interface between numerical controls and industrial machines** (£6.40) is a new standard published by the British Standards Institution and is identical with IEC 550.

The interface is divided functionally into four groups (or types); drive commands, interconnections with measurement systems and measuring transducers, power supply and protective circuit and ON/OFF and coded signals. The standard gives requirements which are common to all groups and specific recommendations and requirements for each type.

The objective is the enhancing of safety aspects, standardizing certain compulsory features to ensure the safe use of equipment from different manufacturers and to define certain electrical signal characteristics at the interface. Connection and installation recommendations are also given.

ELECTROSTATIC SENSITIVE DEVICES

A new standard just issued by BSI's Electronic Components Standards Committee relates to handling of electrostatic sensitive devices. BS 5783 **Code of practice for the handling of electrostatic sensitive devices** (£2.60) is concerned with the precautions for storage, transportation, handling and testing of all kinds of electrostatic sensitive devices, circuits and assemblies. A foreword emphasizes that static electricity of many thousands of volts can be generated in many ways, and describes some of them. It points out that this static is usually not even noticed by operatives, but that its transfer can be sufficient to destroy a device.

SAMPLING OSCILLOSCOPES

The British Standards Institution has just issued BS 5788 **Method for expression of the properties of sampling oscilloscopes** (£8.80). This standard is identical with IEC 548 and is of particular interest to laboratories and the process control industry. It is the first of three standards for oscilloscopes: others dealing with the expression of the properties of cathode-ray oscilloscopes in general and of storage oscilloscopes will appear shortly.

BS 5788 is applicable to equivalent-time sampling oscilloscopes for measuring electrical quantities, to oscilloscopes for measuring non-electrical quantities (when it is possible to express their performance in terms of an electrical quantity), and to some multi-trace sampling oscilloscopes. It standardizes the methods of expression of the properties of these oscilloscopes and, more particularly, the special terminology and catalogue data, the specification of conditions and methods for testing in order to verify compliance with the properties claimed or specified by the manufacturer.

Safety requirements are to be dealt with in a revision of BS 4743.

GUIDANCE ON ELECTROMAGNETIC FLOWMETERS

Guidance on the construction, installation and use of electromagnetic flowmeters for the measurement of conductive fluids in pipes is given in a new standard just published by the British Standards Institution. BS 5792 **Electromagnetic flowmeters** (£5.50) specifies the maximum dimensions of primary devices together with the performance requirements and methods of calibration. The standard does not specify safety requirements or include the measurement to flowrate of liquid metals or medical applications of electromagnetic flowmeters. The new standard takes into account the work undertaken by ISO and is compatible with the ISO Technical Report in course of preparation.

ELECTRICAL INSULATION TESTING

BS 5863 Methods of test for electrical resistance and resistivity of insulating materials at elevated temperatures (£2.60) describes procedures for determining the relevant properties of these materials at temperatures up to at least 800°C. It also deals with the preparation of specimens and electrodes, the test equipment (i.e. the heating chamber, specimen holder, measuring leads, temperature control and resistance measuring equipment) and advises on various precautions to be taken during testing. BS 5823 is identical with IEC Publication 345, issued in 1971.

RADIATION THICKNESS METERS

The British Standards Institution has just published BS 5868 **Guide to ionizing radiation thickness meters for materials in the form of sheets, coatings or laminates** (£7.50). This British Standard provides guidance on test methods and procedures for all radioactive isotope meters designed either for continuous operation or for check measurements. The meters are generally custom-built for industrial applications covering a very wide range of industries and specifications.

The aim is to provide a British Standard Guide which facilitates comparison of equipments of differing specifications by laying down general recommendations and test procedures. The guide applies to systems with output signals directly related to the measured quantity as well as to those related to a difference from a set value of the quantity. Safety aspects of equipment are not considered.

ORGANIZATION OF COMPANY STANDARDS

How does a standards engineer go about setting up a company standards department? What is a standards engineer anyway? The British Standards Institution has been shaping up to these questions for many years, conscious that the specifications it produces can only be taken up effectively where the principles of standardization are understood and accepted. The new edition of BSI's PD 3542 **The operation of a company standards department** (£10.50) represents the considerable experience of members of the British Standards Society in applying standards.

The Society, organized by BSI to provide a forum for individuals concerned with standards in practice, is now closely associated with the work of the International Federation for the Application of Standards, and PD 3542 takes account of this important international awareness. The publication is written in terms of company standards and the work of the standards team. It provides guidelines for the requirements, responsibilities and contribution of a company standards function, showing that standardization is essentially a management technique and can be applied to virtually all aspects of a business—for example, design, manufacture, accounting, administration and buying.

PD 3542 opens with an introduction and a set of definitions. The main subject matter is organized into eight sections:

The standards engineer (includes a job specification)
Functions of a company standards department
Organization of standards information
Scope and benefits of standardization
Standardization practice and application
Making Standards effective
British Standards in company standards operations
National and international standardization

Much of this will be familiar ground to many standards engineers (the term is used to describe all people employed in standardization regardless of local titles) but it incorporates a mass of practical advice and information which is not available elsewhere under one cover. It includes, for example, a substantial account of the standards requirements of specific departments of a typical organization with notes and suggested projects associated with the training requirements of each. The practical examples included are taken from actual practice.

Letters

From: **B. Priestley,** C.Eng., M.I.E.R.E.
G. A. Warwick, B.Sc.
A. J. Prescott, Dip.Tech.(Eng.), A.C.T.(Birm) and
Professor W. Gosling, B.Sc., A.R.C.S., C.Eng., F.I.E.R.E.
R. Benjamin, C.B., D.Sc., C.Eng., F.I.E.E., F.I.E.R.E.

The Frequency Shifting Synthesizer

I read the paper 'The frequency shifting synthesizer'* in the March issue with considerable interest. The technique expounded by the authors seems to have several other applications such as the derivation of Sidereal Time.

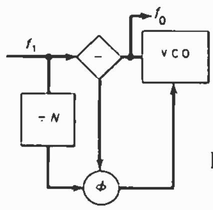

Fig. A

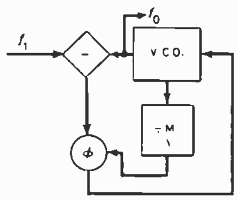

Fig. B

It occurs to me that a trivial rearrangement of the system does give considerable advantages in some cases. Referring to Fig. A, if the position of the divider is changed to Fig. B, then the output frequency is now given by:

$$f_0 = f_1 - f_0/M \quad (f_1 > f_0)$$

so

$$f_0 = f_1\left[\frac{M}{M+1}\right]$$

compared with $f_1\left[\frac{N-1}{N}\right]$ in Fig. B, i.e.

$$M \equiv N-1$$

However, if the reference drops out as in a Droitwich-locked frequency standard, the divider no longer stops and restarts indicating a false phase error. As there is, I believe, a suggestion that Droitwich may go to 198 kHz this technique could have useful applications.

B. Priestley

43 Raymond Road,
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16th March 1980

We welcome and agree with Mr. Priestley's perceptive comments. The alternative f.s.s. configuration which he shows is used in our digital oscillator correction scheme (described in our reference 3†). In the case of complete loss of the input signal, the response of either configuration depends primarily upon the nature of the mixer and phase comparator used.

The suggested change in the frequency of Droitwich transmissions has arisen since the paper was prepared, but an f.s.s. is certainly ideally suited to its adjustment. The most obvious approach, particularly when modifying existing equipment, would seem to be conversion from 198 kHz to 200 kHz. In this case $f_0 > f_1$ and the required divider moduli would be 99 for configuration (a) and 100 for configuration (b), rendering (b) slightly simpler to construct.

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31st March 1980

* Warwick, G. A., Prescott, A. J. and Gosling, W., *The Radio and Electronic Engineer*, **50**, no. 3, pp. 122–6, March 1980.

† Warwick, G. A., Gosling, W. & Prescott, A. J., 'A digital technique for temperature compensation of crystal oscillators'. Conference on Radio Receivers and Associated Systems, Southampton, July 1978, IERE Conference Proceedings No. 40, pp. 207–16.

Matrix-addressed Liquid Crystal Displays

In Reference 1 I discussed novel, relatively simple drive schemes for matrix-addressed l.c.d.s, where the net drive voltage on a polarizable crystal produces an electrical aligning force orthogonal to that generated by surface-affinity effects. More recently the important new technique of two-frequency driving has come to my notice, which permits electrical drives to generate forces either orthogonal or parallel to the containing surfaces.[2] This brief note will therefore indicate how the two-frequency method relates physically to the methods discussed in Ref. 1, what capability it can generate on its own, and how the advantages of the two approaches may be combined.

The direct-response mode may be regarded as the situation where the polarizable molecules align themselves with the instantaneous applied field in polarization and direction, whenever the field is strong enough. In the r.m.s. mode, on the other hand, the fluctuations in the applied field are fast, compared with the rotational response time of the molecule, and so the effective aligning force is defined by the mean product of the field strength itself and the polarization induced by it.

The 'two-frequency' method also uses 'low-frequency' drives, appropriate to one of these modes. However, it associates these with a super-imposed 'high-frequency' drive, whose period is (presumably) short compared with the *polarization* (as well as the rotation) response time of the molecule. This would produce a 90° phase shift in the polarization, and its interaction with the field would generate a double-frequency oscillatory agitation. However, since the torque exerted is greatest at (or near) orthogonal alignment and least at (or near) alignment parallel to the field, the mean of any resultant small cyclic movement would always be biased towards the orthogonal plane. Certainly the established fact is that such a high-frequency agitation actively drives the molecules towards an alignment *orthogonal* to the field. The directions of the ridges on the two face plates will then determine the orientation of the molecules in this orthogonal plane.

Thus the two types of drive oppose each other, and so permit us to use one to apply a large bias, to be overcome by the other. Note that this should not significantly alter the absolute voltage *difference* between the (thus increased) drive magnitudes for the ON and OFF conditions. Hence the *ratio* of the maximum drive level retaining the OFF condition (alignment with the face plates) to the minimum drive level for ensuring the ON condition (alignment with the field) can be brought much closer to unity. This then permits a dramatic increase in the number of picture elements which can be covered by a single feature-dependent drive conductor. The faster relaxation time, due to the high-frequency drive, also makes the direct-response mode easier to realize, if desired.

The two-frequency method can use a high-frequency bias and low-frequency drive or the converse. (The latter appears to offer some advantages.) In all significant respects, the method is equivalent to a low-frequency drive, applied to a material of much enhanced transition-voltage and response-time characteristics. Indeed, in the r.m.s. mode, where the two-frequency method really comes into its own, the better transition-voltage law permits the feature-independent drives to be reduced to a simple high-frequency 'burst' of low 'duty factor', which can be applied to several tens of conductors in a sequential cycle. The feature-dependent drives are then simply the sum of all the drive 'bursts' intersected by the feature conductor, sign-inverted or in phase (i.e. with high or low difference-voltages at the intersection) for 'white' and 'black' respectively. This form of multiplexing can then provide several tens of time-shared feature-independent line drives, sufficient for several character rows (of almost any length).

The scheme of my paper [1] would of course permit the use of even more drive lines, but, with such large numbers, the drive sequences become unduly numerous, long and complex. A compromise scheme would entail multiplexing not single line drives, as above, but groups of 3 or 5 parallel line drives, whose high-frequency 'bursts' would be simple mutually-orthogonal sequences, such as those of Table 5 of Ref. 1. This would still permit a major increase in the number of character rows which can be accommodated within one matrix. Whilst it would also involve a *more than proportionate* increase in the length of the feature-dependent vertical drive sequences, this more modest increase would probably be a very acceptable price for the larger character array.

To generate the control waveforms, any two-dimensional multi-character array must use a table of wanted character names against location, and a table of character names against dot-matrix black and white patterns (such as Fig. 4 of Ref. 1), to define the required vertical black-and-white sequences. If we now assume 9×7 element dot-matrix characters then, analogously to the approach in Ref. 1, the second of these tables could define each character by seven vertical-line features, each comprising nine black or white elements. If we are multiplexing sets of three parallel, mutually-orthogonal drives, these vertical line-features for each character could be partitioned into three triplets, and we could then directly store the appropriate drive sequence (in place of the black/white pattern) for each of these triplets. (See Fig. A.)

The proposed combination of parallel and serial multiplexing would offer a major increase in the number of characters per display panel, with barely any increase in the number of connections to this panel. It would admittedly require a substantial increase in the storage and logic requirements for off-panel drive generation, compared with a purely serial scheme, but, with modern l.s.i. components, this may well be a favourable 'trade-off'. Such compound multiplexing would also be well suited to the oscilloscope application.

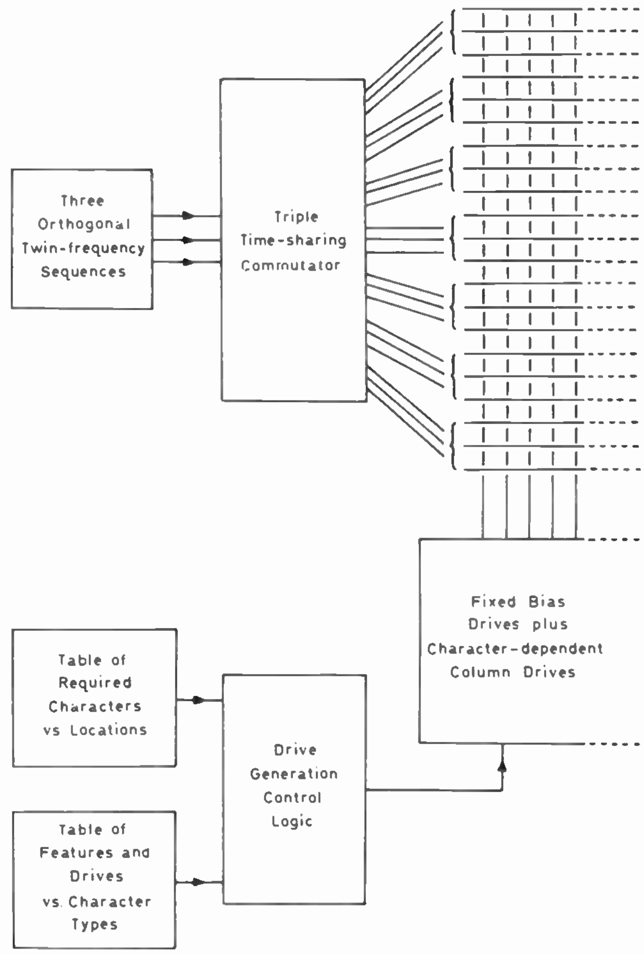

Fig. A. Compound multiplexing.

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20th March 1980

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The future of digital magnetic recording on flexible media

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SUMMARY

Digital magnetic recording on flexible media has grown from ½ in tape in 1959 to include mass storage systems, digital cassettes, data cartridges and flexible disks in 1980. The properties of greatest interest are the capacity of the removable unit, the access time to the data and the rate at which data can be communicated to another device. To the customer the price he must pay per bit of data stored is very important but he must also consider the price of the drive unit and not just the price of the removable unit. The characteristic by which different storage devices are most commonly compared is their storage capacity. To increase this continuously in response to the ever-increasing amount of information that must be stored in machine-accessible form requires a steady improvement in the amount of data that can be packed into each unit of area. The problems encountered in increasing this areal storage density are discussed in detail. Three kinds of limit to bit and track density are considered: the fundamental limit, the technological limit and the economic limit. No estimate is attempted for the economic limit which is strongly machine-dependent but the fundamental limit to bit and track density is given as 20 000 bits or tracks per mm (500 000/in). The technological limit for bit density is given as 4000–8000 b/mm (100 000–200 000 bpi) and for track density 600 t/mm (15 000 tpi).

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1 Introduction

Twenty years ago 'digital magnetic recording' meant the ½ in tape drive. Disks had only recently appeared and although they showed great promise the impact that they would develop throughout the sixties and in particular, the impact on the tape drives, was not yet apparent. Magnetic drums existed; rotating at high speeds and often equipped with multiple read–write heads they were expensive and were used principally in military applications. In just a few years the short access times of disks made them so attractive that tape drives were said to be obsolescent. The use of disks as 'virtual memory' brought about an enormous increase in demand for them. Increases in disk file capacity were obtained by increases in bit density and, more markedly, in track density, but the highest performance files eventually used fixed disks in which the head–disk interface could be controlled more precisely. The Winchester technology permitted heads to take off and land on the disk surface and used a head loading force that was only ~10 g (cf. 350 g in previous disk drives). This made it easier to achieve the lower flying heights needed for higher linear densities (bits per unit length of track). The decrease in the access times of disk drives were less pronounced since head actuators were already operating close to their limits and an increase in the rotational speed of the disks when coupled with their increased linear densities would give data rates too high for the channels to handle. Paradoxically it was the success of disks that caused a resurgence of interest in tape drives. Most of the disk drives used fixed disks and for these a back-up device was needed so that a copy of the disk data could be made frequently and stored off-line. For this purpose the high data rate of the tape drives and the low cost of tape reels were ideally suited.

Tape drives had always been well-matched to batch processing of records but the tape reels had to be brought from the tape library and mounted on the drive by all-too-human operators who frequently mounted the wrong reels. It was not uncommon in some installations to have this happen 30% of the time! An additional problem was caused by the excessive time it took to move a reel from the library shelf to the computer and back to the correct place on the shelf. The solution to both of these problems was the Mass Storage System which used cartridges stored in a honeycomb of cells. Under computer control a cartridge could be transported from its cell to the read–write station and returned after use in an average time of about 10 seconds (Table 1, column 1).

Today, the digital recording devices using flexible media include (in addition to ½ in tape and the Mass Storage System), digital cassettes and cartridges and flexible disks. Digital cassettes grew out of the familiar audio cassette but required extensive mechanical redesign to permit the tape speed to be increased from

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$1\frac{7}{8}$ ips (4·76 cm/s) to 15 ips (38 cm/s). They provide an inexpensive way of loading programs into personal computers and they are widely used in intelligent terminals. A smaller version, the mini-cassette, has a capacity greater than 64 Kbytes and so it can be used to load programs into microprocessors. It is also used in remote, low-power data-collection devices and in portable word processors.

The $\frac{1}{4}$ in tape cartridge was designed for use in small digital systems where it performed the same function as the $\frac{1}{2}$ in tape in larger computers. The cartridge contains its own tape tension and drive equipment and needs only a single motor. The cartridge operates at higher tape speed than the cassette and has much greater capacity; up to 75 Mbytes. It is widely used on communications terminals and in telephone systems and may well find its principal application in the future as a back-up device for the smaller fixed rigid disks. The cartridge has a miniature version which was designed as a device for loading programs into desktop calculators and for supplying 'program patches' to operating systems.

Finally, the most popular of the new flexible media storage devices is undoubtedly the flexible disk. Over a million drives were produced in the USA in 1979.[1] The 8 in (203 mm) diameter version is used principally in word processing systems but it is also widely used as the removable disk file for minicomputers. Each of its 154 tracks can hold a complete page of typing. The 5·25 in (133 mm) diameter flexible disk is also used in word processing systems. The average annual growth rate in world-wide revenues from flexible disk drives has been forecast[2] as 32% at least until 1981; the 5·25 in version is increasing more rapidly than the 8 in.

2 Performance Factors

The most important factors by which flexible digital recording devices may be compared are: the capacity of the removable unit in megabytes, the data rate in kilobytes per second, the average access time in milliseconds and the price of storing a bit in the removable unit. These and other related quantities are shown in Table 1 for a number of flexible media devices ranging from the IBM 3850 Mass Storage System to the minicassette. The access time to a particular record is a key functional parameter but it is determined by the characteristics of the drive rather than medium and so it will not be considered further here. The data rate is the product of the linear recording density and the head-to-medium velocity. The linear density will be discussed at some length but no additional treatment of data rate will be given. Comparison between different flexible media formats and devices is perhaps most easily made by giving the capacity in bytes of the removable unit. This quantity will be considered first.

2.1 Capacity

It has been estimated[3] that by 1983 there will be $2{\cdot}5 \times 10^{14}$ bytes stored in machine-readable form in the United States and that even this vast quantity of data will represent only 3% of all the information which might possibly be stored in such a form. Man's desire to store data is apparently insatiable and a manifestation of this over the last twenty years has been the steadily increasing capacity of magnetic tape devices and disk files.

For example, flexible disks were introduced in 1970 as the microprogram loading device for the controller on the IBM 3330 disk drive. The capacity of 81 Kbytes was increased in 1971 to 242 Kbytes when the device was introduced in Europe as a component of a data entry terminal (3740). In 1973, the capacity of the single-sided diskette using phase encoding became 400 Kbytes. By changing the MFM and M^2FM encoding methods, the bit density was increased by another factor of two and a further doubling was achieved in 1976 by using both sides of the diskette. Thus the unformatted capacity of the double-sided, double-density diskette is presently 1·6 Mbytes. There is already a diskette with a capacity of 5 Mbytes and one feels that the 10 and 20 Mbytes diskettes cannot be far behind.

2.2 Price Per Bit

A 2400 ft (730 m) reel of $\frac{1}{2}$ in computer tape sells for about $10 and has a capacity of 10^8 bytes or 10^{-6} cents per bit. This sets a standard which is difficult to match, as shown by Table 1. It will be noticed that in general the price per bit is lower the greater the data capacity of the removable unit and, unfortunately, the greater the cost of the drive. The minicassette, in contrast to the $\frac{1}{2}$ in tape reel, holds (in one version) only 64 Kbytes for which the price per bit is 8×10^{-4} cents.

When price per bit is of paramount importance, we should clearly pack the largest possible amount of data into the removable unit and this leads us to consider the factors which limit the achievement of high-areal density.

3 Limits to Recording Density

We must distinguish between three kinds of limits:

The Fundamental Limit which is imposed by the Laws of Nature and is concerned with what is possible.

The Technological Limit which is imposed by considerations of what is practical of things that are possible.

The Economic Limit which is imposed by considerations of what is profitable of those things that are practical.

