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RADIO TECHNIQUES AND SPACE RESEARCH

British participation in space research is still in the embryo stage, and it may be argued that it is too early to arrange a major technical conference on this subject. Nevertheless this subject was deliberately chosen by the Council with the dual purpose of providing a forum for discussing the significant work which is already being done and as a vehicle for presenting to the nation's radio and electronic engineers the challenge which lies ahead.

The programme of the Convention has been carefully compiled with these two objectives in mind. A proportion of the time has been allocated to physicists from Universities and Government Establishments, who will describe the objectives of their various studies and in doing so will point to those areas where they seek the support of their radio and electronic engineering colleagues. Other papers from industry will describe satellite and rocket instruments under development or already in use in various joint enterprises with the Universities.

The constraints on the scientist or engineer involved in a satellite programme are formidable. There are the difficulties of establishing his instrument capsule in space, followed by the problems of feeding it with power, maintaining communications and carrying out such actions as attitude stabilization, sun-seeking, etc. Dominating all this is the problem of ensuring the instrumentation survives the launch phase and achieves a high standard of reliability in the space environment. All these aspects are being covered in the Convention and papers will be presented on the capabilities of various launchers, the engineering problem of the satellites themselves, the power sources which can be used and techniques for achieving reliability.

One of the most dramatic ways in which radio has been applied to space research is, of course, in Radio Astronomy. This is an area where Britain and the Commonwealth enjoy an acknowledged lead and there will be contributions by workers in this field. British and Commonwealth scientists have also played a leading role both in tracking the artificial satellites launched by the U.S.A. and the U.S.S.R., and also in extracting important scientific conclusions from the data obtained.

Initially satellites were used primarily as tools for pure research, but more recently they have been used in more operational roles. The most exciting of these is the use of satellites as relay stations to provide global communications. Another field of great potential importance is the use of satellites for navigation purposes, and already there are pilot experiments in being which are demonstrating the scope of this technique. These are some more of the topics which will be covered by the Convention.

Clearly no major technical conference on any aspect of space research would be complete without contributions from the two nations already so well advanced in the subject. The Convention Committee has been fortunate in obtaining several important contributions from the U.S.A. and it is hoped that there will also be papers from the U.S.S.R.

The Convention will be held in Christ Church, Oxford, and arrangements are being made for informal or subsidiary meetings in addition to the formal presentation of papers. There will be an exhibition of equipment and material appropriate to the subject of the Convention and films dealing with space technology will also be shown.

I. M.

EXTRAORDINARY GENERAL MEETING

NOTICE IS HEREBY GIVEN that an Extraordinary General Meeting of Corporate Members of the Institution will be held at 6 p.m. on Wednesday, 7th June, 1961, at the London School of Hygiene and Tropical Medicine, Keppel Street, Gower Street, London, W.C.1, to transact the following business:—

1. In pursuance of Article 34 of the Institution's Articles of Association the President and Vice-Presidents having tendered their resignation, the Council, by virtue of Article 35, announce that the vacancies arising in the office of President and Vice-Presidents have been filled as follows:—

The President:—

Admiral of the Fleet the Earl Mountbatten of Burma, K.G., P.C., G.C.B., G.C.S.I., G.C.I.E., G.C.V.O., D.S.O., LL.D., D.C.L., D.Sc. (*Member*).

Vice-Presidents:—

L. H. Bedford, C.B.E., M.A., B.Sc., F.C.G.I. (*Member*)
W. E. Miller, M.A. (*Member*)
Professor E. E. Zepler, Ph.D. (*Member*)
J. L. Thompson (*Member*)
Colonel G. W. Raby, C.B.E. (*Member*)
Professor E. Williams, Ph.D., B.Eng. (*Member*)
Air Vice-Marshal C. P. Brown, C.B., C.B.E., D.F.C. (*Member*)

2. **To consider and, if thought fit, pass the following resolution with or without amendment as a SPECIAL RESOLUTION.**

SPECIAL RESOLUTION

"That the provisions of the Articles of Association of the Institution be amended by the deletion of the existing Article 16 and substituting a New Article 16 as set out below.

16.

SUBSCRIPTIONS

The corporate and non-corporate members of the Institution shall pay an annual subscription, irrespective of the date of election, the amount of which shall be determined from time to time by the Council but shall not exceed:—

Members	£20	Graduates over 35 years ...	£16
Associate Members	£16	Graduates 25-35 years ...	£12
Companions	£20	Graduates under 25 years ...	£10
Associates	£16	Students	£10

A Member, Associate Member or Companion may compound for his annual subscription by the payment to the Institution in one sum of an amount which will, from time to time, be determined by the Council. The amount shall be at least ninety pounds and shall not exceed a sum equal to fifteen times the annual subscription payable by a Member."