For example, the fundamental limit to bit and track densities is determined by the size of the smallest particle for which the state of magnetization is stable. This yields a limiting density of about 20 000 bits or tracks per mm (500 000 per in). The technological limit to track density is determined by several factors but the most important is how accurate we can position the read–write head over

Table 1

Main parameters of magnetic recording stores

			Diskettes		Cartridge		Cassette	
	IBM 3850	½ in tape	8 in	5·25 in	¼ in	mini	standard	mini
capacity of removable unit (megabytes)	50	⩽100	1·26	0·322	15	0·27	0·81	0·15
bits per inch	~7000	6250	⩽6536	⩽5456	6400	1600	1600	800
per mm	~280	246	257	215	252	63	63	32
tracks per inch	67	18	48	48	16	13	13	13
per mm	2·6	0·7	1·9	1·9	0·6	0·5	0·5	0·5
bits per square inch $\times 10^{-6}$	0·5	0·1	0·25	0·2	0·1	0·02	0·02	0·01
average access time (seconds)	~10	72	0·45	0·4	30	15	20	15
data transfer rate (Kbytes)	1350	1250	62·5	31	24	6	8	2
price of removable unit (dollars)	20	10	3·5	2	25	16	6	3
price per bit (cents)	5×10^{-6}	$1{\cdot}25\times10^{-6}$	3×10^{-5}	5×10^{-5}	2×10^{-5}	7×10^{-4}	$1{\cdot}6\times10^{-4}$	$2{\cdot}5\times10^{-4}$
price of drive in small quantities (dollars)	>250 000	10 000	300–500	200–300	800–1000	200–400	500	125

the track. In principle this might be done with servo tracks that are located and followed by a light beam whose motion is linked to that of the magnetic head. Track densities of 590 t/mm (15 000 tpi) have been achieved in optical video recorders (but without the mass of a magnetic head). However, it may happen that the price to be charged for the recording machine precludes this solution and requires that the servo information be carried by the magnetic particles of the coating and that a two-gap head be used to obtain the position error signal.

For example, a disk surface may be divided into data sectors (85%), and servo sectors (15%), and track densities of perhaps 40 t/mm (1000 tpi) obtained in this way. But the use of any servo may be out of the question for the least expensive machines and in that case the economic limit to track density will be determined by the degree of interchangeability required between machines and by the dimensional stability of the substrate.

The fundamental limits may be simply calculated but the technological limits depend on judgement factors and therefore must be more speculative. It is possible to 'fly' a head over a smooth disk at an average height of 0·375 μm (15 μin) but is a flying height of 0·25 μm possible?; or 0·125 μm? It is still more difficult to make general, quantitative comments about the economic limits of magnetic recording since it depends very strongly on the design of a specific machine.

3.1 Fundamental Limit

An estimate of the fundamental limit to bit and track density may be made in the following way. Since the signal obtained from the recorded medium depends on n, the number of magnetic particles per unit volume, and since the particle noise depends on $\sqrt{n}$ we can maximize the signal-to-noise ratio by maximizing n. We would like to have the largest possible number of particles in each unit of volume. To achieve this, we can work with smaller and smaller particles but there is a limit to this process that is set by the size of the particles at which stable ferri- or ferromagnetism ceases to exist. The time for which a given magnetic state is stable is an extremely sharp function of the ratio of the volume of the particle to the absolute temperature. The anisotropy of the particles is also involved, so that the limit is lower for highly acicular particles than for spherical ones and is lower for materials of higher magnetocrystalline anisotropy. However, the stability time is such a steep function of volume/absolute temperature (it decreases by ten orders of magnitude for a factor of two change in this ratio) that almost all shapes and compositions of particles become unstable at room temperature when the size of the particle drops below about 200 Å. Choosing 500 Å as an average particle size at which the state is stable for archival times (30 years), we find that ultimate bit density is 2×10^4 b/mm (5×10^5 bpi).

3.2 Technological Limit to Bit Density

The technological limit to both bit and track densities is determined by the requirement that the data be recorded on the medium and read back without error. Errors can take the form of missing bits or extra bits and either of these can be permanent or temporary (i.e. they may disappear on re-reading). There are many possible causes of missing or extra bits, for example bumps or voids in the coating, pressed-in debris, airborne contaminants, etc. Even when great care is taken to eliminate defects during the manufacture of magnetic media, recording and reading errors can still occur as a result of increases in the distance between the head and the medium. Increased separation leads not only to reduced signal levels but also to broader pulses and thus can cause both amplitude and timing errors. A shift in the time of a read-back pulse can be caused by the overlapping of adjacent pulses and by coherent noise coming from previous records that have been incompletely erased or from signals on adjacent tracks.

The head–medium separation, the magnetic properties of the medium and the resolution of the head are all important in controlling the peak shift but so also is the tracking ability of the head which affects not only the highest track density which can be achieved but also the bit density.

A particulate recording medium for the highest densities will be characterized by well-dispersed, uniform and highly oriented small particles. A good dispersion is a prerequisite for a smooth surface and a homogeneous interior which minimize the modulation noise, while the use of small particles reduces the contribution of particulate noise. Uniform, well-oriented particles will have a narrow range of particle switching fields and thus reduce recording demagnetization.[5] Orientation is also needed to minimize the dead layer on the surface of a particulate media.[6] Self-demagnetization will be moderated by the proper choice of the ratio of the coercivity of the particles to their remanence and of coating thickness. Since the flexible medium is in frequent if not constant contact with the head surface, its durability is always being tested and the problem can only become more severe as higher densities require still more intimate contact. A coating which is to be used at high linear densities is usually less than 0·25 μm (100 μin) thick. There are two reasons for this: the first is to reduce the demagnetizing field, and the second, which is particularly important in disk recording, is to reduce the current needed to overwrite (and therefore erase) previous recordings. Unfortunately, thin coatings are usually less durable than thicker ones and medium wear may become a limiting factor.

It is evident that the higher the bit and track densities, the more susceptible the recording and reading processes become to defects. The elimination of errors caused by the medium will of necessity become more expensive as the bits become smaller and it may be cheaper to accept a certain number of media-related defects as inevitable and to use error correcting procedures to improve the system error rate.

The technological limit to bit density may be as high as 4000–8000 b/mm (100 000–200 000 bpi).

3.3 Technological Limit to Track Density

The problem in working at high-track densities is not how to write and read narrow tracks but how to find and follow them. The flexible substrates in digital recording media are almost invariably made of cast and tensilized films of polyethylene terephthalate (PET). The material makes substrates which are strong and light but it is dimensionally unstable and anisotropic. These factors are scarcely large enough to pose serious problems at the track widths used today (0·52 mm, 20·5 mil), but they are likely to become very troublesome within the next five years.

The 0·076 mm (3 mil) thick film used in flexible disks will have typically a thermal expansion coefficient of 24×10^{-6}/degC in one direction while in the orthogonal direction it is 12×10^{-6}/degC. Thus in a 203 mm (8 in) diameter diskette for a temperature increase of 10 degC, one axis will change by approximately 0·05 mm (2 mils) and the other 0·025 mm (1 mil). At a track density of 8 per mm (200 per inch) an ellipticity of 20% of the track width is very significant. Not only will the signal amplitude be modulated but also the amount of coherent noise from adjacent tracks will increase.

In addition to the dimensional changes with temperature, polyethylene terephthalate film undergoes significant changes with relative humidity and those changes are also anisotropic. The range in a given sample is typically 9×10^{-6}/% r.h. to 15×10^{-6}/% r.h.

The properties of a cast and tensilized film of PET are not uniform across the web; the highest values of thermal expansion coefficient are usually found at the edges of the web (typically 21×10^{-6}/degC) and the lowest value at the centre ($\sim15\times10^{-6}$/degC). However, the most isotropic region is usually found at a distance from the edge of the web of about one-third of the web width. There is some hope of the film manufacturers being able to make a modest improvement in these properties; perhaps to $12\pm2{\cdot}5\times10^{-6}$/degC and $12\pm0{\cdot}6\times10^{-6}$/% r.h. There is little prospect of a new substrate material having greatly superior properties to PET being developed in the near future. The most promising alternative substrate material for flexible disks is polyparabanic acid film which is isotropic but costs about $100/lb (cf. PET $10/lb).

The head should be designed to be as insensitive as possible to off-track noise and usually this means that the pole pieces should be as small as possible. The thin-film head makes it possible to make at least one dimension of the head extremely small and the use of run-length codes helps here. A technique which has proved to be very successful in reducing the sensitivity of a head to the information on adjacent tracks is to use two heads skewed by approximately 15° with respect to one another.[7] However, the correction achieved by this technique depends quite strongly on the density of the recorded pattern. It would also be very desirable to improve the vertical resolution of the head, so that it sensed only a small fraction of the total thickness of the coating. Unfortunately, there is no known method of achieving this.

Track densities of 600 t/mm (15 000 tpi) have been achieved in optical recording and this is probably close to the technological limit for magnetic recording on flexible media.

4 New Recording Materials

The lower limit of size for stable single-domain behaviour is not a strong function of particle composition—thus the fundamental limit to bit and track density is similar for all particles. However, the technological limits are by no means the same for all particles and this is the reason for the continuing search for new particles.

The remanent magnetization, M_r, and the thickness determine the maximum amount of flux which can be obtained from the coating and the strength of the demagnetizing field. The coercivity, H_c, describes the ability of a magnetic coating to resist demagnetization. The uniformity of particle size and the degree of dispersion and of orientation can affect the magnetic properties but they also have an important effect on the roughness of the coating and therefore on the modulation noise and the separation between the head and the medium.

Advances in recording materials have been reviewed recently.[8,9] The most frequently used particle is iron oxide (γ-Fe_2O_3). It has been in use since the 1930s but has been remarkably improved in the last ten years. It is more readily dispersible, more highly orientable and coercivities as high as 27·8 kA/m (350 Oe) are now readily available. The use of the Berthollide compounds with compositions intermediate between Fe_2O_3 and Fe_3O_4 leads to even higher coercivities. About 12 companies worldwide are active in developing iron oxide particles. Table 2 summarizes the key magnetic properties of the particles most commonly used for high-density recording and gives a qualitative comparison for other important properties.

Chromium dioxide offers excellent high-density recording performance but it is very expensive and shows time-dependent properties. These are a source of concern but there is no published evidence that they have caused problems. The particles are abrasive but this may be ameliorated by the proper choice of the polymers that occupy about 60% of the volume of the magnetic coating. Controlled abrasivity can be an advantage; chromium dioxide particles do an excellent job of cleaning heads and are often deliberately used for that purpose. The challenge is to find a binder polymer that allows the particles to keep the heads clean without wearing them out. Four companies are developing chromium dioxide particles.

Cobalt-impregnated iron oxide particles are a recent variant of the old cobalt-doped particles. The difference is that in cobalt-doped particles the cobalt and the iron are co-precipitated and so the sobalt is in the particle from the beginning. Particles made in this way are usually equiaxed rather than acicular and have very pronounced temperature- and stress-dependent properties which can be moderated to some extent by the incorporation of zinc into the particle. The possibilities of using equiaxed particles in which the direction of

Table 2

Comparison of properties of magnetic recording media

Properties	Particles	Standard Fe_2O_3	Fe_2O_3/ Fe_3O_4	CrO_2	Cobalt-substituted γFe_2O_3	Cobalt-impregnated Fe_2O_3	Metal particles
particle	length, μm	0·2–0·5	0·2–0·5	0·2–0·3	0·2–0·5	0·2–0·5	unknown
	acicularity	5 : 1	5 : 1	10 : 1	1 : 1	5 : 1	
coercivity range	Oe	250–350	300–470	450–575	300–600	300–600	1000–1150
	kA/m	19·9–27·8	23·8–37·3	35·7–45·7	23·8–47·6	23·8–47·6	7·94–91·3
saturation magnetization	e.m.u./g	74	74–82	70–80	~70–74	70–74	160
	μWb m/kg	92	92–102	87–99	87–92	87–92	198
size uniformity		poor	poor	average	poor	poor	unknown
dispersibility, orientability		average	average	good	average	average	poor
temperature sensitivity		good	good	average	poor	poor	good
stress, impact sensitivity		good	average	good	poor	poor	good
high-density output		poor	average	good	average	good	good
abrasivity		average	average	poor	good	good	good
head-cleaning		average	average	good	poor	poor	average
time dependence of magnetic properties		good	average	poor	poor	average	unknown
pass dependence of recording performance		average	poor	good	poor	average	unknown
approximate cost, $/lb		0·50	1·0	5·00	1·50	2·00	9·00 (est.) not yet available as particles

magnetization more faithfully follows the recording vector field seems to have been overlooked. Cobalt-impregnated particles are made by preparing conventional acicular particles of γ-Fe_2O_3 or Fe_3O_4 and then depositing a cobalt salt on them. The salt is then dissociated by being warmed and the cobalt diffuses into the surface of the particle modifying its magneto-crystalline anisotropy. In a way which is not yet understood, the surface anisotropy adds to the shape anisotropy of the acicular particles producing a higher coercivity than could be obtained by either anisotropy acting alone. To reach a coercivity of, say, 59·6 kA/m (750 Oe), one needs to add less cobalt to an acicular particle than to an equiaxed one and thus the undesirable temperature- and stress-dependent properties which characterized the earlier particles are reduced in the cobalt-impregnated ones. About 12 companies are developing cobalt-modified iron oxide particles.

Metal particles have seemed attractive in principle because of the high moment densities and high coercivities which could be obtained thus permitting great flexibility in the choice of properties. Unfortunately metal particles are normally extremely reactive, both with the atmosphere and with the binders commonly used in making flexible media. The particles must be stabilized and must stay stabilized in use. Metal particles are used commercially in a small number of audio cassettes. The coercivity of these particles is so high (87·3 kA/m, 1 100 Oe) that ferrite heads cannot be used with them and this presently limits their use. A head material having a higher saturation magnetization intensity such as Alfesil or some of the amorphous materials is used. Five companies are known to be developing metal particles.

Thin, continuous metal films have been worked on sporadically for at least twenty years. Their attraction lies principally in their thinness: 0·025–0·125 μm (1–5 μin) compared with 1–3 μm (40–120 μin) for particulate recording surfaces. This leads to more favourable demagnetization geometries at high bit densities and to better off-track performance at high track densities. Furthermore, the thinner the recording surface the easier it is to overwrite completely old data. Since the film consists only of ferromagnetic material it is to be expected that the moment densities of films is greater than that of particulate coatings but what is not obvious is that the fundamental noise level of thin films should be low.[10] However, these considerable advantages have thus far been offset by three very serious problems. These are: (1) the difficulty in getting reproducible properties; (2) the propensity of a thin film to corrode, and (3) its relatively poor wear resistance. Thin films are used in some recording machines but this use is limited to instant-replay video disks, a few low-density digital applications and some high-performance drums.

Within the last five years, a new approach to magnetic recording has been proposed in which the coating consists of particles aligned perpendicularly to the substrate plane or alternatively it consists of a thin film with its easy direction in the perpendicular direction. The attraction of this structure to many workers is that it apparently decouples bit demagnetization and coating thickness thus permitting high densities in thick coatings. However, another school strongly disagrees with this argument and it appears that the question will have to be decided by experiment.

5 The Cost of the Drive

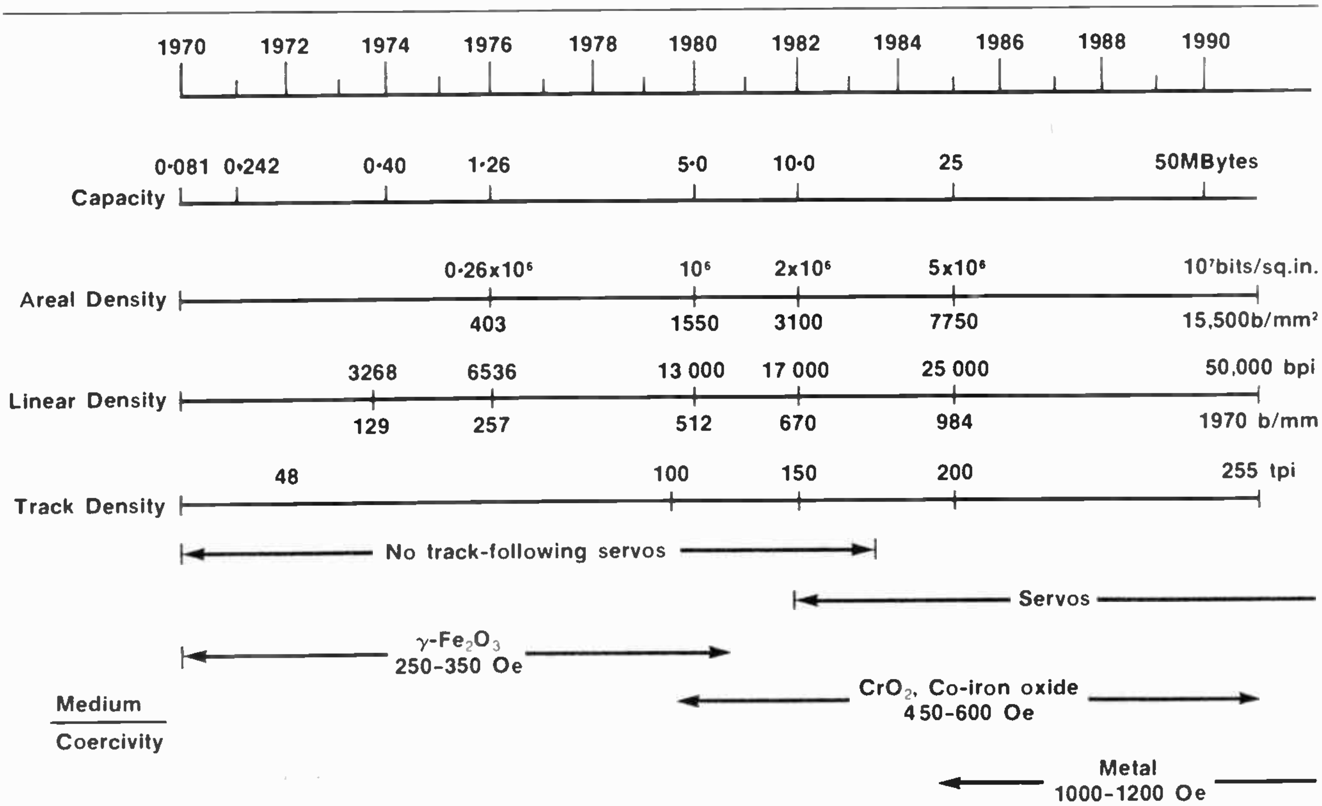
We frequently read of a new technology which promises many more bits for fewer cents than have been available previously. The latest example is the use of a laser to burn holes in a tellurium film,[11,12] where prices are in the range of 1·5 to 5×10^{-8} cents/bit. The recording process is irreversible which is a distinct disadvantage for many applications but an equally important disadvantage is the high cost which the user must pay per drive. We see from Table 1 that in general the larger the device capacity the lower is the price that has to be paid to store each bit. Users of mini- and micro-computers, using cassettes ($1{\cdot}6 \times 10^{-4}$ c/bit) or 5·25 in diskettes (5×10^{-5} c/bit), are not attracted by the possibility of improving these numbers by one or more orders of magnitude if the drive price is tens of thousands of dollars. The price per bit is a relevant quantity when comparing very large storage subsystems but the price per drive is much more important to the small-system user.

6 Future Developments

It is almost axiomatic that capacities of flexible media products will continue to increase but will do so at rates rather less than those which we have seen since 1970. Table 3 shows the changes that have occurred in some key parameters of flexible disks and suggests what may happen in the next decade. Between their introduction in 1970 and the double-sided, double-density diskette in 1976 the compound growth rate of capacity was 57% p.a. Between 1976 and 1980, when it is expected that the 5 Mbyte diskette will be widely used, the growth rate will have been 41% and will probably decrease to 26% by 1982 (10 Mbytes) and to 22% by 1990 (50 Mbytes). No increase in size is expected for the standard diskette and so the capacity increases will be obtained as a result of increasing the bit and track densities. At track densities above about 6/mm (150 tpi), track following servo methods will be necessary because of the dimensional instability of PET film. The sector servo approach will probably be used at first but will later give way to more space-effective schemes. Conventional iron oxide particles will be replaced by the cobalt-modified form (or

Table 3

The history and possible development of flexible disks



by chromium dioxide if the price is considerably reduced) for bit densities above about 500 bit/mm (12 000 bpi). Metal particles will probably be used at densities higher than about 1000 bit/mm (25 000 bpi), assuming that there are no problems of availability and that suitable head materials can be found.

Digital cassettes will continue to be used in applications where low drive price and low medium price are of paramount importance. It is probably more appropriate to compare data cartridges with ½ in tapes than with digital cassettes; they combine high reliability with sophisticated well-engineered drives. The most important use that is emerging for cartridges is to store off-line the data from fixed disks. This segment of the disk business is presently growing at nearly 120% a year[13] and is expected to continue growing through the decade.

A ¼ in tape cartridge holding 140 m (450 ft) of tape on which are recorded four tracks of data at 252 bit/mm (6400 bpi) can hold 17 Mbytes of unformatted data. It is conceivable that during the next five years the bit density, the track density and the length of tape could all be doubled, giving a capacity of 140 Mbytes in a removable cartridge having the approximate size of a thin paperbacked book. It is, of course, also possible that the format could be changed by the use of wider tapes, leading to even greater capacities in a device whose usefulness is only just being recognized.

While no increase in the diameter of the larger flexible disks is foreseen, it is distinctly possible that 3 in or even smaller diskettes will be used in portable equipment. Packs containing four or more diskettes might also offer an attractive method of increasing capacity without changing the diskette's format very drastically. This would have to be done at low cost, or one of the principal reasons for the popularity of diskettes would be lost.

7 Alternatives

The achievement of capacities in the range of 10–20 Mbytes will bring flexible disks into direct competition with the smaller rigid-disk files which have the advantage of reliability, of access time (85 ms versus 450 ms) and of data transfer rate. However, the price of a flexible disk drive is from $300 to $800 and is significantly cheaper than the estimated $1500 to $2700 for the recently announced 8 in rigid disk drive.[4] Furthermore, flexible disks are removable and need no additional device to store information off-line. It thus appears that rigid disks and flexible disks are complementary rather than competitive and that both will continue to grow rapidly in an expanding market.

Magnetic bubbles must be considered next among the possible alternatives to flexible media magnetic recording. The performance characteristics of bubble memories are impressive particularly with respect to access times but the cost is high. Today 1 Mbyte costs $3900 and while this price will decrease, it is projected[2] to be $1600 by 1985. Bubble memories really belong in a different level of the storage hierarchy and thus they also are not in competition with flexible media magnetic recording unless a reduction in price of more than a factor of ten were realized.

In the last twenty years, there have appeared at least a dozen different beam-addressable memories each of which, if the publicity were believed, could replace magnetic recording for data storage. They are listed in Table 4.

Table 4

Beam addressable memories

		WRITE	READ
1959	thermoplastic recording	electron beam	light or electron beam
1965	Gd Fe garnet	laser	laser
1966	ferro-electric-photoconductor	laser	electrical
1969	magneto-optic reading of magnetic tape	magnetic head	light
1970	Mn Bi	laser	laser
1971	Mn Ga	laser	laser
1973	video disk—Philips	—	laser
1973	video disk—Teldec	—	piezo-electric
1973	video disk—RCA	—	capacitance
1975	BEAMOS (two versions)	electron beam	e-beam/electrical
1978	optical data disk	laser	laser

At present, the only technology which appears capable of posing a threat to magnetic recording is that used in the optical data disk.[11,12] Even there the application is principally to store in read-only form, large quantities ($>2\times10^{10}$ bits) of archival data. The price per bit is low but the price per drive is high. The best method of storing large amounts of data cheaply, conveniently and reliably is by magnetic recording on flexible media and such is its potential for improvement that it is likely to remain the preferred method for many years to come.

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Indexing Terms: Frequency synthesizers, Phase lock loop, Voltage controlled oscillators

Wide range frequency synthesizers with improved dynamic performance

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SUMMARY

In a simple phase lock loop type of frequency synthesizer the closed-loop dynamic performance dictates the settling time after a frequency step, the close-in noise spectrum, the achievable cancellation of microphony in the output oscillator, and the flatness of the frequency modulation characteristic. The dynamic performance is controlled mainly by the proportional and not the integral part of the loop gain. Several techniques are described for ensuring a constant loop gain and hence good overall performance for the synthesizer over a wide frequency range of operation.

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1 Introduction

A frequency synthesizer which can operate over a wide frequency range can reduce the cost of a mobile radio. It may then be required to have a frequency range ratio of at least two-to-one or occasionally much more. A synthesizer can be made low cost by keeping it simple; the phase lock loop (p.l.l.) digital frequency synthesizer is probably the simplest type.[1] The major problem is then that it is difficult to keep the closed loop gain constant over the (wide) frequency band and this can degrade the system dynamic performance. The closed loop dynamic performance dictates: the settling time after a frequency step, the shape of the close-in noise spectrum, the ability to cancel any microphony in the voltage controlled output oscillator, and the flatness of the frequency modulation characteristic of the synthesizer.

The loop gain variation over the frequency band of operation derives from two sources. The first is the varying ratio of the programmable divider used to determine the output frequency. The second is the non-linear voltage to frequency characteristic of the output voltage controlled oscillator. This paper discusses ways of overcoming both sources of variation.

One major requirement of a frequency synthesizer is that the output spectrum should be free from spurious frequencies and have low noise content. Unfortunately the only available suitable low-noise output oscillators are those containing a tuned circuit (either LC or cavity) which is varactor controlled. Varactors in general have very non-linear control and in any case the frequency versus capacitance law of a tuned circuit is not linear.

Typically for a frequency synthesizer covering a frequency band ratio in excess of two to one the loop gain can vary by a factor of ten to fifteen mainly as the result of the non-linear varactor law.

2 Single Loop Frequency Synthesizer

Figure 1 shows a block diagram of the type of frequency synthesizer under discussion. It is basically a phase lock loop where the output signal at frequency f_o from the voltage controlled oscillator, after division by a programmable ratio N, is phase compared with a stable fixed reference at frequency f_r. Any frequency or phase error detected in the phase comparator corrects the v.c.o. frequency until a stable phase-locked condition is achieved. Then we have for the output frequency, $f_o = Nf_r$.

The p.l.l. is clearly a feedback control system which can be designed and optimized by application of the standard control methods.[8] Thus in Fig. 1 we see that the loop filter can be described as having 'proportional' and 'integral' control terms giving the overall loop path transfer function as indicated.

The function of the low-pass section of the loop filter is to reduce any spurious frequency components at f_r and harmonics of f_r which are generated by the sampling or

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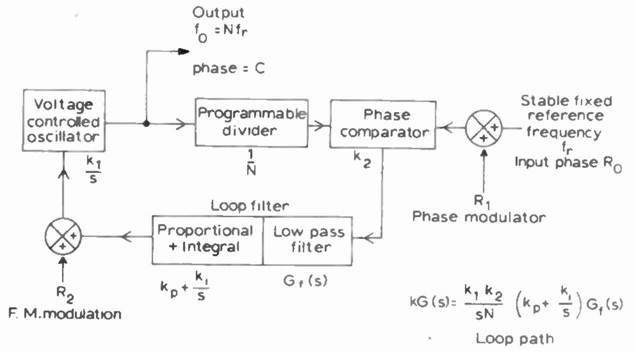

Fig. 1. Single loop frequency synthesizer.

switching action of necessity taking place in the phase comparator. This filter normally has an all-pole type of low-pass transfer function denoted by $G_f(s)$. The cut-off frequency of this filter is chosen to lie between the loop cut-off frequency f_c and the reference frequency f_r. In practice $G_f(s)$ may be distributed with, for example, part if not all of it being on the v.c.o. side of the proportional plus integral (p+i) section.

As in any feedback loop the dynamic performance of a p.l.l. can to some extent be selected by the magnitudes of the proportional and integral loop gains K_p and K_i. For example, if we make K_i zero we have a type 1 control loop signified by a single pole at $s = 0$ in the denominator of the loop transfer function. Alternatively, we can select a value of K_i such that an optimum type 2 system with a double pole at $s = 0$ is obtained. There is also an intermediate system here called 'quasi-type 1' which is strictly speaking a type 2 system but which has some of the properties of a type 1 system. In this the integrator gain K_i is much lower than its optimum value as will be seen in Section 6.

Figure 2 shows a version of the proportional plus integral control section using an operational amplifier, which is used in many p.l.l. synthesizer designs.

It is the loop cut-off frequency that has the dominant effect on the dynamic performance of the synthesizer and the proportional loop gain K_p effectively sets the loop cut-off frequency. The integral gain K_i has little effect provided it does not exceed an optimum (type 2) value.[2] The loop cut-off frequency essentially determines the synthesizer settling time, the bandwidth over which the close-in noise is improved and the frequency up to which microphony can be cancelled. The proportional loop gain also has a direct effect on flatness of the frequency modulation characteristic as described in References 3 and 4.

Figure 3 shows how the close-in natural v.c.o. noise can be reduced out to sideband frequencies equal to the loop cut-off frequency f_c, in this case about 800 Hz. Any v.c.o. microphony is also cancelled to the same frequency by the same feedback action that reduces the close-in noise.

In many respects there is little to choose between the type 2 and type 1 p.l.l. dynamic performance. All else being equal, the type 1 system can have a marginally faster settling time for frequency steps small enough so that no saturation effects occur in the system.[2] The maximum difference is a factor of 1·5. However, any white noise from the phase comparator is more rapidly attenuated going above the loop cut-off frequency in the type 2 system.

The main difference between the type 2 and type 1 system is in the phase error introduced after a frequency step. A frequency step, resulting from a sudden change in the division ratio N, corresponds to the appearance or disappearance of a phase ramp input at the reference. A type 2 system ensures that this results in no phase offset error. However, in most communication systems the presence or absence of a final phase error is quite academic after the output frequency has settled to a steady value.

There may be other reasons for choosing a type 2 system. One of these is given in Section 6 of this paper. Another reason arises when a low noise phase comparator as described in Reference 1 is used. Such a phase comparator has quite a small phase range before saturation occurs. It also requires that the proportional gain of the loop filter be quite low. It is then impossible, given the possible range of output voltage from the phase comparator, for the v.c.o. frequency to be controlled other than over a very small frequency range. In this case a type 2 or quasi-type 1 system is mandatory; the pure type 1 system is unusable.

The following Sections outline some useful ways in which the p.l.l. synthesizer loop gain or at least the proportional part of the loop gain can be kept constant over a band of frequencies.

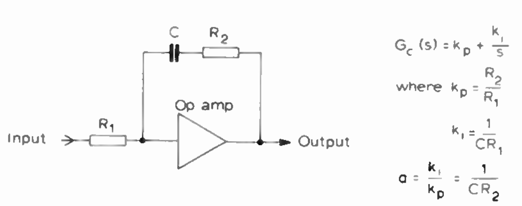

Fig. 2. Proportional plus integral filter section.

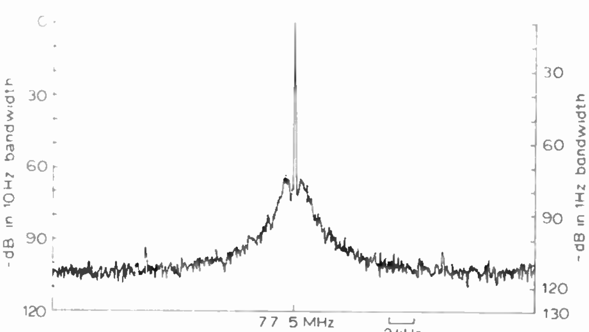

Fig. 3. Typical output spectrum of synthesizer. (Analyser limit approx. −115 dB Hz.)

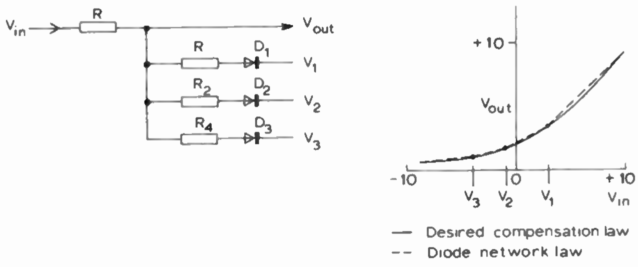

Fig. 4. Varactor law compensation method.

3 Diode Linearizing Networks

Figure 4 shows a network with a piecewise adjustable input–output characteristic.[5] This can be arranged either to compensate for the v.c.o. characteristic or to compensate for the total loop gain variation including the effect of the varying divider ratio N. The latter case is more usually required for the simple p.l.l. frequency synthesizer, however the former case of just linearizing the v.c.o. is required for the fast switching frequency synthesizer described in Section 5.

For high positive input voltages in Fig. 4, all the diodes are back-biased and so the network provides no attenuation and has unity gain. As the input and hence the output voltage descends to V_1, diode D_1 conducts so connecting a resistor of value R effectively in shunt across the output. The network then has an attenuation, i.e. a gain of a half. As the input voltage further descends when the output voltage reaches V_2 the second diode conducts introducing an extra attenuation so that the gain becomes a quarter. And so on as the output voltage descends to V_3 and below.

The voltages V_1, V_2 and V_3 are chosen for the best fit of the piecewise characteristic to the desired characteristic. They can be derived from single transistor open-emitter follower stages which have a low source impedance. Alternatively the diode can be connected directly to a pair of resistors across the supply rails, which are chosen to give the desired voltage V_i together with the desired source impedance R_i. Further simplification is often possible by using a single multiple resistor potential divider to provide the correct voltages and then using series resistances to bring the respective source impedances up to the desired values.

The use of binarily-related resistor values is quite convenient in that this ensures that the gain discontinuity of the network never exceeds a factor of 2. Other ratios can easily be used if required.

A major disadvantage of the passive network as shown in Fig. 4 is that because part of the characteristic is obtained by attenuation, of necessity the input voltage range has to be larger than the output voltage range. However, with the p+i filter of Fig. 2 a compensating amplification can be obtained by reducing the input resistor value R_1. However, a diode compensation network with an inverse characteristic can alternatively be buffered and placed in the feedback path in order to avoid amplifier saturation.

Some care has to be exercised in some cases to keep the impedance levels of the network low. Otherwise it can happen that the Johnson noise from the resistors is sufficient to excessively degrade the v.c.o. spectrum by frequency modulation. Typically with a 1 MHz/V v.c.o. the resistor values should be kept below about 10 kΩ unless other precautions are taken.

4 Distorted Ramp Type 1 System

Figure 5 shows a sample-hold type phase comparator for use in frequency synthesizers. The first sampler samples the ramp triggered by the R input at the time of the leading edge of the V input. The voltage held on the capacitor at the end of the sampling pulse is therefore proportional to the time, hence phase, difference between the V and R inputs. The (optional) second sample-hold removes the sampling glitch present on the output of the first sampler, which results from the finite first sampler pulse width. This method removes from the output practically all the unwanted sampling frequency components.[7]

The input–output characteristic of a sample-hold phase comparator can be made non-linear by distorting the shape of the ramp. Figure 6 shows how an exponential-shaped ramp can be produced to be used either as it is, in (b), or with a linear section added as in (c). In principle further distortion can be carried out by diode networks as in Section 2 but this can become more difficult to implement for circuitry operating dynamically at the sampling frequency.

In order to use such a phase comparator with a distorted characteristic to linearize the loop gain, the synthesizer loop must be of type 1. Because then there is no integral action in the loop filter ($K_i = 0$), it is the output voltage from the phase comparator that has to set the v.c.o. varactor voltage appropriate to the selected

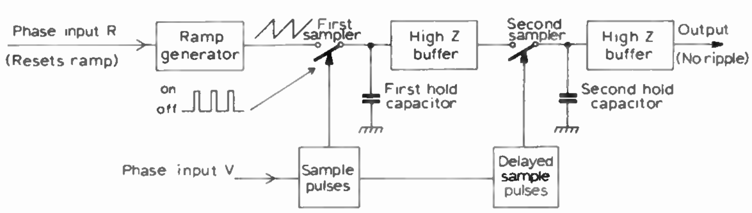

Fig. 5. Double sampling phase comparator.

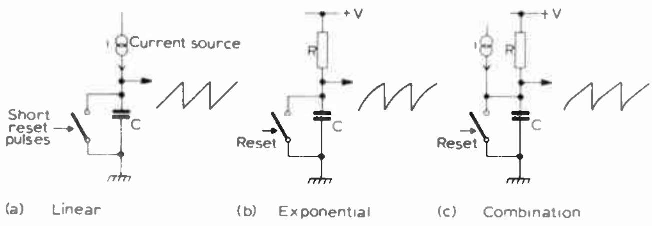

Fig. 6. Ramp generator types.

frequency. By suitable choice of loop filter and of gain distribution around the loop, it is possible to arrange that at a given frequency, and hence phase comparator voltage, the phase comparator slope compensates for the loop gain variation caused by the divider and the v.c.o. varactor. At the same time it is usually possible to choose the loop filter so that an optimum dynamic response is obtained, critically damped for example.

This method can be difficult to design in practice and suffers from the further disadvantage that there is no suitable place for inserting frequency modulation in the loop if the two-point modulation method of Reference 4 is to be used.

5 Fast Switching Frequency Synthesizer

Provided that a linear v.c.o. is available or the v.c.o. is linearized by the method shown in Section 3, it is easy to use a hybrid multiplier to compensate for the loop gain variation caused by the divider. Figure 7 shows a way of doing this that has the added advantage of predictively setting the v.c.o. nearly to the correct frequency when a frequency change is commanded.[6]

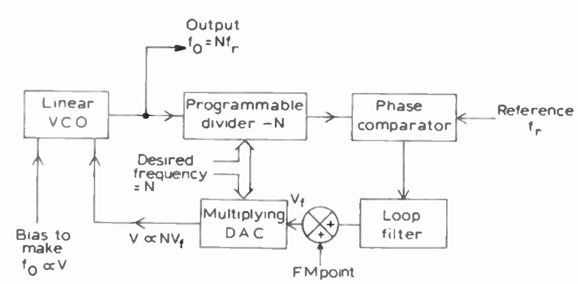

Fig. 7. Fast switching p.l.l. frequency synthesizer.

The multiplying digital-to-analogue convertor (d.a.c.) is a hybrid multiplier where the analogue input (reference) signal is multiplied by a digitally set fractional number. If this number is the same (integer) ratio N as is fed to the programmable divider the hybrid multiplier will exactly compensate for the divider loop gain variation of $1/N$. Furthermore, provided that any v.c.o. offset is cancelled by a suitable bias it is found that the steady state input voltage to the hybrid multiplier is the same for all frequencies. Thus any required change in the v.c.o. voltage V is provided instantly by a change in the value of N fed to the multiplying d.a.c. The feedback loop then only has to correct for any slight inaccuracies in the v.c.o. linearity. Because the loop gain is constant, the settling time for these small errors can be kept as short as is theoretically possible over the whole synthesizer frequency range.

With this method care must be taken to ensure that any white or wideband noise from the multiplying d.a.c. is low enough not to overmuch degrade the v.c.o. spectrum. Filtering of this noise can only be achieved at a cost in frequency setting time which would negate some of the advantage of the method.

6 Split-loop Frequency Synthesizer

As mentioned in Section 2, it is the proportional part of the loop gain and the cut-off frequency which dominates the p.l.l. dynamic performance; the integral gain can often be relatively unimportant. The split-loop frequency synthesizer shown in Fig. 8 makes use of this fact.[2]

The proportional plus integral loop filter section in Fig. 8 is split into two parts as shown. Each of these respectively feed voltages V_f and V_c to the two control inputs of the LC type v.c.o. Because of the presence of the integrator, in the steady state the output of the phase comparator becomes equal to the mid-rail voltage $\frac{1}{2}V$. The control voltage V_f is then also clearly equal to $\frac{1}{2}V$ and this is true whatever the v.c.o. frequency. The v.c.o. frequency is then set solely by the integrator output voltage V_c supplied to the varactors C_1 and C_2. At the voltage $\frac{1}{2}V$ the capacitance voltage sensitivity of the varactor C_3 and hence of C_m is constant (C_m is the series combination of C_3 and C_4). It is possible to choose C_1, C_2 and C_m so that the frequency variation caused by the variation in C_m is directly proportional to frequency to a good approximation over the v.c.o. range. For small perturbations this ensures that the total proportional loop gain is kept practically constant.

Table 1 shows the best relative values of C_m in the case of the varactors C_1 and C_2 having roughly the same values over their ranges. In a practical case, C_4 can be chosen to set the correct value of C_m with $V_f = \frac{1}{2}V$. Table 1 also shows that remarkably small variations in the proportional gain K_p occur over quite wide v.c.o. ranges. For example, over a two-to-one frequency range K_p varies by about $\pm 0{\cdot}7$ dB. Varactors C_1 and C_2 need

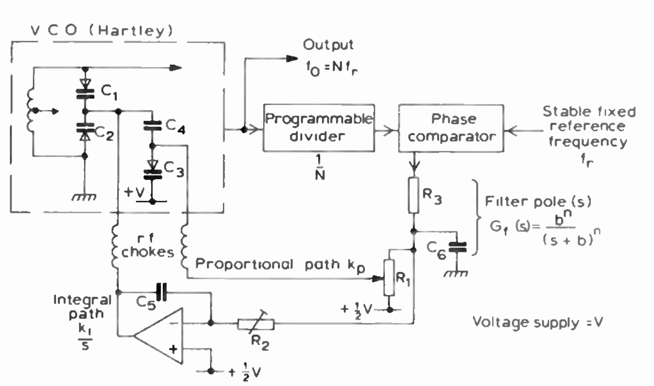

Fig. 8. Split-loop frequency synthesizer.

Table 1

Max. v.c.o. frequency / Min. v.c.o. frequency	Max. C_1 / Min. C_1	C_m / Min. C_1	K_p variation (dB) (peak-to-peak)
1·1	1·26	1·59	0·028
1·2	1·55	1·76	0·102
1·3	1·89	1·94	0·210
1·4	2·25	2·12	0·344
1·5	2·65	2·30	0·497
1·6	3·09	2·49	0·663
1·7	3·57	2·67	0·841
1·8	4·09	2·86	1·025
1·9	4·64	3·04	1·215
2·0	5·24	3·24	1·408
2·2	6·54	3·62	1·799
2·4	8·00	4·00	2·190
2·6	9·62	4·39	2·578
2·8	11·39	4·78	2·957
3·0	13·32	5·16	3·328
3·4	17·67	5·94	4·040
4·0	25·37	7·12	5·033

then only have a maximum to minimum capacitance ratio which exceeds 5·24; this is not greatly in excess of the ratio of 4 which would have been required to give a two-to-one frequency variation had C_m been absent.

The Bode diagram of the loop gain of the split-loop system is shown in Fig. 9. This shows that at the 0 dB point the asymptotic cut-off frequency is determined solely by the proportional loop gain. The integrator ceases to have an effect on the gain asymptotes above the frequency $f_a = a/2\pi$ where a is the zero in the loop transfer function given by $a = K_i/K_p$. The real 3 dB or 90° cut-off frequencies are slightly dependent on the integral gain K_i but only by a few per cent.

The loop cut-off frequency therefore remains practically constant provided the integral gain does not become too high. The usual strategy is to set the correct proportional and integral gains for an optimum type 2 response at the low-frequency end of the v.c.o. range. Then at the high-frequency end a quasi-type 1 response will be obtained with a proportional gain that is low by a factor not exceeding about 3/4. Reference 2 confirms that a good transient response, and hence settling time for frequency errors after a frequency jump, is thus obtained over the whole v.c.o. band.

A major advantage of this method is that the two-point frequency modulation can be used.[4] The f.m. part

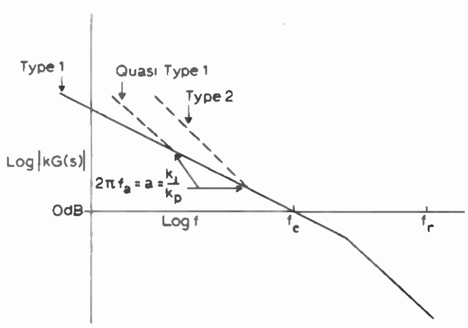

Fig. 9. Bode diagram of system.

of the modulation is added to the proportional control voltage V_f. The flatness is the same as the K_p variation. Table 1 shows that the step in the f.m. characteristic that occurs at the loop cut-off frequency will be less than 0·7 dB for the 2 to 1 frequency range case.

7 Extended Range Frequency Synthesizers

Given a p.l.l. frequency synthesizer with a 2 to 1 frequency range, its range can be extended downwards almost indefinitely by placing a divider (followed by a low-pass filter if necessary) in the synthesizer output as shown in Fig. 10. Provided that the system is operated with a binary number representation of frequency the decode and command logic can be quite simple. For the more usual BCD number system this logic becomes more complicated and a major part of the system.