Corporate Members who are unable to attend this meeting are reminded that they may appoint another Corporate Member as their proxy to attend and vote on their behalf. Such forms of proxy must be deposited at the Offices of the Institution 48 hours before the time of the holding of the meeting.

By Order of the Council.

GRAHAM D. CLIFFORD,
General Secretary

15th May, 1961
9 Bedford Square,
London, W.C.1

A copy of this notice has already been sent to all corporate members together with a form of proxy.

Radio and Television Broadcasting in Great Britain†

(This survey has been compiled by the Technical and Television Group Committees of the Institution as a contribution to the deliberations of the Committee on Broadcasting.)

Contents

Introduction

In August 1960 the Government appointed a special committee to advise on the future of broadcasting. This body invited recommendations and proposals from all organizations concerned with the future of sound and the vision services in Great Britain: the Institution was one of the bodies invited to give evidence. The Council of the Institution has decided that the best service that it can render to the Committee on Broadcasting is firstly to review the conditions and circumstances which led to the present broadcasting systems and secondly to examine possible developments.

1. Sound Broadcasting

The B.B.C. now transmits a Home Service which is regional in character, a national Light Programme, and a national Third Programme.

Until 1940 sound broadcasting throughout the world was mainly confined to the low, medium and high frequency bands using amplitude modulation. A new system of broadcasting employing frequency modulation was introduced in America in the early 1940's after a period of experimental transmissions. The chief advantage of this system of transmission is the relatively good reception available in areas where medium frequency reception is very poor because of interference and severe fading.

In 1955, after field tests, the first British v.h.f. sound broadcasting station was established at Wrotham. The present twenty v.h.f. stations duplicate the service at present transmitted on the medium and low frequencies and may ultimately replace it entirely.

Reception difficulties of signals due to overcrowding and other forms of interference make it impracticable to expand broadcasting services in the medium frequency band. Thus any expansion is likely to be in the v.h.f. bands.

In order to minimize interference both to public broadcasting as well as other essential radio services, it is imperative that international agreements on allocation of frequencies are strictly observed by all transmitting authorities.

1.1. *Future Sound Programmes*

Band II, part of which is already in use, gives adequate and suitable provision for an expansion of national or local sound broadcasting. Such expansion might include cultural or educational programmes and stereophonic broadcasting services. Any eventual stereophonic system of broadcasting must be compatible with monaural systems existing at the time of its introduction.

With regard to local broadcasting there is a great deal to be said for the provision of alternative programmes of a local character under appropriate control.[1] This development was recommended by the Beveridge committee in 1951 but has not yet been implemented.[2] A recent B.B.C. application to introduce local broadcasting has been rejected by the Postmaster-General pending the report of the Committee on Broadcasting.[3]

2. Television Broadcasting

The first experimental television transmissions were made by the B.B.C. in 1932 using a 30-line system. This was soon abandoned in favour of high definition systems and in November 1936 came the official inauguration in Great Britain of the world's first regular television service, using a 405-line standard.

Other countries, following the British lead, developed television systems. Whilst these systems are basically similar to the British system, certain characteristics are different, and these are compared in Appendix 2. In 1940 and after extensive study a television service was opened in the U.S.A.[4] on a standard of 525 lines 60 fields per second, a standard which was, after the war, adopted by Canada, Japan, and certain other

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† Approved by the Council for publication on 2nd May 1961 (Report No. 21).

countries. In France the television service was started on 441 lines 50 fields per second but was later changed to a higher line standard of 819 lines 50 fields per second.†

On the recommendation of the Hankey Committee, the television service in Great Britain was resumed in 1946 on the pre-war standard of 405 lines. The Independent Television Authority which commenced operations in 1955 uses the standard British 405-line system.

In most other countries of Europe a 625-line system has been adopted although some of these countries employ standards which differ in certain characteristics. Australia, New Zealand, and some countries on the African continent also use a 625-line system similar to that common in Western Europe.

3. The Availability of Frequencies for New Television Services

Television transmission requires very wide bandwidth, and even with careful geographical siting of stations using the same channel so as to minimize mutual interference, the number of transmissions in each band is nevertheless limited.

At the present time Bands I and III are used for broadcasting the existing 405-line programmes. The B.B.C. transmits on Band I and I.T.A. on Band III. Both organizations give a national coverage but there is room in Band III for a further network of transmitters to give a third programme. This is only possible, however, for a coverage of approximately 90% if the standard is maintained at 405 lines.

For a third programme using 625 lines the coverage would be about 70%. If a complete change to 625 lines is made then Bands I and III will have to be re-engineered and because of the greater bandwidth demands of the 625-line system, only two nationwide programmes will be possible in Bands I and III.

Any major expansion of national broadcast programmes over the air will necessitate the use of u.h.f. Bands IV and V on any system. As stated in the Television Advisory Committee Report 1960 (Appendix 1),[5] there has been some international agreement on the use of an 8 Mc/s channel in Bands IV and V which allow the setting up of three or four different national networks. The u.h.f. band as allocated to television will permit four national programmes but there has been some encroachment of other services which will at the moment limit the frequencies available. Assuming Bands I and III to be re-engineered on 625 lines thus providing a further two programmes, there will then be the possibility of a total of five or six different programmes (see Appendix 3).

Unfortunately, by comparison with Bands I and III, the u.h.f. bands suffer from two major disadvantages: (1) the reduced service area for a given power; (2) propagation is less reliable, shadow effects of tall buildings, hills, etc., being more marked.

A great deal of research into u.h.f. problems is at present being carried out. Due to overcrowding of the v.h.f. bands on the Continent of Europe some countries, notably Germany and Italy, have been forced into the u.h.f. bands to provide a second television service. In those countries considerable progress has been made in the development of u.h.f. receivers using 625 lines and f.m. sound.

The problems of transmission at ultra high frequencies have been dealt with at length in the report of the Television Allocations Study Organization[6] of the F.C.C. of America, and the B.B.C. experiments serve to confirm some of the difficulties experienced.

It is for these and economic reasons the Institution considers that the u.h.f. bands should not be used in Great Britain for a national television service except after further detailed study. The u.h.f. band may immediately be utilized for television transmissions of a local character.

4. Wired Systems as an Alternative to U.H.F. Transmissions

In the circumstances described in Section 3 the possibility of a nationally *wired* (i.e. non-radiating) television system becomes attractive. Such a system would generally provide superior signals with several different programme channels. Both 405 and 625-line signals could be transmitted over the same network if such a requirement exists during the transition period. In rural areas, or areas of low population density, the network could be supplemented by low power u.h.f. stations.

Wired television networks are already being installed or operated in many towns. Several different methods are used, some of which are not capable of carrying an adequate number of programmes. Others need special receiver arrangements. There is, therefore, a need for a gradual change to agreed national standards for wired television. The requirements of such standards should be investigated at an early date.[7, 8, 9]

Developments in the field of waveguide transmission indicate that within a decade it should be possible to transmit signals at frequency of 35 kMc/s with a bandwidth of 1000 Mc/s over long distances by this technique. For the first time it is possible to foresee a national distribution system of sound and television programmes on a wired basis. The main waveguide

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† It is understood that it is proposed that the French second service will be established on a 625-line standard but with different parameters from West European standards.

trunking system will serve distribution centres from which subscriber lines could be laid to individual homes.

The achievement of such a distribution system would offer substantial benefits to the public, such as:

(1) Virtually no fringe areas.
(2) No interference.
(3) A selection of programmes at a constant signal strength.
(4) The possibility of significantly improved pictures of increased definition.
(5) Planning without the necessity for international agreement.

The value of a wired system of the type envisaged is independent of the radio spectrum and of the regulations governing the use of the various radio frequencies. It would appear from existing knowledge that this method of distribution would be competitive with the cost of installing transmitters for broadcasting an equivalent number of programmes.

The prospects are sufficiently near to achievement to make it highly desirable to consider these implications in any long term planning.

5. Television Standards

In terms of picture quality the main deficiency of the 405-line standard is in respect of line visibility and moiré effect. There would be a worthwhile improvement by an increase in the number of lines but the practical limit to the number of lines is set by the bandwidth available. With an 8 Mc/s bandwidth, upon which there has been some international agreement 625 lines is a good engineering choice (see Appendix 1).

Although the B.B.C. tests[10] show that there is no significant improvement in overall picture quality between the 405 and 625-line pictures, the 625-line standard holds the promise of increased picture resolution with further development of transmitting and receiving equipment.

With the introduction of additional programmes and colour transmissions, new transmitting and receiving equipment would be required. It would therefore be opportune at this stage to change the standard and come into line with one of the European 625-line standards. The predominant advantage in the change will be the reduction in observable line structure especially when television pictures are viewed on a large screen—21-in. or more. Furthermore the change to 625 lines will render interchange of programmes technically easier to achieve since the vast majority of television systems in Europe and the Commonwealth are already operating on 625 lines.

A change to a 625-line Western European standard would not give a marked improvement in horizontal resolution.