The main disadvantage of this system is that the step size is different on the divided down ranges. Figure 11 shows a way round this problem. Here the range divider is placed in the phase-lock loop. It is then as if the range divider were part of the v.c.o. and the output frequency is determined only by the programmable divider ratio N. However, because the overall division ratio does not change by more than 2 to 1 the gain variation due to divider ratio is small.

Some decoding logic is needed to decide when the range divider should be switched but this is relatively easy to implement. The usual strategy is to base the decision on the number of leading zeroes in the binary or

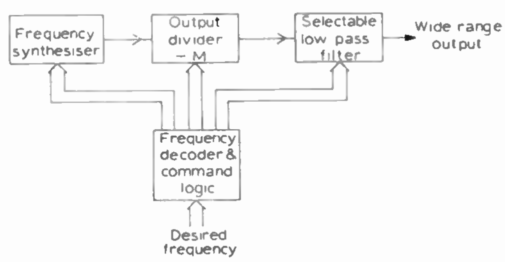

Fig. 10. Range extension by output division.

BCD desired frequency number. It is also sometimes helpful to introduce backlash to minimize range switching, at the expense of requiring a slightly larger v.c.o. range.

In a practical synthesizer required to cover several decades, a combination of the two-range divider methods is likely to produce the best overall system. The second technique can be used to provide a synthesizer with a decade range or slightly more and this then can be extended downwards in decade steps by the first method.

Oscillator phase noise is improved by a divider. The improvement is the division ratio. Thus the noise sidebands of an oscillator divided by a ratio of ten are $20 \log_{10} 10 = 20$ dB lower (in power density). This can be seen on the basis of the oscillator phase fluctuations which cause timing fluctuations in the first stage of the divider. These same timing fluctuations appear at the

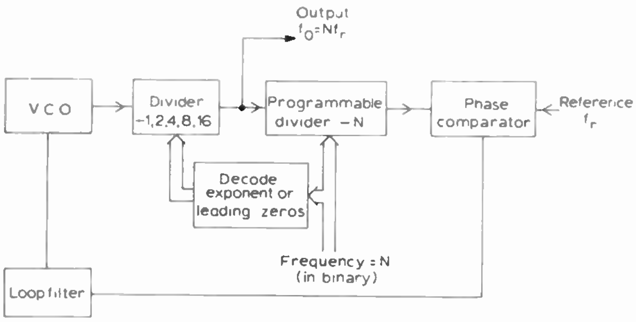

Fig. 11. Extended range frequency synthesizer.

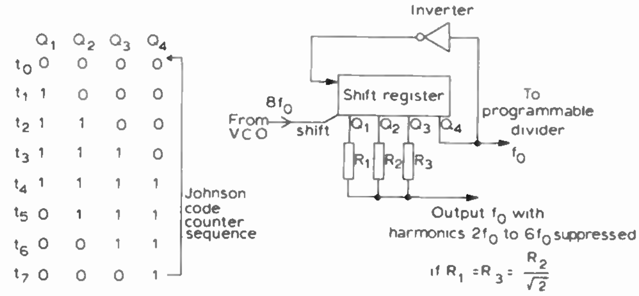

Fig. 12. Divider with harmonic suppression.

divider output but then referred to the much lower frequency with a period M times longer corresponding to a phase noise reduced by the ratio M. The phase noise is ultimately limited by the signal-to-noise ratio effectively present at the divider input. The phase component of any input noise is added to the phase noise already present on the input (from the v.c.o.).

The square wave output from a divider can be a disadvantage in that it is rich in odd harmonics. Figure 12 shows a method of cancelling some of the harmonics when the divider is implemented by a Johnson code counter. The resistors and the shift register effectively form a transversal filter with nulls at the harmonic frequencies. A further advantage is that phase-shifted versions of the divider output can be obtained using a further set of resistors connected to different Q and $\bar{Q}$ outputs.

This method is usable for both range extension methods.

8 Conclusions

The techniques that have been described here make it possible for a single-loop frequency synthesizer to have a good dynamic performance over a wide frequency range. Improved closed loop dynamic performance results in better v.c.o. close-in noise, less v.c.o. microphony effects, faster frequency setting, and a better frequency modulation characteristic. These techniques enable the simple type of digital frequency synthesizer to cover a much wider range of applications with a good overall performance.

9 Acknowledgments

The author would like to acknowledge the efforts of several colleagues in confirming the efficacy of the techniques outlined in this paper, namely R. I. H Scott and P. A. Jordan of PRL, T. G. Giles of MAL Mitcham, and T. Ault, N. Walters and J. Butler of MEL Crawley.

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Indexing Terms: Satellites, weather, Data transmission, Demultiplexer, Algorithm

A signal demultiplexer for *Meteosat* primary data user stations

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SUMMARY

Digital signals received from the *Meteosat* satellite consist of a serial bit stream which contains the output of the satellite's spin scan radiometer after computer processing at Darmstadt, Germany. This paper describes a system which recognizes the particular format being transmitted, picks out a chosen channel from the data, and turns it into a form suitable for display on a facsimile recorder or similar instrument.

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1 Introduction

The European *Meteosat* satellite was first launched in November 1977. It is one of five geostationary satellites, spaced at about 72° intervals round the globe, which give continuous coverage of meteorological data from equatorial regions up to about 60° of latitude on a 24 hours, 365 days a year basis from an altitude of 35,600 km above the Earth surface.[1,2]

The prime sensor of this satellite is the spin scan radiometer camera[3] which scans the complete Earth disk from south to north in three wavelengths once per half hour. The resultant signals are first digitized and then transmitted by radio link to a ground station at Darmstadt, Germany. These signals are then computer processed, and subsequently retransmitted to the satellite, which acts as a relay, and broadcasts the data in a variety of formats to specially designed receiving stations on the side of the Earth facing the satellite.

There are two main types of reception facility:

(a) Secondary Data User stations (or SDUS) which have the ability to receive data in analogue format only.

(b) Primary Data User stations (or PDUS) which are designed to receive data in digital form.[6]

The digital transmissions permit complete Earth disks to be received in one transmission, whereas the analogue transmissions only allow small segments to be received at one time. In addition to the Earth disks, the digital transmissions permit a sectorized view of Europe to be obtained. Both transmission formats allow the reception of data from the American *GOES*† satellite situated at 75°W. This paper is concerned with the decommutation of the digital output of a PDUS.

2 Description of Signals

2.1 Transmission Method

The digital signal is transmitted at a frequency of either 1691·0 or 1694·5 MHz, since *Meteosat* has two transmission channels available. The data is contained as phase modulation of this carrier at a bit rate of 166·667 kilobits/second. The digital code used is bi-phase-L which permits reasonable bandwidth conservation, and enables the clock frequency to be extracted at the receiver (Fig. 1, refs. 4, 5).

The radiometer output data are transmitted as 8-bit serial words, permitting a resolution of one part in 256 which is adequate for the type of signal being transmitted. In fact the visible channel only requires 6 bits as 64 levels of grey scale is sufficient. In this case the two lesser bits are not used to transmit any information, but are retained so that at all times the word length is 8 bits. The word rate is therefore 166·667/8 = 20·833 k words/s at all times, although some words in the

† Another synchronous meteorological satellite (*GOES* = Geostationary Operational Environmental Satellite).

0033–7722/80/060297+10 $1.50/0

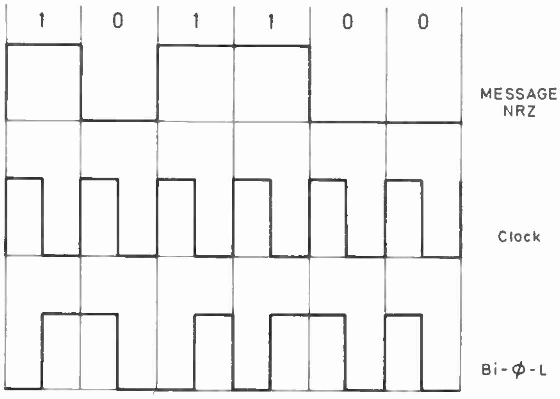

Fig. 1. Digital coding for *Meteosat* signals.

message are not radiometer data, but other data which are required to enable the transmission to be correctly decommutated.

It is clearly necessary to be able to locate the beginning and end of each word, and also to keep a tally of the word count. To do this, the words are sent in groups, and every 364 words a synchronizing word, of 24 bits duration, is inserted. These sets of 364 words are called frames.

In *Meteosat* data, the frame length is always the same. However, the next larger grouping of data is described perhaps misleadingly as sub-frames, because sub-frames consist of either 8 frames or 4 frames duration.

The frame number within a sub-frame is numbered by means of an identification (ID) word and every first frame of a sub-frame contains a label, which indicates the type of data contained therein.

A typical 4-frame sub-frame is shown in Fig. 2(a).

2.2 Information Types

There are three types of radiometer data transmitted in *Meteosat*—these are:

(1) visible channel sensitive to wavelengths in the range 0·4 to 1·1 μm
(2) water vapour channel (5·7–7·1 μm)
(3) infra-red channel (10·5–12·5 μm)

These types of data may be transmitted separately or together in a multitude of possible formats.

2.3 Data Formats

There are three basic types of digital data format with *Meteosat*. These are:

(a) A format, giving complete Earth disks
(b) B format, giving the European area
(c) X format, giving the Americas (see Fig. 3)

The A and B formats can be transmitted in any of the three wavelengths: visible, water vapour or infra-red. Since the X format originates from the American *GOES* satellite, which does not have the water vapour channel, only visible and infra-red data are available in X format.

A formats consist of 8-frame sub-frames and both other formats consist of 4-frame sub-frames.

2.4 System Resolution

The visible channel is capable of a 2·5 km resolution at the satellite sub-point. The infra-red and water vapour channels have a resolution of 5 km at the satellite sub-

4 - Frame Sub-frame

SYNC	ID	LABEL	SP1	IDENT.	SP2	INTERPRETATION
SYNC	ID	INTERPRETATION				
SYNC	ID	INTERPRETATION				
SYNC	ID	INTERPRETATION			SP3	

←364 bytes→

8 - Frame Sub-frame

SYNC	ID	LABEL	SP1	IDENT.	SP2	INTERPRETATION
SYNC	ID	INTERPRETATION				
SYNC	ID	INTERPRETATION				
SYNC	ID	INTERPRETATION			SP3	
SYNC	ID	SP4				
SYNC	ID	SP4				
SYNC	ID	SP4				
SYNC	ID	SP4				

←364 bytes→

Fig. 2. (a) Header sub-frames.

4 - Frame Sub-frame

SYNC	ID	LABEL	SP6	DATA
SYNC	ID	DATA		
SYNC	ID	DATA		
SYNC	ID	DATA		GRID

←364 bytes→

8 - Frame Sub-frame

SYNC	ID	LABEL	SP5	DATA
SYNC	ID	DATA		
SYNC	ID	DATA		
SYNC	ID	DATA		
SYNC	ID	DATA		
SYNC	ID	DATA		
SYNC	ID	DATA		
SYNC	ID	DATA		GRID

←364 bytes→

Fig. 2. (b) Data sub-frames.

Spectrum: VIS, IR, WV; frequency: every 3 h; duration: 6–30 min.

Spectrum: VIS, IR, WV; frequency: every ½ h; duration: 1–4 min.

Spectrum: VIS, IR; frequency: every 3 h; duration: 7 min.

Fig. 3. *Meteosat* dissemination formats (reproduced from *Meteosat* Ground Segment Handbook).

point. The resolution varies with latitude as suggested by Table 1.

2.5 System Dynamic Range

The word length used in the digital transmissions is 8 bits giving a total of 256 possible levels transmitted. However, the visible channel only makes use of 6 bits, giving 64 possible levels from black to white. The infra-red transmissions are arranged so that a large signal corresponds to a low temperature: a small signal indicates a high temperature. The temperature range covered is considerable; from the highest cloud tops to the hottest desert surfaces on the Earth the observed temperature may vary from about 5 K to 320 K.

3 Transmission Formats

3.1 Format Details

One of the abilities of the *Meteosat* system is to transmit data in one, two or three wavelengths simultaneously. Thus a B_{IVW} transmission indicates a transmission of the European area in infra-red, visible channel, and water vapour channel. The actual transmission formats used are described in Table 2.

Table 1

Ground resolution as a function of latitude

Latitude°	Resolution IR/WV (km)	Resolution Visible (km)
0	5	2·5
10	5·11	2·56
20	5·51	2·75
30	6·26	3·13
40	7·56	3·78
45	8·55	4·28
50	9·90	4·95
55	11·81	5·91
60	14·68	7·34
65	19·38	9·69
70	28·30	14·15
80	253·0	126·0

Note: (1) It is assumed that ground longitude is same as spacecraft.
(2) If longitude not same, read latitude column as geocentric angle between spacecraft and ground point.

When the visible channel is transmitted on its own or along with one other wavelength (e.g. B_{IV}) the full resolution data format is transmitted. Because the visible channel has double the resolution of the infra-red or water vapour, twice as many lines have to be transmitted. As there are twice as many data points along a line, altogether there will be four times as many words in a transmission as the other data. Therefore, for example, in A_{IV}, there is one infra-red sub-frame followed by four visible sub-frames (two of Vis 1 half-lines and two of Vis 2 half-lines).

When all three wavelengths are being transmitted, the visible data are transmitted at full resolution in the direction of scan, but the number of lines of visible data transmitted is the same as the infra-red or water vapour channels, giving reduced vertical resolution at this wavelength.

A summary of the transmission formats used is given in the *Meteosat* Handbook.[6]

It can be seen that the number of possible formats transmitted is very large, and consequently the task of devising a system to output the data to a facsimile machine requires a device with some decision making capability and not just the ability to recognize the format.

Considering the A_{IV} format, we have one 8-frame sub-frame for each infra-red line, and two 8-frame sub-frames for each visible line. But there are two visible lines for each infra-red line.

Now bit rate is 1 bit every 6 μs or 166·6667 k bits/s, i.e. 166·6667/8 = 20·8333 k words/s.

Each frame contains 364 words, therefore frame rate is 20,833·33/364 = 57·2344 frames per second (this figure is the same for all transmission formats).

Table 2

Complete Earth disks:	A_V	A_{IV}	A_{IVW}	A_I	A_{IW}
European area:	B_{IV}	B_{IVW}	B_{IW}	B_I	
USA:	X_{IV}				

Sub-frame rate is 57·2344/8 = 7·1543 sub-frames/s
= 429·258 sub-frames/min
(since there are 8 frames per sub-frame. But there is only one IR sub-frame in each 5 sub-frames, therefore:
IR sub-frame rate = 429·258/5 = 85·8515 lines/min.
Visible channel rate = 85·8515 × 2 = 171·7030 lines/min

If the same calculations are performed for all other formats Table 3 results.

Table 3

Format	Vis line rate	IR line rate	WV line rate
A_V	214·6296	—	—
A_I	—	429·2582	—
A_{IW}	—	214·6296	214·6296
A_{IVW}	107·3148	107·3148	107·3148
A_{IV}	171·7033	85·8516	—
B_I	—	858·5182	—
B_{IW}	—	429·2582	429·2582
B_{IVW}	214·6296	214·6296	214·6296
X_{IV}	343·407	171·703	—

It is apparent from Table 3 that with the exception of the B_I format, the data rate is such that the images may be printed on a facsimile machine which has a choice of two speeds, namely 429·258 and 343·407 lines per minute. This is because the line rates may be divided into these speeds to give an integer result, except for the B_I format. The only way to print the B_I format directly would be to use a facsimile machine running at 858·518 lines per minute.

As the technology was available in the laboratory to modify an existing machine to run at the two lower speeds, it was decided to ignore the B_I format. If the requirement arose to print B_I data directly, this could be met by recording it on tape and playing back at half speed.

3.2 Description of Header

At the beginning of each transmission, several sub-frames are transmitted which contain information about the data which is to follow. These are either 8-frame sub-frames (A format) or 4-frame sub-frames (B or X format)—refer to Fig. 2(a). The ID word is incremented by 1 for each subsequent frame within a sub-frame, permitting the system to keep track of its position. Following the sync and ID words, there is a label. The label is common to header and data sub-frames and contains the following information:

Number of frames in sub-frame (4 or 8).
Total number of sub-frames in message.
Current sub-frame number (1 for header).
Image line number (0 for header and conclusion).
Image number from start of mission.
Format indicator (A, B, X).
Vis 1, Vis 2, IR, WV, Grid, Text present indicators.
These indicators state data present if the corresponding word is all at 1's (FF hex). The meaning is slightly different for the header sub-frame to the data sub-frame. If the indicator states data present in the header, it indicates the data content of the overall transmission. If the indicator states data present in the data sub-frame, it refers to content of the current data sub-frame.

The label takes up a total of 24 words, although only 19 are currently used. Referring to Fig. 2(a), SP1 consists of 8 bytes all set to zero, identification data which give the data and various other information, followed by SP2 (16 spare bytes set to zero), and then interpretation data. This gives information about corrections added to the spin scan radiometer data and also calibration data.

The current design of system is not designed to read these latter data out, although this could have been done, had the program length of the existing system permitted. However, a microprocessor system to read out the data from the serial/parallel output has been developed, and provides these data when required.

A detailed description of the header is given in the *Meteosat* Ground Segment Handbook.[6]

3.3 Description of Data Sub-frames

Referring to Fig. 2(b), the A format data sub-frame starts with sync and ID words followed by the label of 24 bytes. SP5 contains 40 bytes all set to zero, and then data start at word 68 (starting from the sync word first byte as word 0). The data continue until the end of the frame, when there is a sync word and ID word, followed by a complete frame of data.

When 2500 words of data have been transmitted (word 47 of 8th frame), the grid data are transmitted. Each bit of these data corresponds to one word of the message. The grid takes up 2500 bits + 28 spare bits, i.e. 316 bytes. The end of the grid corresponds to the end of a sub-frame.

For B and X sub-frames, there is half as much data to be transmitted, i.e. 1250 bytes, which necessitates a smaller sub-frame of 4 frames duration. Referring to Fig. 2(b), SP6 contains 8 bytes, the data start at word 36, continuing until word 204 of the 4th frame, at which point the grid data are encountered. These consist of 1250 bits + 14 spare bits, i.e. 158 bytes. If the grid bit is zero, then there is no grid at the corresponding point. If the grid bit is 1, then there is a grid at the corresponding point.

B formats also have, at the end of the message, some annotation, which may be printed out (if required). A detailed description of the data sub-frames appears in the *Meteosat* Ground Segment Handbook.[6]

4 Serial/Parallel Converter Design

4.1 Design Strategy

The output of the receiver consists of two signals; an NRZ serial bit stream, and the clock, the NRZ having been generated by EXCLUSIVE OR-ing the bi-phase-L signal with the clock. This unfortunately creates a phase ambiguity, because the recovered clock may be of the incorrect phase, resulting in the recovered NRZ being inverted.

In order to decommutate the synchronous serial digital signal, the positions of start and finish of data words must first be found. This is accomplished by feeding the serial bit stream into a shift register and clocking it through by means of the clock. Synchronism of the system is determined by a synchronization word detector, attached to the various shift register stages.

Detection of the synchronization word performs two tasks:

(1) establishing the exact time when the 8-bit data word is in its correct position in the shift register;
(2) presetting the word count within the frame to 2.

In the case of *Meteosat*, the synchronization word is 24 bits long, and to detect this, a 24-bit shift register and associated AND gate and inverters is required. The actual word is 0506DE (hex).

As explained earlier, there is some ambiguity in the phase of the clock which is produced by the digital receiver. To resolve this, a separate 24-bit detector is used which is set to decode the inverse of the synchronization word (FAF941). The system is so arranged that if two or more inverse synchronization words occur within about 20 ms, then the phase of the signal is reversed, at which point the normal synchronization detector will operate on the receipt of subsequent pulses.

A divide-by-eight circuit keeps track of when the word is in its correct position in the shift register. When the bit counter indicates zero a data latch is enabled, thus producing an 8-bit parallel word.

As the output of the data latch can only change at bit count zero, a word valid latch is set whenever the word count moves from zero, indicating that the parallel data are valid. The eight lines of parallel data are buffered before being sent out from the serial/parallel converter.

It was indicated above that the synchronization word detector also indicates to the system that it is at word count two in the frame. In the *Meteosat* frame there is a total of 364 words, and the word counter is set to count 2 on detection of valid synchronization pulses.

4.2 Additional Facilities

Also built into the serial/parallel converter are a large number of tests, which are in the main of two types:

(a) those which detect specific words in the parallel data stream;
(b) those which detect specific counts in the word counter output.

These tests are simply carried out by hard wired NAND gates set to detect the appropriate codes. The results of these tests are available at all times on wires which are connected to the algorithmic state machine.

A block diagram of the serial/parallel converter is given in Fig. 4.

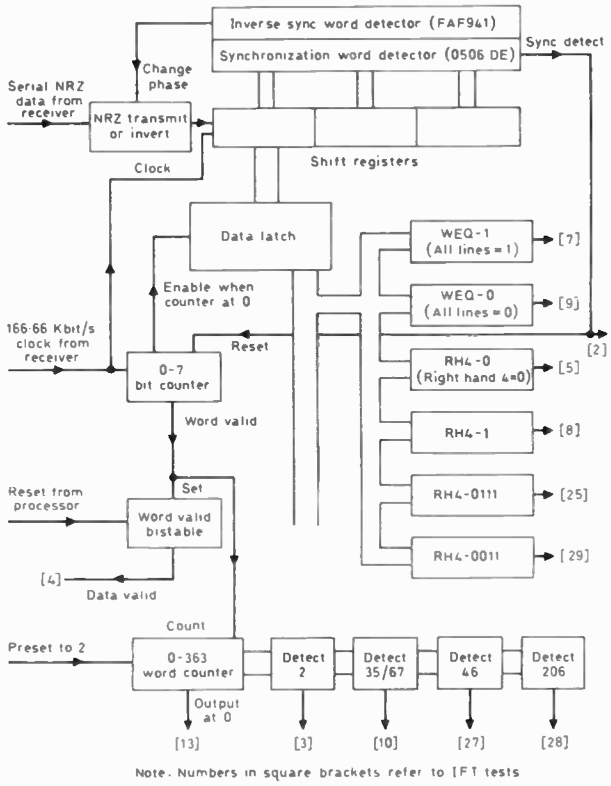

Fig. 4. Serial/parallel converter.

5 Algorithm State Machine System (ASM)

5.1 Block Diagram

A block diagram of the ASM logic system is shown in Fig. 5. It is contained on one card. (See also Ref. 8.)