Some advantages might accrue by the adoption of an 8 Mc/s channel for a 625-line standard. A standard of this type has been proposed in Appendix 3 of the 1960 T.A.C. Report.[5] This proposed standard does not make recommendations as to the modulation polarity of the vision signal or the type of modulation for the sound signal. These aspects of the standard should be the subject of further investigations.

The major television-producing country in the world, however, the U.S.A., uses a system of 525 lines 60 fields per second, i.e. different from either Europe or Great Britain, and a further difference is that the number of complete pictures per second is thirty (60 fields per second) and not twenty-five (50 fields per second) as in the case of Europe. The choice of picture frequency is strongly influenced by the mains supply system in use in the country in which the television system is being installed, and this accounts for the difference between the American system and those adopted in Europe (see Appendix 2).

If a change were made to a 625-line system in this country, very little technical difficulty would be experienced in broadcasting and receiving, on the same system, recordings which have originated in the countries using the 525-line standard. There is a good argument for permitting broadcasting authorities in this country to transmit on either 625 lines 25 pictures per second or 525 lines 30 pictures per second. This would necessitate precautions being taken in the design of television receivers to accommodate the two standards and is likely to increase the receiver cost by about £2. It is essential, however, to carry out further field tests to establish the practicability of the proposal.

6. Changeover of Television Standards

If it is decided that the present television standards should be changed from 405 to 625 lines then the introduction of the new system must be effected in such a way that it gives no break or reduction in the quality of service to viewers and as little disruption as possible to the British radio industry.

As far as the studios and transmitters are concerned multi-standard equipment is already being manufactured and is in use. In recent years British broadcasters have installed studio equipment capable of accommodating 625 lines. It would, however, be a major practical and economic problem to change existing equipment such as transmitters and aerials.

The method of changeover will greatly affect manufacturing policy and, of course, viewers. Whatever method is adopted the existing 405-line transmissions in Band I and III must remain unchanged for 5–10 years after an alternative service is provided to allow for receiver obsolescence. Three possible methods of making the change are:

1(*a*) Duplicate the existing two programmes, i.e. B.B.C. and I.T.A. on 625-lines in Band IV.

(*b*) Commence any new programme *only* on 625-lines Band IV. Existing 405-line receivers continue to receive two programmes. New v.h.f./u.h.f. receivers receive three (or more) programmes. (*Note:* Dual-standard receivers will not be essential.)

(*c*) Band I and III transmissions on 405 lines cease after the new service has been established and a reasonable time has been allowed for receiver obsolescence. These bands will then be available for re-engineering them additional 625-line programmes.

2(*a*) Duplicate the existing two programmes on the 625-line system in Band IV.

(*b*) Commence the new programme on the 405-line system in Band III, and the same programme on the 625-line Band IV (i.e. all programmes could be received on the present 405-line receivers or on the new 625-line receivers).

(*c*) Cease 405-line transmissions as previously stated 1(*c*).

3(*a*) Install cable distribution in all big towns for say three 405-line and three 625-line duplicated programmes on Band I (or lower) frequencies. Simplified 625-line or existing 405-line receivers can be used.

(*b*) To cover rural areas fill in with Band IV 625-line transmitters giving all programmes.

(*c*) Cease 405-line transmissions as previously stated 1(*c*) if desired.

It will be apparent that methods 2 and 3 with complete duplication of all programmes on 405 and 625 lines would produce the minimum of disruption in service or obsolescence in receivers. This would be at the cost of the additional transmission facilities required to duplicate both the existing programmes and any new programme. On the other hand, neither of these methods require dual standard receivers so that there would be a considerable saving in the cost to the public of new receivers.

7. Economics of Change

Three programmes broadcast at u.h.f. would probably need between 100 and 200 transmitters sites to provide programmes to cover the country. The building of these stations will take at least five to seven years even assuming an early acquisition of the sites and will cost a very substantial sum (at least £50M). The acquisition of between 100 and 200 suitable sites is likely to take a very long time judging by past experience. Suggestions for alleviating this problem are given in Section 11.

A u.h.f. receiver for either 405 or 625 lines is estimated to cost between £5 and £10 more to the public than existing 405-line receivers. Dual standard u.h.f. receivers would on the same basis cost £10–£15 more than an existing 405-line receiver.

8. Colour Television

There is general agreement that colour television should be permitted as soon as a decision has been reached on the future television standards to be used for monochrome transmissions. It has been stated by the B.B.C. that colour transmissions could be started on the present standards using the N.T.S.C. system within twelve months of being given permission to go ahead.

If the decision is taken to leave the standard as at present (405 lines) this statement by the B.B.C. means that the United Kingdom could within 12 months be given a colour television service of a few hours a week. If a change to a 625-line system is made then colour television could only be introduced on those standards when the new 625-line transmissions are commenced and would follow the natural growth of the system.

It is likely, however, that even if colour television were introduced, the growth of colour television receiver sales would not, for reasons of receiver cost, be very great. The major difficulty is in the production of an inexpensive colour display tube. The only available colour tube is several times the cost of a monochrome tube and the receiver is likely to cost at least £200 initially for either of the standards discussed.

Standards conversion for colour television is extremely difficult and if programme exchange on 625 and 525-line standards is to be made then this is a strong argument for conforming with one or other or both of these standards.

9. Programmes and Educational Services

It is possible that the Committee on Broadcasting will recommend that there should be an extension of programmes available, both for sound and television. As a professional body interested in education, the Institution considers that broadcasting facilities should be made available for purely educational programmes. There is a place for more formal educational programmes on television on the lines of the present schools broadcasts, and more could be done in investigating its value as a direct teaching aid. There is also a need for more adult education by television. A serious programme for viewers with only a modest educational background would do much to improve the common culture.

9.1. *Programmes*

A strong body of opinion regrets the competitive nature of the two authorities responsible for radiating television service for public acceptance. This has resulted in two programmes which do not offer sufficient choice of subject matter. This is, in fact, a common experience where public viewing numbers are taken as a criterion of success. It is doubtful if expansion on these competitive lines is desirable since it will produce a dilution of talent that has the greatest public appeal with a consequent degradation of programme quality.

Any expansion of television should therefore make some attempt to overcome this natural convergence of programme types and provide the public with a choice of different programme material. The present restricted broadcasting hours represents under-exploitation of a considerable capital investment in studios, television networks, and receivers. It is, therefore, suggested that by removing the restriction on broadcasting hours this country could benefit from educational broadcasts during the hours not usefully employed for entertainment.

9.2. *Educational Television*

The use of television techniques for formal education has only received limited attention in the United Kingdom through the schools broadcasts of I.T.A. and the B.B.C. In other countries, however, notably the United States of America, there have been many years of experience in using broadcast and closed circuit television for formal education at all levels. This experience seems to indicate that television provides an opportunity to present educational material to greatly increased numbers of students and that students' achievement on formal courses of study organized on television is not significantly different from conventional training. In fact, under appropriate conditions the use of television as a teaching aid shows an impressive superiority over conventional methods. Television, as used in the United States, does not replace the teacher, but supplements his teaching. The success of these operations depends upon co-operative planning between television and classroom teachers.

It is important that channels should be made available for educational television and, additionally, that existing authorities should be permitted and encouraged to broadcast for longer hours to cater for educational needs. Such a service reinforced by extensive closed circuit systems will greatly improve our teaching methods and possibly raise our educational standards.

10. Subscription Television

It is recommended that subscription television should not be granted radiating (over-the-air) licences for national coverage owing to the lack of channel space.

Subscription television is most easily achieved over a wired system and could be introduced if desired as an adjunct to a system distributing all other programmes. Where national coverage is obtained by the use of u.h.f. transmitters for rural areas these facilities should be available to subscription television.

11. The Future Overall Control of Broadcasting

At the present time we have, in this country, two broadcasting organizations, namely the B.B.C. and the I.T.A., a national communications organization, the G.P.O., and several programme producing companies. The G.P.O., in conjunction with the Government Committee on Frequency Allocation, takes decisions on frequency allocation. Systems are decided either by a natural growth resulting in a *fait accompli* or by Government approval following recommendations from the Television Advisory Committee (T.A.C.) which in turn is advised by the Technical Sub-Committee of the Television Advisory Committee (T.S.C./T.A.C.).

The administration of frequency allocation is at present handled by the G.P.O., which is also the representative of the United Kingdom in international discussions. It is placed in an invidious position when it is the licensing authority and also an important user of the frequencies it allocates.

In the United States of America the Federal Communications Commission provides a single body with ultimate authority to take decisions on all matters relating to systems standards, frequency allocation, interference between services, etc. Such a body in the United Kingdom, if properly constituted, could solve many of the problems into which a series of committees are now enquiring or have enquired.

It is apparent therefore that a considerable amount of study is required on all aspects of broadcasting, particularly in television, and this work could be carried out through such a central authority and financed by the Government.

It is the opinion of this Institution that this country should have a single Government Committee called possibly the National Communications Authority (N.C.A.) to take decisions on all matters relating to system standards, frequency allocation, interference between services, service area, etc.

Moreover, it is recommended that there should be a sub-committee called possibly the National Broadcasting Systems Committee (N.B.S.C.) to advise the N.C.A. on technical matters.