The heart of the system is an 8-bit program counter connected to the address lines of two 256 × 8-bit programmable read-only memories. There are 16 parallel lines from these p.r.o.m.s, and every time the program counter changes, a different 16-bit word appears on the lines. This word is entered into a data latch so that it may be retained during changing of the program counter contents.

As can be seen from Fig. 5, the three left-hand bits contained in the p.r.o.m. decode into six possible instruction types by means of a 3 to 8 demultiplexer (two codes are not used). The next 5 bits are used to select a device by means of an address bus which is common to many (but not all) of the six instruction types. The right-hand 8 bits in the p.r.o.m. are connected to the LOAD inputs of the program counter (which is a synchronous pre-settable type).

5.2 Instruction Types

The first instruction is the INIT (initiate) instruction, normally encountered at the start of a program which sets all counters, latches (except program counter) to zero. The second instruction is the decision-making capability of the system and is the IFT instruction which means 'if specified test is true, go to AA', AA referring to

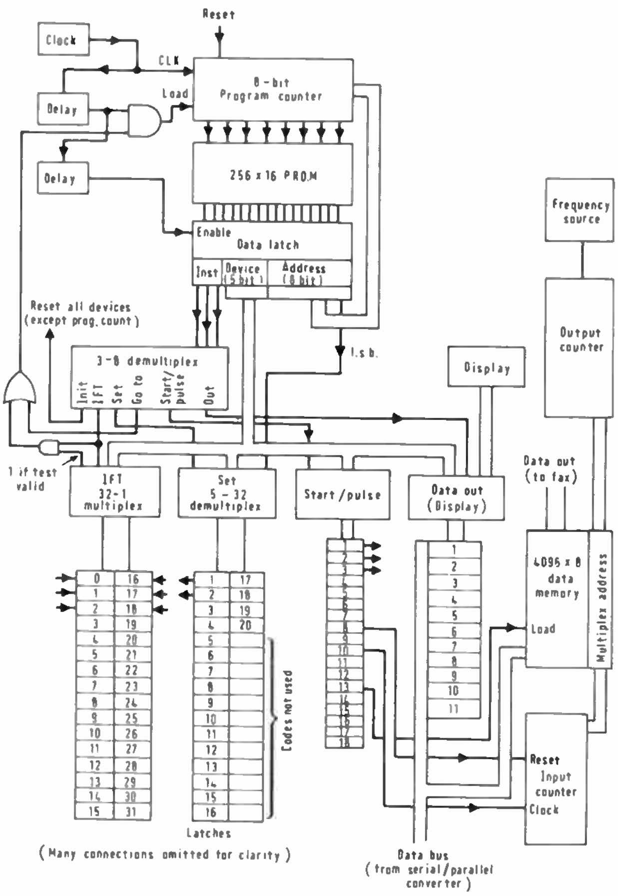

Fig. 5. ASM logic system.

the program counter address contained in the right-hand 8 bits of the p.r.o.m. The specified test is that which is addressed by the 5-device address lines. If the result of the test is false, then the program continues to the next step (i.e. the next count as specified by the program counter).

Each of the 5 address lines is connected to a 32–1 multiplexer, which has 32 single lines coming into it. Each of these lines is connected to a separate hardware test. In Fig. 6, all tests are labelled, e.g. RH4 1 means 'right hand four bits are all at 1' of the data.

The next instruction is the SET LATCH instruction which has the ability to set up to 32 separate latches to 1 or 0. These instructions produce an output from the system, and can, for example, light up a l.e.d. lamp labelled 'A format' indicating that this is the format of the data being received.

The DATA OUT instruction permits a numerical display of certain data contained in the label, which is a part of every sub-frame. The 8-bit data word is fed to a data latch, and because much of this data is in the form of 16-bit words, two sets of such latches are normally addressed by contiguous data words. The resultant binary data is connected to a 16-bit hard-wired binary-to-decimal converter, which can display integer numbers in the range zero to 65,535 (decimal) on a l.e.d. display. It is useful, for example, while monitoring a transmission to watch the line count as the transmission proceeds. The particular display chosen is determined by the operator who controls a set of switches on the ASM front panel; the data are retained from one label to the next (contained in every sub-frame) by the data latches.

The START/PULSE instruction decodes into 32 possible instructions, e.g. increment memory, reset all latches, etc. In fact only 19 of these possible instructions are used.

5.3 ASM Connections

The signal inputs to the ASM system are concerned solely with the IFT instructions. Many of the tests are hard-wired tests contained on the serial to parallel converter card, either to do with data, or to sense the word count. By this means, it is unnecessary to bring the 8 actual data lines on to the ASM card. The other tests concern the setting of the user-operated front panel switches (e.g. to see whether Vis, IR or WV has been selected by the operator) or the setting of the SET LATCH data latches.

The SET LATCH data latches are on the ASM card itself, resulting in no external connections.

The DATA OUT circuits (including the binary-to-decimal converter and display) are also on the ASM card.

The data memory, input and output counters, and grid memory are all contained on card 3. Therefore many of the connections from the ASM system go to card 3 to control the many functions which exist there.

The facsimile decoding connections are determined by the setting of some of the data latches on the ASM card. Connections are therefore required between these and card 3.

5.4 Speed of Operation of ASM System

The speed at which the ASM system operates is determined by means of a free-running R-C oscillator. The maximum theoretical frequency of this is governed by the switching time of the TTL circuits, and also by the rate at which the memory can be operated. It was, however, decided to use a much lower frequency which simplifies circuit design, and a frequency of about ten times the word rate was chosen. As the word rate is 20·8 kHz, the ASM oscillator was set to 200 kHz, which has been found adequate.

The output of the oscillator is a TTL square wave, and it is connected directly to the clock input of the program counter. The counter is clocked on the rising edge and the p.r.o.m. data latch is arranged to be enabled on the falling edge. The timing diagram is shown in Fig. 6.

A delayed 2·8 μs pulse is used to load the program counter, when required by the program, and to enable the instruction demultiplexer.

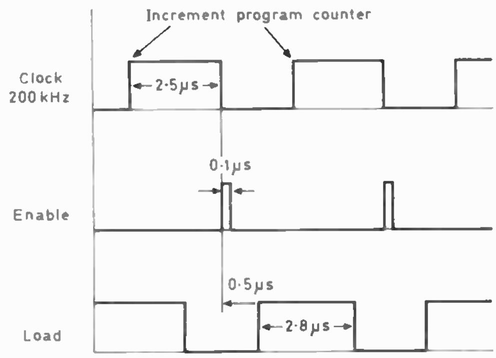

Fig. 6. ASM timing diagram. Sequence of events: (1) increment program counter: (2) enable data latch (also enables data out latches from memory): (3) load program counter (if instruction appropriate) enable instruction demultiplexer.
Notes: Data latch always valid except during (2). Load signal to counter must overlap $0 \rightarrow 1$ of clock.

6 Flow Chart

Figure 7 shows an overall flow chart to which the system has been designed. After the I/O reset, the system waits until valid synchronization pulses have been detected, at which point the next word (identification or ID word) is examined. If the right-hand three bits are zero, then the first frame is indicated, which contains the label. If the label is present, then the display data (subsequent words) are sent to the latches corresponding to their destination, and to the display requested according to the setting of the front panel switches. The word count is automatically incremented on the serial/parallel card every time a new word enters the data latches, and the program is arranged to examine each word in sequence.

Examination of words 5 to 8 indicates whether the current sub-frame is a header or data type. This information is entered in the header latch.

Examination of word 13 indicates the data format (i.e. A, B or X) and the appropriate latch is set at this point. Words 14 to 19 contain the information necessary to decide what data are present in the message. As explained in Section 3.2 these words have a different meaning depending on whether the sub-frame is a header or data sub-frame. If the sub-frame is a header, then the information which follows is ignored in the current design. These data contain information concerning calibration of IR data, deformation matrices, etc., which require a special memory and printer to output them. It was considered at the design stage that this could be performed by the existing system, but was not strictly necessary as far as displaying the images was concerned.

Thus, in this design, the header information after word 25 is ignored, and the flow chart indicates that the system keeps going back to the beginning until data sub-frames are detected, before taking any further action. As each data sub-frame has a label similar to that used for the header, this is used to identify which wavelength is in the sub-frame, and if it is the same as that requested on the front panel switches, then the data words are loaded into memory.

This operation only lasts for one sub-frame when IR or WV data are being requested, but with Vis data, two sub-frames are necessary for each line. This implies that the memory must not be reset to address zero after the first sub-frame, and therefore a system for detecting the presence of the first or second half of each line is required. This is accomplished by two bistables, one for Vis 1 lines and one for Vis 2 lines. Although it might have been possible to design a system with no bistables, since a unique code is transmitted for the second halves of lines, this information was not available during the original design exercise. The system as designed works satisfactorily, and there seems to be little point in changing it.

At the end of each sub-frame, the program causes the grid to be read into its memory. After reading in a complete line, the program initiates memory read-out, although this must start at such a time so that the output data flow is even, and not at an uneven rate as the input data flow can be.

7 Memory System

7.1 Overall System (Fig. 8)

The system memory consists of a random access memory of size 4096 × 8 bits. The memory is connected via its address lines to a multiplexer. The multiplexer is connected to two counters; one the input counter under program control, and the other the output counter. The

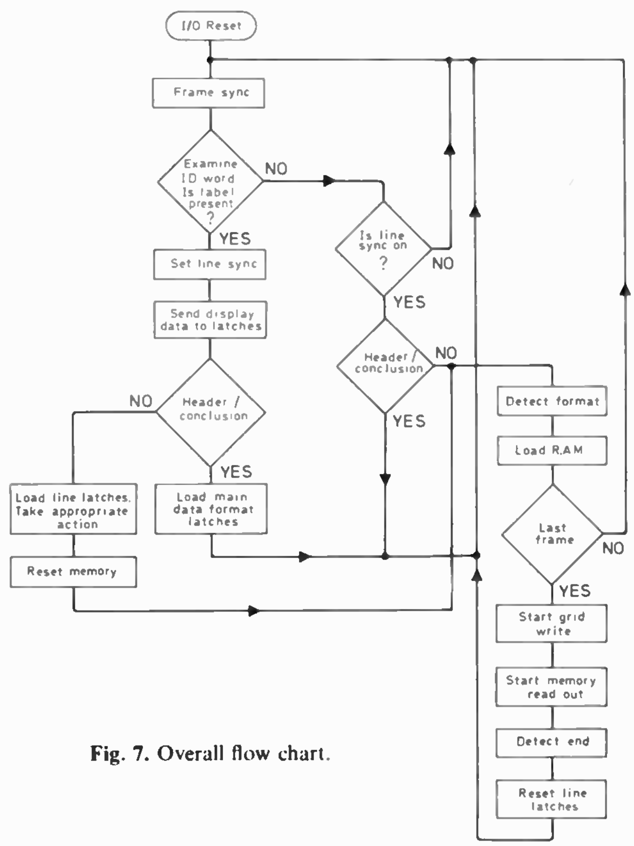

Fig. 7. Overall flow chart.

multiplexer is arranged so that the output counter is normally connected to the memory. The only time when the input counter is connected is during a LOAD RAM instruction which lasts for 2·5 μs.

The functions in the memory which are under program control are:

(1) reset input counter
(2) increment input counter
(3) load r.a.m.
(4) set start of output cycle.

The size of memory required is determined by the maximum length of a line. The maximum length of a line is 5000 words for A_V format, but by starting the read-out before the write operation is finished, a 4 K memory is adequate.

7.2 Input Counter

The input counter (see Fig. 8) is a 12-bit binary ripple counter. It is coupled to the memory via a multiplexer, and is connected to the memory just prior to and during write operations which are controlled by the ASM system as a separate instruction (increment r.a.m.).

The input counter is reset to zero at the beginning of each line. In the case of visible transmissions where the line is made up of two sub-frames, the program is arranged so that the reset only takes place on the appropriate alternate sub-frames.

The input counter is incremented at word rate, but there are interruptions at the end of each frame while the synchronizing word and ID words are presented to the system. Obviously these must not be loaded into memory.

7.3 Output Counter

The position in the cycle of operations at which the memory read-out starts is determined by the program. Because of the fact that the output frequency has to be precisely constant, the read-out frequency (and hence the output counter clock) are derived from a crystal oscillator and divider totally independent of the ASM system. The frequency of the read-out must clearly be correctly set in order that the output image has the correct width.

The point at which the memory stops reading out has also to be closely controlled. There are three possible line lengths: 1250, 2500 and 5000 words. There are, therefore, three separate gates which are automatically selected according to format, and which switch off the output bistable when the required number of memory samples have been output.

Because of the fact that the memory system overflows at 4096, the output counter has an extra bistable to encompass the count of 5000. On output count 4096 to 5000, the memory positions 0 to 904 are being addressed, these having been loaded with the correct data by the time they are being output.

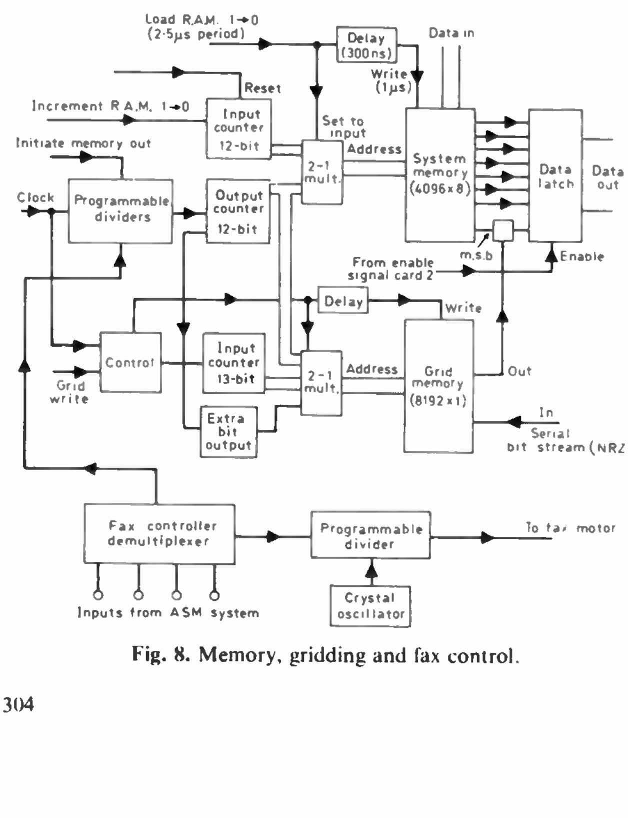

Fig. 8. Memory, gridding and fax control.

8 Gridding Circuits

8.1 System Design (see Fig. 8)

The grid data are contained in a serial bit stream which is placed immediately after the signal data words. Each bit in the grid data corresponds to a particular word—if the bit is zero then there is no grid at that point; if the bit is one then there exists a grid point. At the wish of the user, the grid may be superimposed on the image or the video printed without grid.

In order that the grid data are most visible, it is desirable that in areas where the video data cause the resultant tone to be dark, the grid is white and vice versa. This is accomplished by means of a set of comparators which measure whether or not the resultant video is greater or less than a certain grey level and replace the video sample with the appropriate shade.

8.2 Circuit Detail—Input Arrangement

A separate one-bit plane of memory is used for the grid data. The read-in of these data is initiated by the program, but because of the high speed of the data, the program does not have sufficient precision to stop it at the correct point. The data read-in is at clock rate (166·667 k bit/s).

The circuit is so arranged that the address lines of the grid memory are normally connected to the output counter of the main memory. When data are being read in, the last output sample is held by a data latch. Because the input and output frequencies are simply related, there can be no interference between the read and write cycles, and the input and output signals are transparent to one another.

8.3 Grid Output Arrangement

The addressing of the grid plane uses the same counter as the main memory output counter, but with one

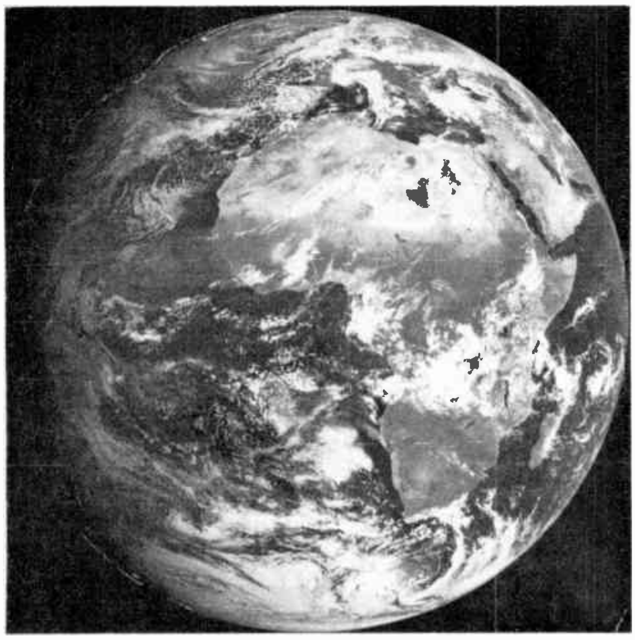

Fig. 9. A format Visible channel.

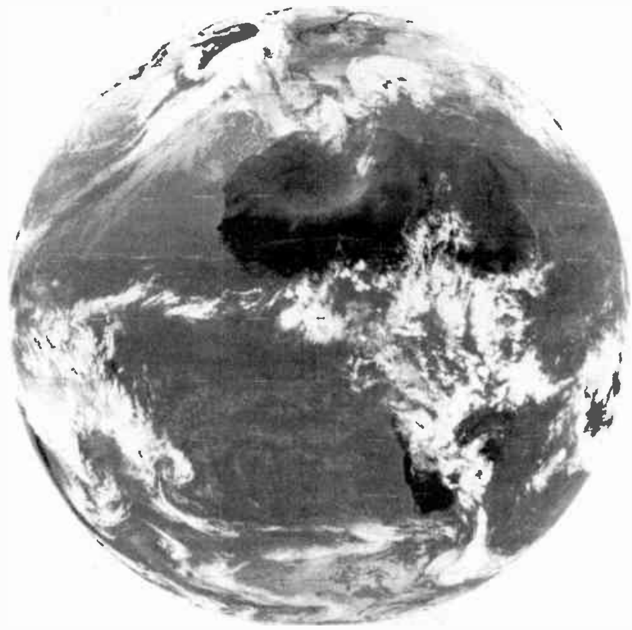

Fig. 10. A format Infra-red channel.

important difference. This is that one extra bit of that counter is caused to address the second chip of the grid memory, and not to return to the first bits as is done with the main memory.

9. Interface to Facsimile Machine

9.1 Output Frequency Calculations

As explained in Section 3.1, the facsimile machine used has two speeds: 429·258 lines/min and 343·407 lines/min. It has been shown also in Section 3.1 that the rates at which the various data channels appear are determined by the specific transmission formats used.

In order to print the data correctly on the machine running at these speeds, it is often necessary to print for one revolution, and then to blank the machine for several lines. For example, when A_{IVW} is being transmitted and the IR channel is selected, from Table 3 we can see that the output line rate is 107·3148 lines/min. This is one-quarter of the facsimile machine rate, therefore the output frequency used should be such that a complete line of data is read out during one revolution. The machine is then blanked (set to white) for the following three revolutions. The memory output circuit is arranged automatically to read out the correct length of line at the appropriate frequency. When the output bistable is reset, the facsimile machine is blanked by a switch arranged in the output circuits.

9.2 Method for Smoothing Data Flow

In certain formats, e.g. B_{IV}, the visible data channel is read into memory in such a way that there is an unequal gap between the alternate lines. This is because after the Vis 2 line has been decommutated, there is a line of IR. Therefore, we have four contiguous sub-frames of visible data (making two lines) followed by a gap.

In these cases, it has been found necessary to produce a special timing circuit divided down from the clock, which switches on the output bistable at equal intervals chosen so that the average line rate remains constant. It is important that this timing circuit is correctly phased, so that the data start to read out at the appropriate point in the memory cycle.

In order to ensure that all formats are correctly read out, two data selectors controlled by the A latch, Vis request switch, Vis present and W present decode to give the correct frequencies and phasing data.

10 Signal Outputs

10.1 Processing and Display Arrangement

The memory read-out is connected to a data bus which can be taken to any digital parallel entry device, e.g. computer compatible tape machine. There is also a digital-to-analogue converter output, which is connected to the facsimile machine and also to an oscilloscope monitor, both via a special processing circuit.[9]

11 Results

11.1 A Format Examples

Examples of A_V and A_I are shown in Figs. 9 and 10. The resolution of the A_V formats can be observed to be greater than the A_I formats. For the A_V formats, the processing (see Section 10.1) was left constant at less than 5% (i.e. virtually no processing). To produce the optimum contrast for A_I and A_W formats, the processing was continuously adjusted during the transmission. The annotation can be observed on the top right of the images, and is clearly rather small in the case of A_V.

11.2 B Format Examples

The B format images (Figs. 11 and 12) can be observed to be sectorized portions of the complete Earth disks. By

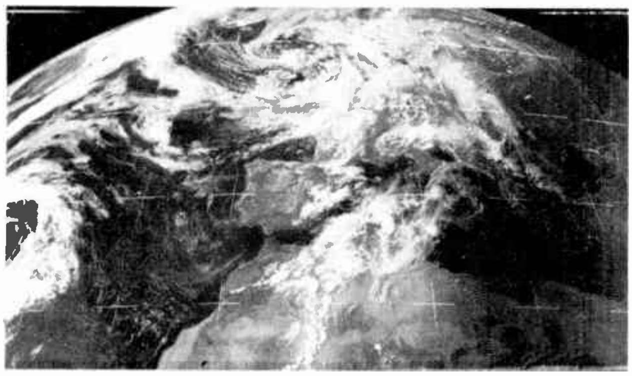

Fig. 11. B format Visible channel.

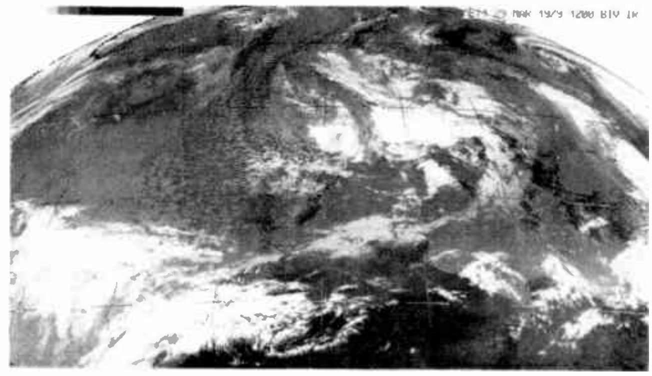

Fig. 12. B format Infra-red channel.

adjusting the line feed of the facsimile machine, the formats have been expanded slightly in a North–South direction, which gives a slightly better projection at Northern latitudes.