It is further recommended in this context that this N.B.S.C. should be financed by the Government through the N.C.A. to undertake specific studies for the purpose of providing clear future planning of systems under such headings as:

(*a*) Standards for monochrome television

(*b*) Standards for colour television

(*c*) Standards for wire broadcasting.

It is suggested that the N.B.S.C. would allocate specific budgeted tasks to the most appropriate laboratories in the country.

Assuming that the National Communications Authority is appointed it will be responsible for the control of all transmitting services. This can take two forms, either on the lines of the past development, in which case a national authority has acquired sites, erected transmitters, etc., in an attempt to achieve national coverage. Alternatively, the same coverage could be achieved by allowing Universities, Local Councils, private concerns and the like, to apply to the N.C.A. for permission to erect and operate transmitters. In this way, the problem of site acquisition would be alleviated because only when a locality wanted the service would they apply and thus make every effort on a local basis to acquire the necessary site. This has proved highly satisfactory in America, where the service is determined by public demand.

12. References

1. "A Plan for Local Broadcasting in Britain", Pye Ltd. (Cambridge), Issue No. 2, 1960.
2. Report of the Broadcasting Committee 1949 (Cmd. 8116), 327 pp., 1951.
3. E. L. E. Pawley, "Local broadcasting—the engineering problem", *Ariel*, **6**, No. 1, p. 4, January 1961.
4. D. G. Fink (Ed.), "Television Standards and Practice" (McGraw-Hill, New York, 1943).
5. "Report of the Television Advisory Committee 1960' (H.M.S.O.).
6. "Engineering Aspects of Television Allocations." Report of the Television Allocation Study Organization to the Federal Communication Commission. 16th March 1959.
7. F. N. H. Robinson, "Wired Television". A Report prepared for the Radio and Television Retailers' Association, 1961.
8. H. E. M. Barlow, "A Report on Wire Broadcasting Systems". Prepared for the City of Leicester Corporation, November 1960. (Available from Rediffusion Ltd., London.)
9. K. A. Russell and F. Sanchez, "A common carrier multichannel television wire broadcasting system", *J.Brit.I.R.E.*, **20**, pp. 497–512, July 1960.
10. "Field Television Trials of 405-line and 625-line Systems in the U.H.F. and V.H.F. Bands 1957–8", British Broadcasting Corporation, May 1960.

THE RECOMMENDATIONS OF THE INSTITUTION'S COMMITTEES

Sound Broadcasting

1. Because of present overcrowding on the medium and long wave bands, it is recommended that there should not be any expansion of broadcasting services on these bands. Additional services should be accommodated in the v.h.f. Band II.

2. The Institution's committees support the proposals for the provision of additional and alternative sound programmes of a local character, but envisages the necessity for the control of such transmissions to come under the guidance of such authorities as indicated in paragraphs 11 and 12 of these recommendations.

3. The Institution's committees are unanimous in urging that in providing for additional sound services, provision should be made for a considerable extension of present educational and cultural programmes. For these to succeed in their application to all forms of education, the Institution advocates the setting up of an independent committee on educational broadcasting with representatives drawn from educational and cultural authorities.

Television Broadcasting

4. On the utilization of the u.h.f. Bands IV and V for television broadcasting, the Institution's committees recommend:

(*a*) These channels should not be used to provide a National television service except after considerable further study of the transmission problems involved.

(*b*) These bands are suitable for radiating television programmes of a local character.

(*c*) As an alternative to transmissions in the u.h.f. band, wired television systems should be considered.

Wired Television

5. Progress in the development of waveguide techniques makes it possible to foresee a National distribution system of both sound and television programmes, and the use of wired distribution would simplify any change in television standards. In this event there will be a need for a gradual change to agreed National standards for wired television.

Television Standards

6. For exclusively technical reasons, the Institution's committees agree that a 625-line, 8 Mc/s band-width standard is a good engineering choice. It is recommended, however, that certain aspects of the standard should be the subject of further investigation, e.g. vision modulation polarity, type of sound modulation and the interchange of 625- and 525-line systems.

The report describes three possible methods of changing standards, but it is recognized that the cost of changing standards of existing programmes would be very substantial. The minority view of the Institution's committees is that the high cost is not justified in view of the small advantage that would accrue to the viewer.

It is emphasized that should the British Government authorize an additional programme to the present two, such further services can be accommodated in Band III provided there is no change of television standards. More than one additional programme will require a move to Bands IV and V and/or the use of wired distribution as and when it would be opportune to change existing standards.

If a change of standards is made, then existing 405-line Band I and III transmissions should remain unchanged for 5-10 years after an alternative service is provided.

Colour Television

7. The Institution's committees recommend that colour television should be permitted as soon as a decision has been reached on future television standards. If the standards are changed then 625-line colour transmissions could commence as stations became available. If the standards are not changed, then national coverage of colour television could be achieved within 12 months. The high cost of colour receivers in either case will restrict rapid growth.

Educational Television

8. The Institution wishes to record its emphatic and strong recommendation that any extension of television services should first take into account the provision of facilities for supporting juvenile, youth and adult education. The Institution regards the services which could be given in this medium as being of paramount importance which should take precedence over providing further entertainment.

Programmes

9. Any expansion of entertainment television should make an attempt to overcome convergence of programme types and provide the public with a choice of different programme material.

Subscription Television

10. It is recommended that subscription television should not be granted radiating licence but should be confined to wired systems.

National Communications Authority (N.C.A.)

11. The Institution's committees recommend the setting up of a National Communications Authority (N.C.A.) to take over and replace the control and planning activities of the many existing authorities. Such an Authority would take decisions on all matters relating to system standards, frequency allocation, interference between services, service areas, wired broadcasting licences, etc.

National Broadcasting Systems Committee (N.B.S.C.)

12. As an adjunct to the above Authority, it is recommended that there should be a sub-committee, called possibly the National Broadcasting Systems Committee, to advise the N.C.A. on technical matters.

Appendix 1

Optimum Line Number for a Television System with an 8 Mc/s channel bandwidth on the basis of unity ratio of horizontal to vertical resolutions utilizing the vision to sound carrier spacing suggested by the T.A.C. in its 1960 report.

This is given by:

$$a = \sqrt{\frac{2fb}{nRK}}$$

where a = optimum number of lines.

b = ratio of active horizontal scanning period, to active vertical scanning period

$$= \frac{82{\cdot}5\%}{93{\cdot}6\%} = 0{\cdot}88.$$

f = video bandwidth of received television signal, which in a practical receiver would be in the order of 5·0 Mc/s.

n = number of pictures per second = 25.

R = aspect ratio of the picture = 4/3.

K = "Kell" factor = 0·64 †

Substituting figures in the above equation we have

$$a = \frac{2\times 5\times 10^6\times 0{\cdot}88}{25\times 1{\cdot}33\times 0{\cdot}64} = 640 \text{ lines.}$$

It is usually convenient for technical reasons to choose a line number whose prime factors are small. In this case a suitable choice of line number would be 625.

---

† The Kell factor (after R. D. Kell of the R.C.A. Laboratories) is the ratio of observed vertical resolution to the number of lines making up the picture. It is less than unity in view of the fact that the random vertical picture information does not fall exactly on the scanning lines of the picture. This factor was determined by experiment.

Appendix 2

A Comparison of World Television Standards

Television standard	B.B.C./I.T.A.	Russian O.I.R.	W. European	F.C.C.	French
Principal countries of use	Gt. Britain and N. Ireland	Russia, E. Germany, Hungary, Czechoslovakia, Poland	Most of W. Europe excluding Britain and France	U.S.A., Iran, Canada, Japan, Thailand and certain Latin American countries	France, Morocco, Algeria, Tunisia
No. of lines	405	625	625	525	819
Channel width	5·0 Mc/s	8·0 Mc/s	7·0 Mc/s	6·0 Mc/s	14·0 Mc/s
Sound–Vision carrier spacing	3·5 Mc/s	6·5 Mc/s	5·5 Mc/s	4·5 Mc/s	11·15 Mc/s
Transmitted video bandwidth	3·0 Mc/s	6·0 Mc/s	5·0 Mc/s	4·0 Mc/s	10·4 Mc/s
Line frequency c/s	10 125	15 625	15 625	15 750	20 475
Field frequency c/s	50	50	50	60	50
Active line scanning period	82 μs	52 μs	52 μs	53 μs	40 μs
Time for one picture element	0·166 μs	0·084 μs	0·100 μs	0·124 μs	0·048 μs
Number of active vertical scanning lines (mean of tolerances)	376	576	576	490	737
†Vertical resolution (active vert. lines × 0·7) at best.	0·7 × 376 = 263	0·7 × 576 = 402	0·7 × 576 = 402	0·7 × 490 = 343	0·7 × 737 = 516
†Horizontal resolution (video freq. × active line scanning period × ¾ × 2)	369 (308)‡	468 (429)‡	390 (352)‡	319 (280) ‡	625 (592)‡
Ratio hor./vert. resolution	1·4 (1·17)‡	1·16 (1·06)‡	0·97 (0·88)‡	0·93 (0·82)‡	1·21 (1·15)‡
Total number of picture elements (vert. res. × hor. res. × 4/3).	129 000 (108 000)‡	251 000 (230 000)‡	208 000 (188 000)‡	146 000 (128 000)‡	430 000 (406 000)‡
Elements per picture width	490 (410)‡	625 (570)‡	520 (470)‡	425 (374)‡	835 (790)‡
Vision modulation	A.M. positive	A.M. negative	A.M. negative	A.M. negative	A.M. positive
Sound modulation	A.M.	F.M.	F.M.	F.M.	A.M.