11.3 X Format Examples

Figure 13 shows an X_V image, transmitted from the American *GOES* satellite (75°W) to a receiving station at Lannion, France, after which the data is retransmitted to the *Meteosat* satellite for relaying to PDUS.

12 Conclusions

The ASM system has been shown to be a suitable system for decommutating *Meteosat* signals. The system produces an economical means of providing primary data images in hard copy form. The output provided could also be fed to a video display unit, if it were desired to display it on a video monitor.

Without a tape recorder, it is not possible with this system to display all types of data from one transmission, i.e. when printing IR from a B_{IW} transmission, the water vapour channel cannot be simultaneously printed (unless a second ASM system were produced). A digital tape recorder would permit the alternate channel to be printed at the end of the transmission.

Fig. 13. X format Visible channel.

13 Acknowledgments

The author acknowledges gratefully the provision of *Meteosat* serial digital signals by P. E. Baylis, who designed, constructed, and commissioned the receiver, and provided much help during the testing phase of the ASM system. Thanks are also due to A. A. Dickie, who provided much help in the initial stages of the design, in particular by his suggestion of the algorithmic state machine method to implement the system.

Acknowledgment is also due to Ferranti Ltd, Manchester, who provided a parametric amplifier unit to give the necessary low noise signals.

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A superposition-based analysis of pulse-slimming techniques for digital recording

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

SUMMARY

The application of pulse-slimming to digital magnetic recording is investigated, and analysed using superposition. Representative criteria are used to determine the maximum achievable packing density both before and after slimming. The results indicate that pulse-slimming is of little value for an already-optimized recording system, but could be used to trade-off timing margin against amplitude margin in a new design.

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1 Introduction

The superposition technique offers an opportunity for many facets of the magnetic recording process to be analysed in non-real time, allowing the recording and replay mechanism to be effectively magnified for greater insight into the detailed changes produced by variation of any of the parameters involved, such as coding technique, packing density, or detection process.

The superposition principle, as applied to magnetic recording, states: 'At all packing densities for which the read-back process is linear, the net flux in the read-coil from any pattern of surface flux-reversals is the algebraic sum of the individual flux contribution from each reversal acting on its own'. This principle provides a very simple means of simulating the effects of any pattern at any packing density, since the isolated reversal response, called here 'the basic pulse', can be stored on a computer as an array of voltage readings, and then any number of these basic pulses can be added or subtracted, at the correct distances from each other, to produce the total output voltage waveform. Measurements can then easily be made on this output waveform to calculate peak-shift, amplitude, etc.

The only phenomena which will render superposition invalid are those which alter the written transition shape in a manner dependent upon the transition density. Morrison and Speliotis[1] report this range of validity to go up to 60 000 bits/in, a packing density out of reach of current technology. Other authors have suggested that this figure is too high, but the alternative to superposition is the dynamic iterative hysteretic model[2-4] which, although more accurate than superposition, particularly at very high packing densities, involves many times the computational effort, and was therefore not considered for this study because of the large number of permutations involved in the analysis.

2 Choice of the Basic Pulse

The heart of the superposition process is the basic pulse, and this must naturally be chosen very carefully. Several expressions have previously been chosen to represent the basic pulse analytically. Hoagland[5] proposed the Gaussian expression $v(t) = \exp(-t^2)$ which was also used by Chu,[6] as well as the Lorentzian $v(t) = 1/(1+t^2)$, used also by Kosters and Speliotis.[7] The mathematical justification for the latter is that the Lorentzian is the derivative of the arctangent function, which has widely been assumed to be a good representation of the magnetization distribution in an isolated transition region.

Sierra[8,9] has also used the Gaussian expression, whilst Jacoby[10] modified this to $v(t) = \exp(-t^{1\cdot 6})$. Several other expressions were also considered by the author.

All the results from the superposition program are normalized to the width of the basic pulse at 50% of the maximum amplitude, i.e. to PW_{50}. The latter is now

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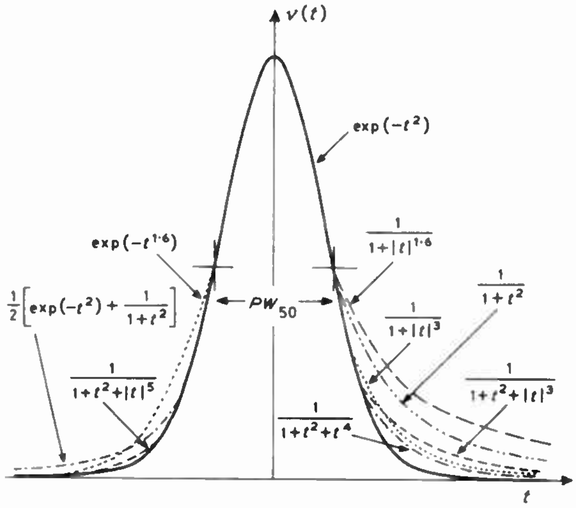


Fig. 1. Analytical basic pulses.

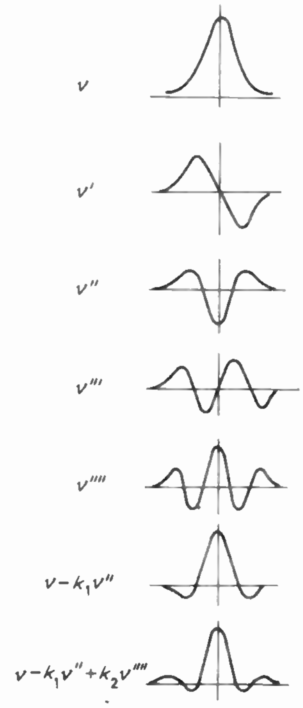


Fig. 3. Pulse-slimming by addition of derivatives.

universally accepted as a sound basis of comparison between pulses; it is easily measured in practice due to the high slope of the pulse in this region, and produces much less error when directly comparing pulses than does the more obvious alternative of basewidth, i.e. PW_0.

The nine basic pulses used are plotted in Fig. 1. Only one curve is given above the PW_{50} point for clarity, as the curves are all very close in this region. It can be seen that the expressions account for almost any shape of symmetrical pulse likely to be encountered, although even an asymmetrical one can be simulated by using different expressions on each side of the origin.

Several different currently-available memories were used to compare the analytical pulses against, though not simply by comparing practical basic pulse shape against theoretical one, as this cannot be done accurately. Instead, for each of the memories available, graphs were plotted of 'all ones NRZ1 amplitude' and 'two ones NRZ1 peak-shift' against packing density, and similar graphs were produced for each analytical expression using a superposition program. The theoretical graphs were then compared with the practical ones for both location and fit. The clear winner in this comparison was found to be $1/(1+t^2+t^4)$, with $1/(1+t^2+|t|^3)$ and $\exp(-t^2)$ fairly good, but the great surprise was that the Lorentzian came out very poorly.

3 The Pulse-Slimming Principle

Since superposition is normalized to PW_{50}, the implication is that the maximum packing density achievable by any code and/or detection system is inversely proportional to PW_{50}. The possibility of slimming the read-back pulse therefore implies an increase in the packing density and storage capacity of a given store. This must certainly be true if the slimming is effected 'mechanically', e.g. by reducing the head-to-surface separation or oxide coating thickness, but the validity of the theory of superposition also allows the slimming to be performed electronically, after the data waveform has been read back from the surface. The pulse-slimming filter then merely represents an extra block in the recovery chain, as shown in Fig. 2.

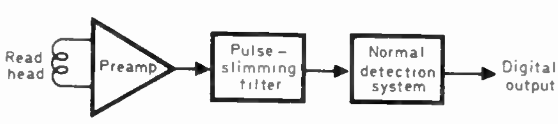


Fig. 2. The waveform recovery chain.

4 Addition of Derivatives

Figure 3 shows how a symmetrical pulse (v) suffers a reduction in PW_{50} by the subtraction of its second derivative (v''), in the correct proportion, but also contains significant baseline undershoot. The further addition of a proportion of the fourth derivative (v'''') to this reduces the undershoot, but also introduces overshoot as shown. If the initial pulse is substantially asymmetrical, odd-order derivatives may be applied to correct this, though this extra complication will not be considered here.

Figure 4 shows the effect of the addition of $-v''$ to v in various proportions, using the superposition program. The two pulses are first normalized so that the peak amplitude of each is unity. They are then added, and the resulting slimmed pulse is also normalized. Its PW_{50} and undershoot amplitude are then measured.

4.1 Differentiation

Of particular importance to this pulse-slimming technique is the differential process. The most accurate way of producing the derivatives of the read-back

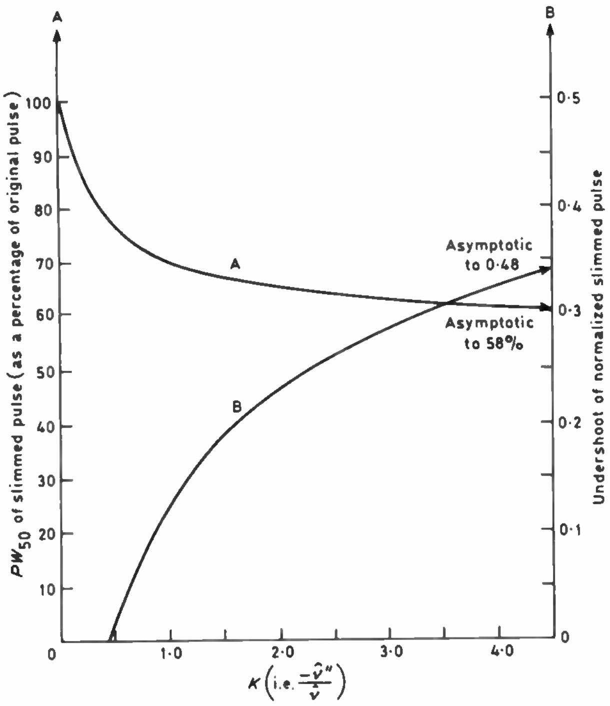


Fig. 4. PW_{50} and undershoot of slimmed pulse.

waveform involves the use of a delay line. This process can be analysed as follows:

Consider a small portion of the read-back waveform (Fig. 5).

If $v = f(t)$, then

$$v + \delta v = f(t + \delta t)$$

so

$$\delta v = f(t + \delta t) - v$$
$$= f(t + \delta t) - f(t).$$

Therefore

$$\delta v/\delta t = [f(t + \delta t) - f(t)]/\delta t$$

or

$$\mathrm{d}v/\mathrm{d}t = \lim_{(\delta t \to 0)} \{[f(t + \delta t) - f(t)]/\delta t\}.$$

This shows that the derivative of the read-back waveform can be formed by subtracting from it a delayed version of itself, and the shorter the delay is, the more accurate will be the differentiation. Unfortunately, any noise superimposed on the signal which is of a higher frequency than the signal fundamental, but not high enough that it can be filtered off, will be doubled in the

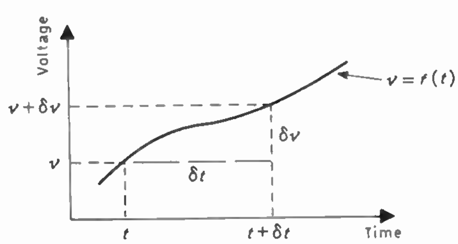


Fig. 5. Read-back waveform.

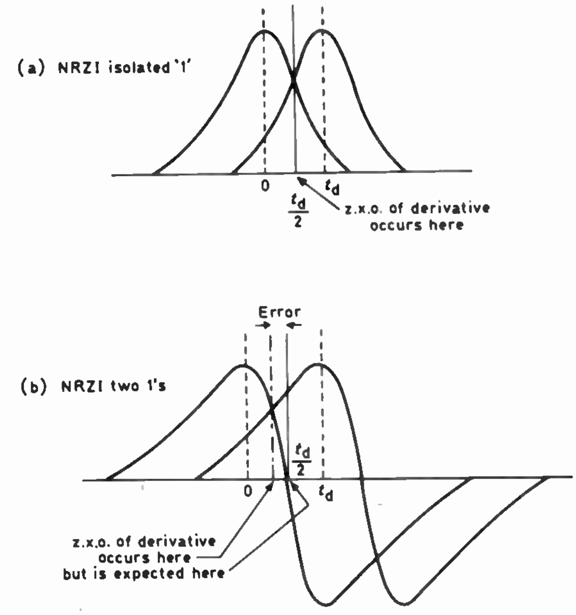


Fig. 6. Worst-case timing error due to non-ideal differentiation.

worst case, as its period will be much less than the delay, and one noise peak could reinforce another. Since the amplitude of the signal derivative falls as the delay is reduced, a compromise must be found between accuracy of differentiation and signal-to-noise ratio.

Figure 6 shows how the accuracy problem arises in practice. For an isolated basic pulse as in (a), the zero-crossover (z.x.o.) of the derivative always occurs at $t_d/2$ for all values of t_d (the delay between the two signals), assuming the pulse is symmetrical. For the worst case pattern of two ones NRZ1 (two isolated transitions), shown in (b), however, the steep gradient on one side of the peak and the gentle gradient on the other combine to give a z.x.o. which is not at $t_d/2$. It is apparent that reducing t_d reduces the error. It should also be noted that, for a given t_d, increasing the packing density will increase the error, as the two gradients mentioned will differ by even more.

The superposition program can again be used to give quantitative answers to this effect. The results of this analysis are shown in Fig. 7, which plots the maximum timing error (as a function of PW_{50}) and the peak amplitude of the derivative of the normalized basic pulse for all values of delay up to $1{\cdot}0 \times PW_{50}$.

It is suggested that a suitable trade-off between accuracy and S/N ratio results from using a delay of $0{\cdot}3 \times PW_{50}$. This yields a maximum timing error of 0·5% PW_{50} at $PF = 1{\cdot}5$, and a peak signal of 0·4 after differentiating an isolated normalized basic pulse. Since the noise has doubled, note that the S/N ratio has been reduced by 14 dB, though the actual extent to which this loss is felt depends on the particular implementation involved. It is clear, however, that the second derivative will have a very poor S/N ratio, and clearly use of the fourth derivative, whilst beneficial in theory, will not be

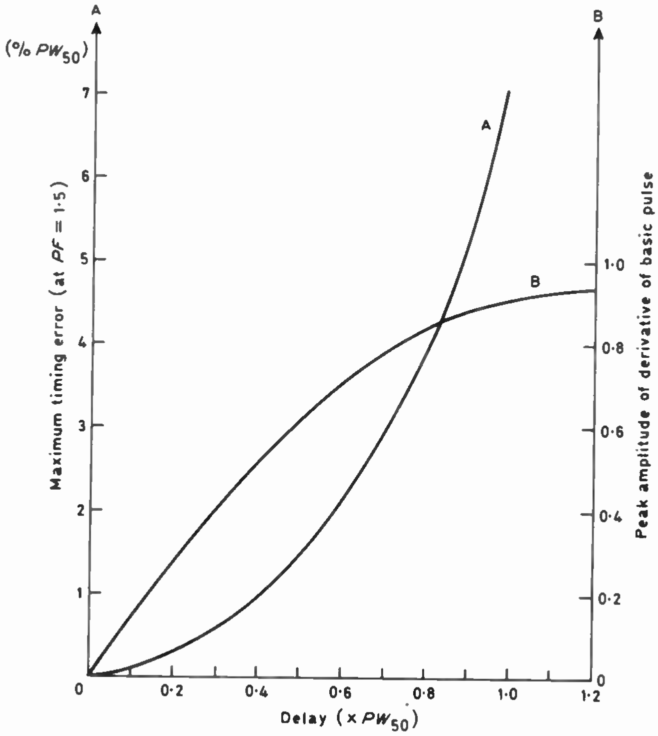


Fig. 7. The effect of using different length delay lines for differentiation.

sensible in practice.

It should be noted that the value of delay suggested above should be selected with reference to the worst-case PW_{50} in a particular system. This means that for pulses with smaller PW_{50}, the effective delay to them is greater than optimum, resulting in a greater timing error, but because these pulses are non-worst-case to start with, they should be able to accommodate this extra error.

4.2 Implementation

The circuit shown in Fig. 8 shows an experimental implementation of this pulse-slimming technique. The short-circuited lumped delay lines perform the differentiation. The delay in the path of the input pulse is to align it correctly with the doubly-differentiated one, which is delayed by $2 \times t_d$ (where t_d is the length of the delay line) relative to its input. The delays are variable (in finite steps) to enable them to be set correctly relative to the PW_{50} of the input pulse. The variable-gain amplifier, the inverter and the summer can all be effected by means of a high-bandwidth dual-beam oscilloscope with trace-addition facilities.

4.3 Effect on Achievable Packing Density

It was shown in Fig. 4 that the maximum possible reduction in PW_{50} using only the second derivative is 42%, so it would not seem likely that packing density increases will exceed this figure. Indeed, if the slimmed pulse was exactly the same shape as the original pulse, and the S/N ratio was unaltered, the calculation would be as simple as that, but the very complex nature of the slimmed pulse means that only detailed practical or theoretical analysis can determine the exact effect on the achievable packing density.

The first step in this process is to find a packing density limit for a given system before slimming is applied. In this instance, this was done using the superposition program, by calculating worst-case peak-shift and amplitude data for a hypothetical recording system using NRZ1, with a read-back S/N ratio of 20:1 (26 dB), i.e. isolated basic pulses have a peak amplitude of unity, whilst the noise has a peak amplitude of 0·05. The detection system postulated was a typical rectify-and-clip process, as shown in Fig. 9. The read-back waveform is first amplified—this may be automatic gain-controlled amplification, but, if so, perfect a.g.c. action will be assumed. The signal is next full-wave rectified, and then clipped to remove baseline noise, which would otherwise be a problem later on, in the squaring process. A certain amount of the noise could be filtered out by correct choice of the frequency response of the linear amplifier, but a problem arises when the noise has components at frequencies lower than the maximum significant frequency in the data. Attempted filtration of these components will result in integration of the data waveform, having the effect of increasing the PW_{50} of the basic pulses, which clearly limits performance.

Waveforms pertinent to this detection system are shown in Fig. 10 for the worst-case amplitude pattern of

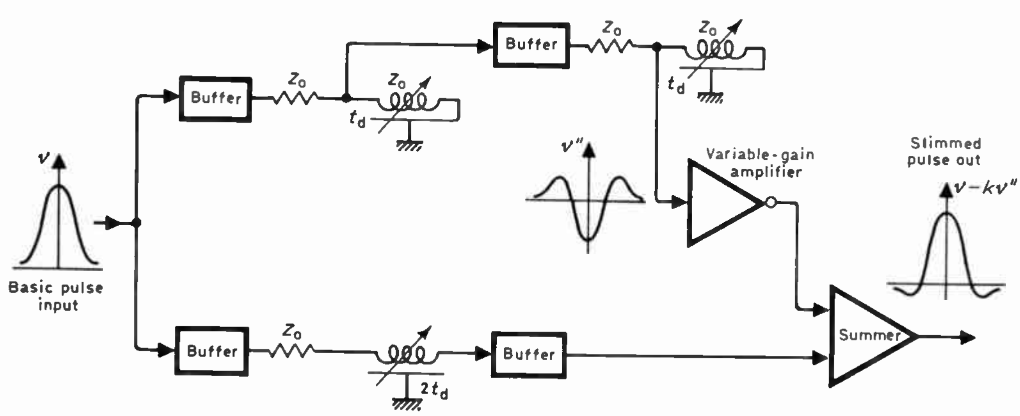


Fig. 8. Active pulse-slimming filter (by addition of derivatives).

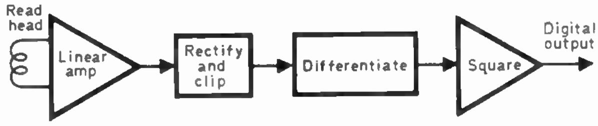
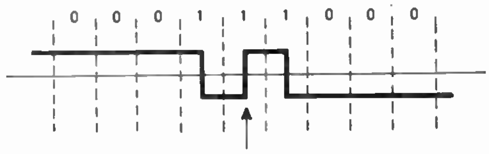


Fig. 9. Rectify-and-clip detection system.

three isolated ones. For the determination of the absolute limit of the system, no margins are allowed, and so the clip level is set to the noise level of 0·05. The differentiation circuit uses a delay line of total delay $0{\cdot}3 \times PW_{50}$, as described earlier. The signal is finally squared to exaggerate the zero crossovers, which are then detected using the ideal timing window of half a bit period.

The packing density achievable using this detection system may be limited by timing problems or amplitude problems. An amplitude limit will occur where the worst-case read-back signal falls to the clip level of 0·05. Figure 11 is a plot, computed by superposition, of the minimum read-back signal against packing factor (*PF*), where *PF* is the packing density normalized to a bit period (*BP*) equal to PW_{50}, i.e. $PF = PW_{50}/BP$. The peak producing this worst-case signal is the centre'1' of three isolated 1's, i.e.



It can be seen from the graph that zero amplitude margin occurs at $PF = 1{\cdot}88$.

A timing limit will occur where the worst-case peak shift plus the differentiator error equals half the timing window. Figure 12 shows the worst-case NRZ1 peak-

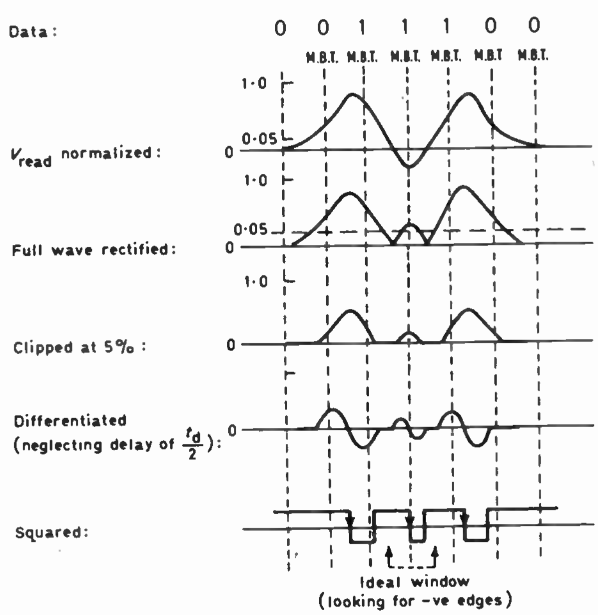


Fig. 10. Rectify-and-clip detection waveforms.

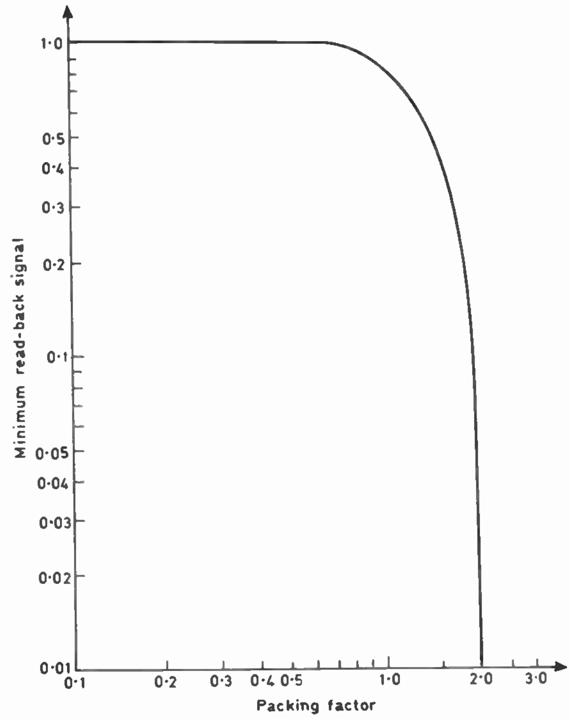


Fig. 11. Minimum read-back signal vs. packing factor for NRZ1 'three-ones'.

shift (for a two 1's pattern) against *PF*, as calculated by superposition. Zero timing margin occurs where peak-shift = 50% *BP*, i.e. at $PF = 2{\cdot}33$. Clearly, the theoretical limit for this code with this detection system occurs at $PF = 1{\cdot}88$ (due to the three 1's pattern).

No allowance has been made so far for clocking inaccuracies due to such factors as crosstalk, differentiator error, incomplete erasure, particle noise and phase-lock-loop errors, all of which can occur in a practical system. Additionally a practical system would always work with a certain margin on top of the known inaccuracies. Taking a figure of 8% PW_{50} as a reasonable allowance for inaccuracies plus margin, it is possible to calculate a new timing limit for the system. A curve

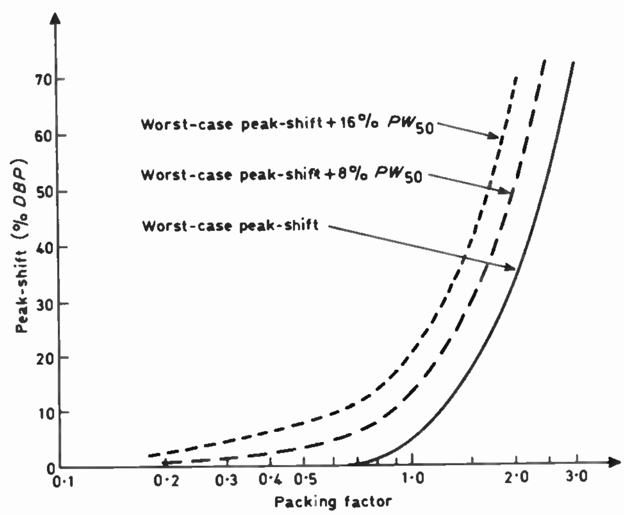


Fig. 12. Worst-case peak-shift for NRZ1.

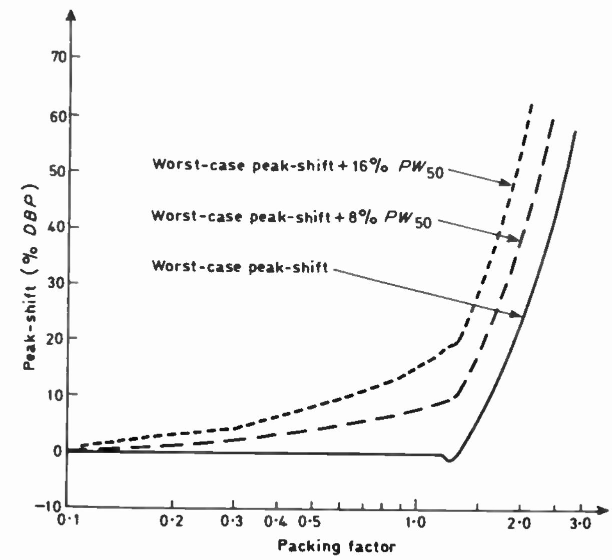


Fig. 13. Worst-case NRZI peak-shift for the slimmed pulse ($v - v''$).

representing this real-time error (*RTE*) on top of the worst-case peak-shift is shown in Fig. 12, from which it can be seen that the new timing limit is at $PF = 1{\cdot}92$. A similar construction can be performed for any value of *RTE*, and if 16% PW_{50} were allowed, for example, the timing limit would be at 1·62.

To summarize these results for clarity:

(a) $RTE = 0$: Although the timing limit is 2·33, an amplitude limit occurs first at 1·88.
(b) $RTE = 8\%\ PW_{50}$: Again, although timing does not limit performance until 1·92, an amplitude limit occurs at 1·88.
(c) $RTE = 16\%\ PW_{50}$: Timing causes breakdown first in this case at 1·62.

Consider now a slimmed pulse based on the ratio $v:v'' = 1:k$. Assuming, as before, a read-back noise amplitude of 0·05, then allowing again for a doubling of the noise in each differentiator, it can easily be calculated that the peak noise amplitude out of the slimmer is $[0{\cdot}05 + 0{\cdot}47k]/(1+k)$, after normalization of the slimmed pulse. For $k = 1$, the noise amplitude is 0·26 (compared with undershoot at this stage of 0·12).

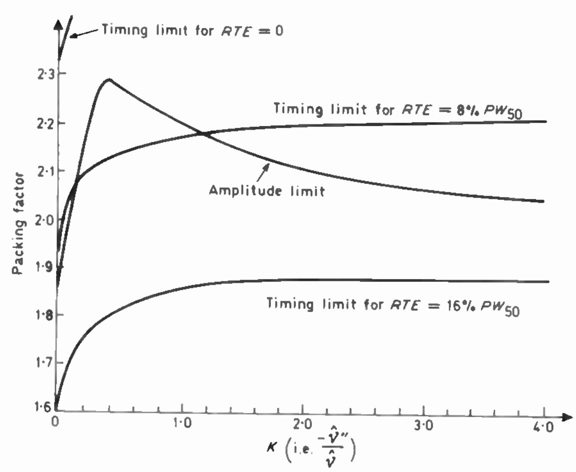


Fig. 14. Timing and amplitude limits after pulse slimming.

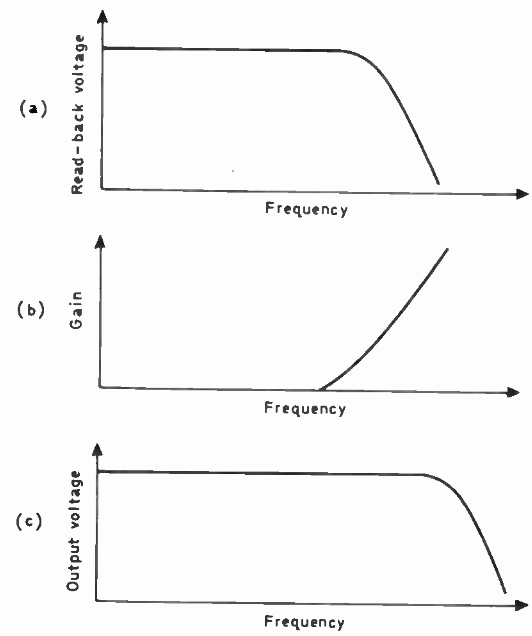


Fig. 15. Pulse-slimming by amplitude compensation. (a) Replay response; (b) Filter characteristic; (c) Improved replay response.

The clip-level in the detection system must be set to (at least) the undershoot amplitude plus the noise amplitude to avoid erroneous triggering, i.e. clip-level to be $0{\cdot}26 + 0{\cdot}12 = 0{\cdot}38$. An amplitude breakdown occurs at $PF = 2{\cdot}20$, where the amplitude of the all-ones pattern (not the usual three 1's pattern), after slimming, falls to 0·38. The worst-case peak-shift pattern changes also. It becomes the three 1's pattern, whose outer peaks' peak-shift is plotted in Fig. 13.

The new packing density limits can be summarized as follows:

(a) $RTE = 0$: Although the timing limit is at 2·57, an amplitude limit occurs at 2·20 (cf. 1·88 before slimming).
(b) $RTE = 8\%\ PW_{50}$: The timing limit now precedes the amplitude limit, and is at 2·15 (cf. 1·88).
(c) $RTE = 16\%\ PW_{50}$: Again the timing limit causes breakdown, at 1·86 (cf. 1·62).

The same analysis can now be applied to other slimmed pulses to determine an optimum value of *k*. Figure 14 shows the results of this analysis, and it can be seen how *k* can often be chosen to produce simultaneous amplitude and timing breakdowns, thus maximizing the system margins. Summarizing:

(a) $RTE = 0$: Although the timing limit is never less than 2·32, an optimum amplitude limit occurs for $k = 0{\cdot}4$, at $PF = 2{\cdot}29$.
(b) $RTE = 8\%\ PW_{50}$: The optimum value of *k* is 1·2, yielding simultaneous amplitude and timing margins at 2·18.

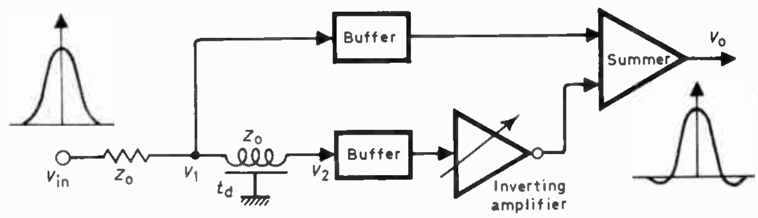


Fig. 16. Active pulse-slimming filter (by amplitude compensation).

(c) $RTE = 16\%\ PW_{50}$: The amplitude limit is greater than the timing limit for all values of k, and the timing limit $= 1{\cdot}88$ for $k > 1{\cdot}6$, so the best choice is $k \simeq 1{\cdot}6$, yielding maximum amplitude margin.

The striking shape of the curves in Fig. 14 for $k < 0{\cdot}4$ deserves some explanation. All the curves show a significant improvement in this area, which then rapidly reduces, or even reverses. Reference to Fig. 4 again shows why this happens: the largest improvement in PW_{50} occurs for $k < 0{\cdot}4$, after which point not only does the slimming improvement reduce, but also undershoot commences, and increases fairly rapidly. It is apparent then that $k = 0{\cdot}4$ will generally prove to be optimum, over a wide range of RTE values, resulting in typical packing density increases of 15%.

5 Amplitude Compensation

The basis of this method is the all-ones response shown in Fig. 15(a). If the read-back signal is passed through a filter having the complementary response shown in (b), the result will approach the characteristic shown in (c), giving a greater bandwidth. Since the 3 dB point is being increased, the implication is that inter-symbol interference is being reduced and thus peak shift will be reduced.

One of the best ways of achieving the required gain characteristic is by the use of transversal filters. This is the name given to a class of transmission line devices which afford constant delay, or linear phase, filtering. In practice, a simple lumped delay line provides this effect. A typical implementation of this technique is shown in Fig. 16. The circuit can be easily analysed:

$$V_1 = V_{in} \times Z_s/(Z_s + Z_0).$$

where Z_s is the sending-end impedance of the delay line, and Z_0 is its characteristic impedance

$$V_2 = V_1/\cos\theta$$

where $\theta = \omega t_d$, and t_d is the electrical length of the line. Therefore

$$\begin{aligned} V_0 &= V_1 - kV_2 \\ &= V_1 - kV_1/\cos\theta \\ &= V_1(1 - k\sec\theta) \\ &= V_{in}(1 - k\sec\theta) \times Z_s/(Z_s + Z_0) \end{aligned}$$

Since

$$Z_s = -\mathrm{j}\, Z_0 \cot\theta$$

then

$$\begin{aligned} V_0/V_{in} &= -\mathrm{j}\cot\theta(1 - k\sec\theta)/(1 - \mathrm{j}\cot\theta) \\ &= (k - \cos\theta)(\mathrm{j}\sin\theta - \cos\theta). \end{aligned}$$

Therefore

$$|V_0/V_{in}| = k - \cos\theta.$$

This has the cosinusoidal form shown in Fig. 17, from which it is apparent that for $\omega t_d < \pi$, i.e. $\omega < \pi/t_d$, the required frequency response is obtained.

However, the circuit can also be analysed in a different manner. With reference to Fig. 15 again:

$$V_1(t = 2t_d) = V_1(2t_d) = 0{\cdot}5V_{in}(2t_d) + 0{\cdot}5V_{in}(0)$$

and

$$V_2(2t_d) = V_{in}(t_d)$$

so

$$\begin{aligned} V_0(2t_d) &= V_1 - kV_2 \\ &= 0{\cdot}5V_{in}(2t_d) + 0{\cdot}5V_{in}(0) - kV_{in}(t_d). \end{aligned}$$

Letting $k = 1 + c/2$,

$$\begin{aligned} 2 \times V_0(2t_d) &= V_{in}(2t_d) + V_{in}(0) - 2V_{in}(t_d) - cV_{in}(t_d) \\ &= (V_{in}(2t_d) - 2V_{in}(t_d) + V_{in}(0)) - cV_{in}(t_d) \end{aligned}$$

Therefore,

$$2V_0 = V''_{in}(0) - cV_{in}(t_d).$$

This method is therefore shown to be exactly the same as the previous pulse-slimming technique, involving the addition of v and v'' in the appropriate ratio (remember that v is delayed by t_d to align it correctly with v'').

6 Lattice Filters

This is, notionally, a different method, involving postulating the effect of a filter on the isolated pulse, deducing the transfer function of the network, and then, by Laplace transform techniques, realizing the filter with inductors, capacitors and resistors. In effect, however, the method can be viewed as another attempt at amplitude compensation without recourse to transversal filters.

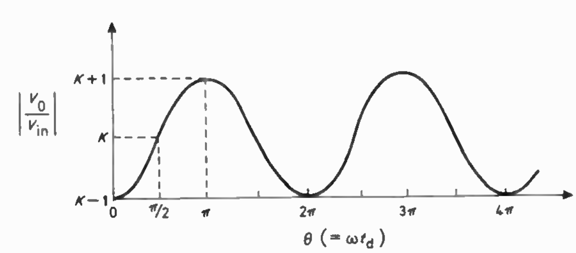


Fig. 17. Pulse-slimming filter characteristic.

Sierra[9] has described a passive, symmetrical lattice filter for slimming the width of a Gaussian pulse by a factor of two. Although the design aim was a 50% reduction in PW_0, and presumably PW_{50}, the results show a much less reduction achieved, which would seem to indicate that the limits of this method are akin to those of the two methods already described. Also, since the lattice contains 24 accurate passive components, including inductors, and is a delicately balanced bridge structure, it would seem a less attractive technique than the others.

A simple active equivalent to the symmetrical lattice was investigated by Dodd[11] and by Whitehouse.[12] The active circuit reduces the number of complex filter-arm impedances by a factor of four, which clearly benefits a parallel system. None of the three authors mentions the effect of the filter on noise or undershoot, but it would appear from the photographs supplied that this method is hampered by these effects to the same extent as in the addition of derivatives technique. Also, the three authors appear intent on reducing the pulse-width as much as possible. As has been shown, a PW_{50} reduction of 42% is possible (and maybe more using, say, the fourth derivative), but this is not necessarily the optimum choice.

7 Slimming of Asymmetrical Pulses

Many magnetic recording devices, particularly 'in-contact' ones, produce markedly asymmetrical pulses, due primarily to the shape of the transition region in the media, and the influence of the vertical component of magnetization from the media.

A similar analysis to the foregoing shows that the improvements to be obtained on asymmetrical pulses are not as great as those which can be achieved with symmetrical ones. In general, the more asymmetrical the pulse, the less the improvement achievable by slimming. This can be explained with reference to Fig. 18. With a symmetrical pulse, the undershoots on each side of the slimmed pulse are of equal amplitude, and since it is the maximum value of the undershoot which contributes to breakdown, this is optimum. With an asymmetrical pulse of equal PW_{50}, then, after slimming, one undershoot is smaller, and one larger, than those for the symmetrical pulse. This causes earlier breakdown. One other way of viewing this is that an asymmetrical pulse is already an asymmetrically slimmed version of a symmetrical pulse, thus leaving less margin for extra slimming.

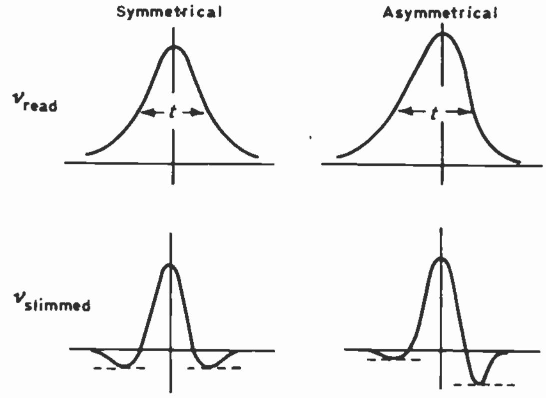


Fig. 18. Comparison of slimmed symmetrical and asymmetrical pulses.

8 Conclusions

It has been shown how superposition can be effectively used to analyse many facets of the magnetic recording process. An analysis of pulse-slimming techniques has shown that only small increases in packing density are possible by their use. Improvements up to 20% are obtainable, though it is doubtful whether such a small improvement is worthwhile for a system already in operation. It may be beneficial, however, to use pulse-slimming to trade off amplitude margin against timing margin, or vice-versa, particularly in the design of a new system, where it may also be possible to radically alter the detection process to cater for the peculiar type of waveform distortion produced by the pulse-slimming.

9 Acknowledgments

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The use of single-frequency Decca Navigator signals for remote position monitoring

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SUMMARY

The paper describes an extension to the Tracer technique for monitoring and tracking the positions of remote and unattended objects, allowing the signals transmitted by Decca Navigator stations to be used. The problems of retransmitting the multi-frequency Decca Navigator signals have been overcome by receiving a single frequency and decoding the time-multiplexed 'multipulse' signals in time-division form.

A number of applications of the system are described and results presented. The accuracy, fix repeatability and skywave sensitivity of the system are seen to compare favourably with those of conventional receivers.

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1 Introduction

Many widely-used radio-navigation systems are of the phase-comparison, hyperbolic type.[1–4] Their operation is best understood by considering that signals are transmitted simultaneously by a chain of spaced transmitting stations. A receiver measures the phase differences between the signals received from pairs of stations and its position is established by referring to a hyperbolic grid of lines of position, each of which is a locus of constant phase difference.

To allow the transmission of the various stations to be separated and identified they must be multiplexed in some way. Time division multiplexing has been adopted in several systems, the various transmitters operating in sequence on a common frequency.[1–3] The receiver and transmitters are synchronized by having one station transmit a signal with clearly-identifiable characteristics such as a frequency shift or a signal burst of unique duration.

1.1 Tracer System

A recent addition to the facilities afforded by time-multiplexed chains is the Tracer system,[5,6] a radio-navigation relay system designed for tracking a wide range of remote and unattended objects (Fig. 1). The signals transmitted by the stations are received at the mobile unit where a Tracer converter changes their frequency from the original radio frequency to an audio frequency—usually 2 kHz—whilst maintaining their relative phase differences. After being amplitude-limited, the audio signal is relayed via radio or land-line to a receiving centre where the position information is extracted by a Tracer decoder.

This retransmission arrangement takes advantage of one of the principal benefits enjoyed by time-multiplexed systems: phase shifts in the various stages of the receiving equipment do not affect the phase differences between the bursts of signal, provided that they are substantially constant throughout each sequence, so the accuracy of the position measurement is preserved. This is true whether the receiver is of the conventional type or, as in Tracer equipment, is divided into two units at widely-separated locations.

2 Decca Navigator System

The signals of the various transmitters of the Decca Navigator system[4] are separated and identified by being frequency multiplexed. The stations forming a chain transmit simultaneously on harmonically-related frequencies which are multiples of a frequency f (approximately 14 kHz) which identifies the chain. Most chains consist of four stations: a master station which transmits at $6f$ and the red, green and purple slave stations, transmitting at $8f$, $9f$ and $5f$, respectively.

A mobile receiver makes phase measurements to establish its position by converting the signals received to

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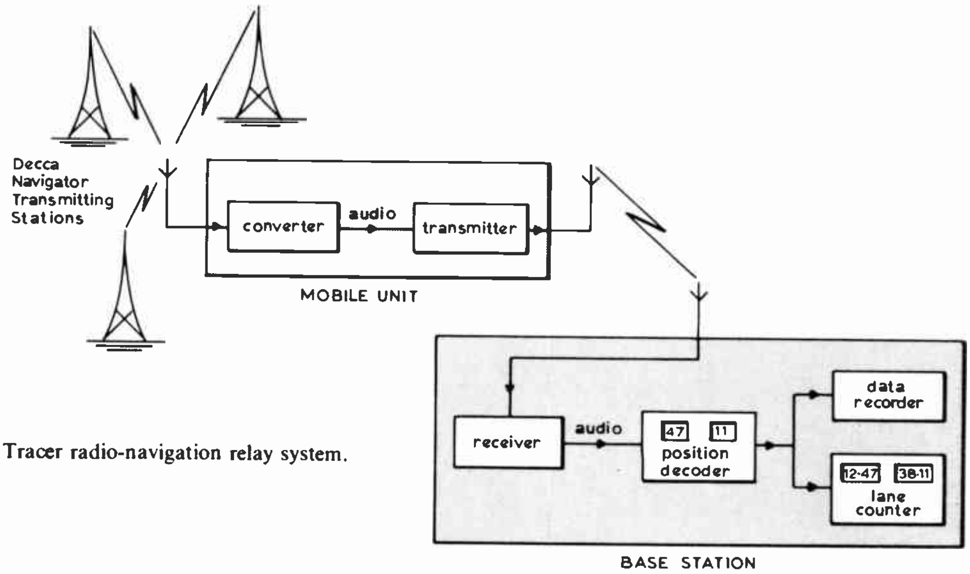

Fig. 1. Tracer radio-navigation relay system.

common comparison frequencies. So, for example, the position of the receiver on the 'red' pattern, generated by the master and the red slave station, may be measured by multiplying the $6f$ master signal and the $8f$ red slave to their lowest common multiple frequency, $24f$. The green- and purple-pattern phase comparisons are made at $18f$ and $30f$, respectively.

Such a receiver requires a separate channel for each frequency. Unfortunately, any drift of the phase delays of these channels leads to changes in the phase difference readings. These errors may be substantially reduced by a 'referencing' procedure in which signals at the various radio frequencies are input to the receiver simultaneously and the individual channels adjusted until their outputs show zero phase difference.

However, if one wishes to use the Tracer technique to retransmit the signals of a frequency-multiplexed radio-navigation system such as Decca Navigator, serious problems arise. Each of the four frequencies can be separately converted to a different audio frequency and amplitude limited before being relayed to the decoder.[7] But the phase delays of the various receiver and communications channels are not, in general, the same, so errors are introduced. It would be possible, in principle, to reference the system, but a suitable multi-frequency signal source would have to be provided at the mobile and synchronized to the decoder. This technique fails to meet one of the fundamental design criteria of the Tracer system: that the mobile unit be as simple as possible and as economical in cost and power consumption. The work described in this paper was undertaken to find a more acceptable way to apply the Tracer retransmission technique to Decca Navigator signals.

2.1 Use of the Lane Identification Transmissions

Section 2 describes only the basic Decca Navigator transmission format. Each phase-difference measurement defines a hyperbolic locus within a 'lane' bounded by lines of zero phase difference. This 'lane fraction' measurement must be supplemented by a 'lane number', identified by generating wider lanes at lower effective measurement frequencies. To do this, the transmissions are regrouped, each station in turn transmitting a 'multipulse' burst of all the frequencies in a sequence lasting 20 s. The multipulse transmissions also include an additional frequency, $8{\cdot}2f$; when this is differenced with the $8f$ transmission from each station, a $0{\cdot}2f$ pattern is generated, having very wide lanes (Fig. 2).

These additional features make retransmission of the Decca signals much more straightforward. Each station of the chain now transmits on each of the five frequencies

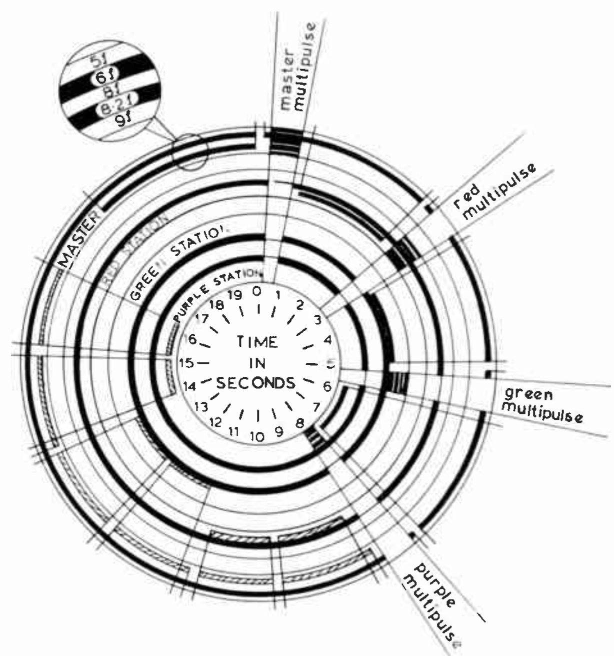

Fig. 2. 20 s transmission sequence of Decca Navigator multi-pulse chain. Hatched areas are supervisory communications.

at least once within the 20 s sequence. It is possible, therefore, to treat the sequence of signals radiated on any single frequency as a time-multiplexed radio-navigation system and to retransmit it using the Tracer technique. Only a single receiving channel is required, instead of the multiple channels of a conventional Decca receiver. This gives a simpler, cheaper receiver of lower power consumption but without lane identification.

A particular advantage of choosing $6f$ or $8{\cdot}2f$ is that the decoder can be designed to synchronize to the incoming signal sequence relatively easily by recognizing the breaks and signal bursts of the transmission sequence. Both $8{\cdot}2f$, the so-called 'orange' frequency (approximately 116 kHz), and the $6f$ master frequency (approximately 85 kHz), are being used successfully for single-frequency Tracer retransmission.

3 Tracer Equipment

3.1 Converter

The converter (Fig. 3) is simple in concept and similar in many respects to that used to receive Hifix or Seafix signals.[5] It filters and amplifies the radio-frequency signals, converts them to the audio frequency of 2 kHz whilst maintaining their phase differences, and substantially eliminates amplitude variations. The phase relationships between the audio signal bursts at the converter output are the same as those at its input provided the frequency of the local oscillator is constant and that the converter introduces negligible phase changes for input signal level changes.

Signals from the Decca stations are received at the mobile on a 1·5-m-long whip aerial. The base matching unit of the aerial, which feeds a 50 Ω cable to the converter, contains a broad-tuned circuit covering either all $6f$ or all $8{\cdot}2f$ channels.

The r.f. amplifier uses a dual junction-f.e.t. long-tailed pair to feed the dual-f.e.t. single-balanced mixer.

The local oscillator must have excellent short-term frequency stability. In particular, phase variations occurring within one sequence of the navigation transmissions must be negligible. Constant and long-term frequency errors, however, are corrected by the decoder (see Sect. 3.2). This allows a simple c.m.o.s. crystal-oscillator/divider with no oven, and hence low power consumption, to be used. An alternative local oscillator employing a programmable frequency synthesizer is fitted where several Decca chains are to be received.

The channel-defining filters of the converter are distributed between the radio- and audio-frequency sections. The r.f. filter uses a single X-cut crystal bar in a half-lattice circuit. The 2 kHz filter is designed from a 3rd-order maximally-flat (or Thomson) prototype and uses three 20 mm pot-cores. The overall −3 dB bandwidth of the converter is ±12 Hz and the time-domain response allows amplitude settling to −100 dB within 100 ms, a time short compared with the 450 ms duration of the multipulse bursts.

The stopband rejection required can usually be estimated by calculating the signal levels of adjacent Decca Navigator chains; the closest interfering frequencies possible are ±75 Hz in the $6f$ band and ±102·5 Hz at $8{\cdot}2f$. The filters give 38 dB attenuation in the worst $6f$ case and 49 dB at $8{\cdot}2f$; if this is insufficient, however, an alternative r.f. filter unit, employing two X-cut crystals in a half-lattice configuration, is available to increase the attenuation to 62 dB and 69 dB, respectively. The r.f. filters also attenuate the image frequency, 4 kHz above the signal frequency, by at least 60 dB.

A two-stage limiter is used, followed by a bandpass filter centred at 2 kHz which restores the fundamental component of the nominally square-wave limiter output. The converter's output amplitude range is less than 3 dB, and its phase delay variation less than 0·02 cycles, for an input signal range in excess of 90 dB.

The version of the converter fitted with a c.m.o.s. local oscillator has a power consumption of only 240 mW at 12 V. When fed from a 1·5 m whip aerial, its input field strength range is from 22 μV/m, corresponding to a range of approximately 500 km over poor soil, to 700 mV/m, found within a few hundred metres of a transmitting station.

3.2 Decoder

At the base station, a Tracer decoder is employed to extract the position information. The decoder is a digital phase-comparator designed specifically to measure the phase differences between the time-separated master and slave signals. Each master–slave phase difference is evaluated in percentage form to provide a direct and

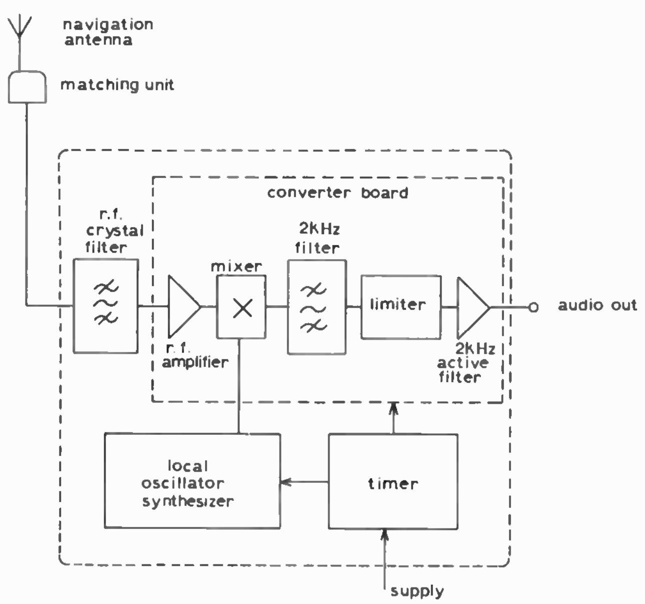

Fig. 3. Block diagram of Tracer converter.

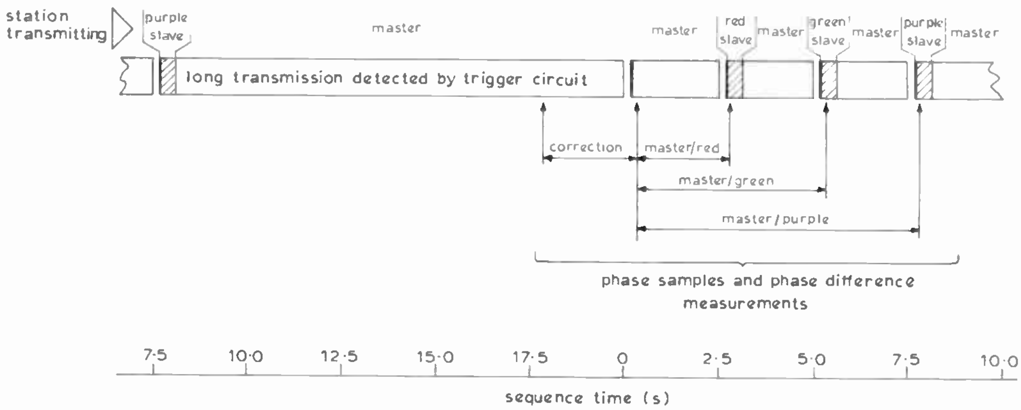

Fig. 4. Timing of decoder phase samples: $6f$ version.

immediate reading of the position-in-lane on the red, green and purple patterns. Two patterns may be selected for display on the front panel and to provide binary-coded decimal (b.c.d.) output data to feed a digital printer, cassette recorder or tape punch, or to drive a lane-counter (see Sect. 4.1).

3.2.1 *Phase-difference measurements*

The master and slave signals of the Decca Navigator $6f$ and $8{\cdot}2f$ sequences are not received simultaneously, so the Tracer decoder employs an unconventional phase-comparison technique which uses a free-running reference oscillator at a frequency nominally equal to the input signal frequency. The phase difference between the master and reference is measured and then that between slave and reference subtracted from it, to give the required master–slave reading.[6]

This measurement procedure is carried out digitally. A phase comparator generates 2 kHz pulses, of duration proportional to the phase difference between the input and reference signals, which are used to gate 2 MHz clock pulses into three cascaded up/down decade counters giving phase-difference counts in the range 0 to 999. The phase difference subtraction is achieved simply by counting up the master–slave phase difference and counting down the slave–reference difference at appropriate times in the sequence (see Fig. 4). The residual count in the two most-significant decade counters gives the position-in-lane in hundredths of a lane, or centilanes (cl).

Generally, the input signal frequency is not exactly 2 kHz, principally because of the frequency error of the converter local oscillator. This results in a phase drift of all received transmissions relative to the reference. To restrict the unwanted phase shift to, say, a negligible 0·01 cycles over the approximately 10 s measurement period would require an unreasonably-high converter local oscillator accuracy of better than 0·01 parts per million. Instead, a frequency correction process is used in which the phase drift rate is measured and corrections calculated for the resulting errors. A frequency error of up to 100 parts in 10^6 can be tolerated.

To measure the phase drift rate, the phase-difference measuring procedure described above is applied to the signal from a single station. Figure 4 shows the two 'correction' phase samples of the master signal, 2·5 s apart. All master–slave phase readings are separated by multiples of 2·5 s, so simple proportional corrections may be applied by feeding extra clock pulses to increment or decrement the phase-difference counters.

This simple decoder is unsuitable for tracking high-speed mobiles. A recent paper[6] shows that the effective instant of measurement of each line of position (l.o.p.) is that at which the appropriate slave phase reading is taken. So the red l.o.p. is measured 2·5 s before the green and 5·0 s before the purple, and a skew error, proportional to velocity, results.

4 Lanes and Lane Identification

The hyperbolic grids generated by each master–slave station pair at $6f$ or $8{\cdot}2f$ are confocal with those shown on the conventional Decca Navigator charts. The width of each lane where it crosses the baseline joining these two stations is half a wavelength at the frequency used for phase comparison. Table 1 shows the principal relationships between the various systems.

The Decca Navigator numbering convention groups 24 red, 18 green or 30 purple lanes into zones of effective comparison frequency f. Zones are labelled A–J. The $0{\cdot}2f$ zone-identification patterns are used to identify the correct zone from a group of five, A–E or F–J. Each 5-zone group corresponds exactly to 30 lanes of the $6f$ pattern or 41 lanes of the $8{\cdot}2f$ one.

Table 1 shows that Tracer lanewidths are the same for all three patterns; Decca lanes are not. To convert from Decca to Tracer nomenclature, one reduces the Decca coordinate to a simple lane number and fraction, then multiplies by the appropriate ratio. Decca-to-$6f$ Tracer conversion is especially simple since each $6f$ lane corresponds exactly to an integral number of Decca

Table 1

Relationships between Decca Navigator lanes and zones and Tracer $6f$ and $8{\cdot}2f$ lanes

		Pattern		
		Red	Green	Purple
Frequency of phase comparison	Decca receiver	$24f$	$18f$	$30f$
	$6f$ Tracer	$6f$	$6f$	$6f$
	$8{\cdot}2f$ Tracer	$8{\cdot}2f$	$8{\cdot}2f$	$8{\cdot}2f$
Baseline lane-width (m)	Decca lane	441	588	353
	Decca zone	10588	10588	10588
	$6f$ Tracer lane	1765	1765	1765
	$8{\cdot}2f$ Tracer lane	1291	1291	1291
Decca lanes per	$6f$ Tracer lane	4	3	5
	$8{\cdot}2f$ Tracer lane	2·926	2·195	3·658

lanes. The Tracer grids may be printed on to transparent overlay sheets or directly on to charts, in the usual way.

The Tracer systems, in common with other single-frequency, time-multiplexed systems, have no lane-identification facilities, so a lane-counter unit has been developed which is connected to the output of the decoder. The lane numbers of the starting location are entered manually and the coordinates are adjusted automatically each 20 s; it is assumed that each new position lies within ± 50 cl of its predecessor.

5 Errors

The errors in Decca Navigator-measured positions have been studied extensively over many years and data sheets and error contours published. This Section identifies the situations in which Tracer errors differ significantly from those given by a conventional Decca Navigator receiver.

5.1 Fixed Errors

Fixed errors are chiefly due to inaccurate assumptions of the velocities of propagation along the transmission paths between the Decca stations and the mobile receiver. These errors have been surveyed in many areas. The results may be used to correct Tracer position fixes, but small discrepancies will remain because while a Decca Navigator receiver compares signals propagating at the master frequency over one path and at the slave frequency over the other, a tracer receiver uses either $6f$ or $8{\cdot}2f$ signals for both.

5.2 Skywave Errors

Skywave errors are due to unwanted ionospherically-reflected components of the signal interfering with the wanted direct or groundwave component. They are the principal source of position-line and fix errors in the Decca Navigator system.[8] Their amplitudes vary both diurnally and seasonally (being greatest during winter nights) and with range and effective ground conductivity. Because of the random nature of skywave reflections, these errors are only predictable as being within certain statistical limits.

Skywave errors depend solely on the ratio of the amplitudes of the reflected and groundwave components and on their phase differences; they are independent of the transmitter powers and the receiver bandwidths. Standard data and conventional techniques for skywave error prediction may, therefore, be used to estimate the accuracy of a Tracer system, although it is again necessary to make small allowances for the differences between the frequencies used on the various paths in the two systems.

The components of error due to the effect of skywaves on the 'locking' paths between the master and slave stations, together with system phase-locking errors due to other causes, are, of course, the same for Tracer and Decca Navigator receivers.

5.3 Random Noise

Noise errors depend on the radiated powers of the transmitters, the locations of the receivers and the time and season. These factors are the same for the two types of receiver. The noise bandwidth of the Tracer system is normally equal to the converter bandwidth, ± 12 Hz. This is similar to the receiving channel bandwidths of conventional Decca Navigator receivers, but their phase-locked loops give an additional reduction of noise. Tracer digital output data may also be smoothed, but the allowable acceleration of the mobile is then limited by the 20 s interval between position samples.

In general, the performance of a Tracer system is at its worst in high noise environments, such as tropical locations, and when tracking rapidly-moving objects. In almost all cases, however, the limitations imposed by skywave errors exceed those of ambient noise and the wider noise bandwidth of the Tracer receiver system is of little importance.

6 Applications

6.1 Pattern Monitoring

A stationary Tracer converter may be used to monitor pattern stability; errors in the apparent position due to phase shifts in the generation or propagation of the signals may be used to correct the measured positions of nearby mobiles. A recent paper[9] gives details of this technique and estimates the extent to which it may be used to cancel skywave errors.

Table 2 compares skywave errors monitored over a period of 60 h at five sites lying within a circle of radius 10 km with predictions calculated from Decca Navigator planning charts. The distribution of random errors in Decca Navigator position fixes is usually Gaussian and the results are expressed in conventional form as the radius of the circle, centred on the mean measured position, within which 68% ($1\ \sigma$) of all recorded points lie.

Table 2

Skywave errors: five fixed sites

Period	68% fix repeatability (m)	
	Measured	Predicted
Winter night	103	329
Summer night	83	237
Dawn/dusk	46	183

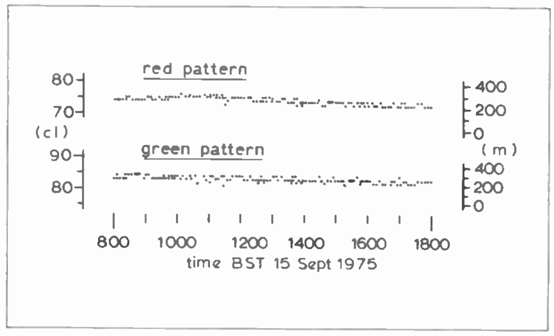

Fig. 6. Lightvessel daytime pattern variations: standard deviation is approx. 30 m.

6.2 Station-keeping

Tracer equipment is employed to monitor the positions of unattended seamarks such as automatic lightvessels and large navigation buoys.[10] Decca Navigator transmissions are especially suitable for this purpose because of their excellent accuracy and reliability. The wide lanes give adequate resolution with minimal lane ambiguity, the maximum excursions of most moored objects being significantly less than one lane.

A detailed study of 672 h of operation of a test installation has been used to check the resolution, accuracy and skywave sensitivity of the system. A Tracer converter was installed on the Galloper Lightvessel in the English Channel to receive the $8{\cdot}2f$ signals of the English chain (5B) and relay them via a 53 km-long u.h.f. link to decoding equipment at the North Foreland Lighthouse. Each 5 minutes a timer activated the converter and transmitter for 32 s, a sufficient time to guarantee reception of at least one set of $8{\cdot}2f$ transmissions.

Figure 5 shows the red and green patterns in the vicinity of the Lightvessel; its nominal position is marked AP and mean measured position, MM. The discrepancy is 1 cl (27 m) on the red pattern and 2 cl (68 m) on the green.

The measured 24 h tracks, of which Fig. 5 is an example, are dominated by tidal and wind forces which swing the vessel about its mooring. Under certain conditions, however, it remains almost stationary and the fix stability can be estimated. Figure 6 shows such a 10 h period during which 99% of all data lay within ± 2 cl, a fix standard deviation of 30 m. The predicted value is in the range 25–50 m.

A similar technique was used to estimate the 'summer night' skywave errors (see Table 3); 99·5% of the 9050 readings recorded lay within limits corresponding to the sum of the anchor chain length and the predicted 99% fix repeatability.

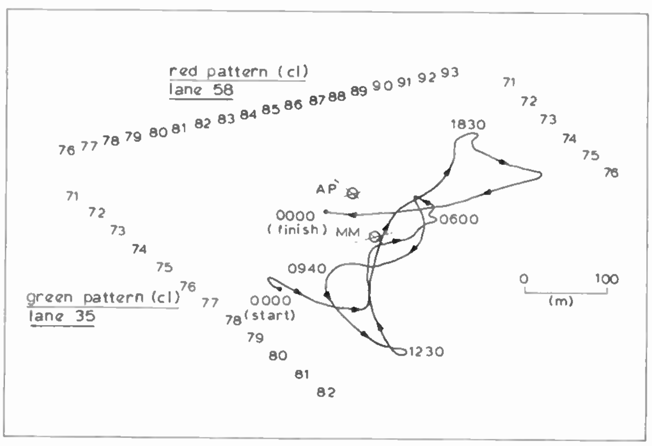

Fig. 5. Track of lightvessel for 24 h; the grid is in units of one hundredth of a lane (cl). AP = assigned position. MM = mean measured position.

Table 3

Summer night skywave errors: Galloper Lightvessel

Pattern	68% fix repeatability (m)	
	Measured	Predicted
Red	26	31
Green	57	47

6.3 Buoy and Ship Tracking

Tracer-equipped free-floating buoys are used to plot marine currents. To minimize power consumption sampled operation is normal, the interval between position fixes being set so that the mobile cannot cross more than 50 cl of the narrowest pattern when travelling at its maximum velocity. If every 20 s fix is used, the limiting velocities are 32 m/s (72 miles/h) when receiving $8{\cdot}2f$ signals and 44 m/s (98 miles/h) for $6f$.

Multiplexed groups of buoys are frequently deployed, time-sharing a common radio channel and decoder. Individual buoys are switched by internal crystal-controlled clocks. A typical buoy track, lasting 15 tidal cycles, is shown in Fig. 7.

6.4 Animal Tracking

An unusual application of the system is to track animals on open pasture land for experimental purposes.[12] Simplified, light-weight Tracer equipment is used.