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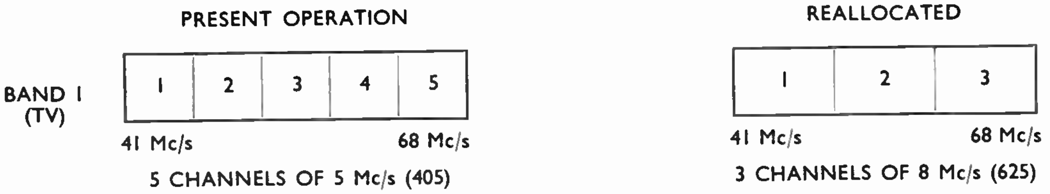
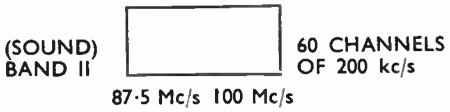
† Vertical resolution is defined here as the number of picture elements which may be resolved in the picture height, and horizontal resolution is the number of picture elements which may be resolved, in a horizontal direction, in a width *equal* to the picture height.

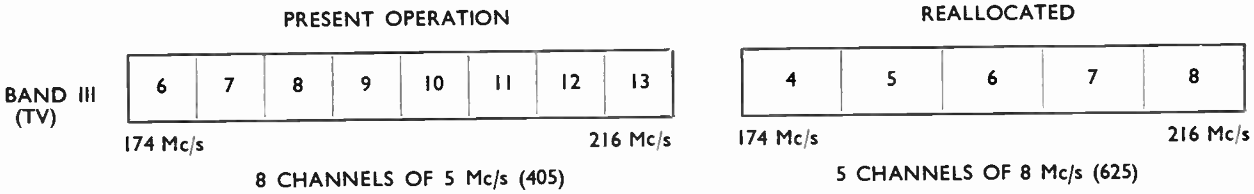
‡ The values in the above table in brackets are based on the video bandwidth likely to be realized in practice, which has been taken as 0·5 Mc/s less than the transmitted video bandwidth.

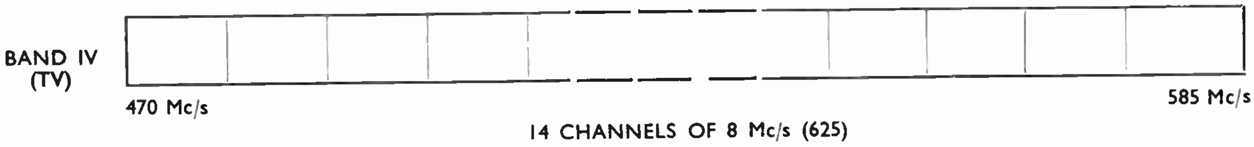
Appendix 3

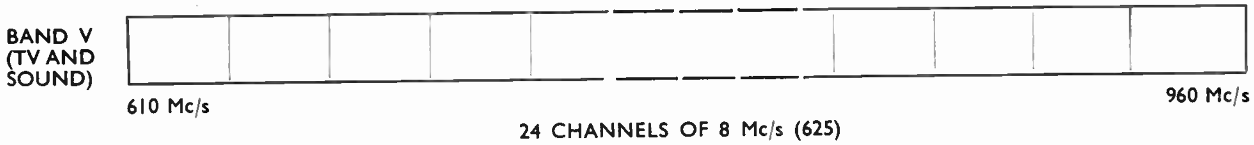
Frequencies Available for Sound and Television Broadcasting in the V.H.F. and U.H.F. Bands











INSTITUTION DINNER 1961

All members have received individual notice of the Institution's Dinner which is to be held at the Savoy Hotel, London, on Thursday, 8th June. Applications for tickets and acceptance of invitations are still being received but the following are some of the members and guests who have already indicated their intention of being present:—

Adorian, P., F.C.G.I. (*Past President*)
Alexander, Sir William, L.H.D., Ed.B., M.A., B.Sc. (*Secretary, Association of Education Committees*)
Anderson, Wing Commander E. W., O.B.E., D.Sc., A.F.C. (*President, The Institute of Navigation*)
Armstrong, Dr. V. (*Scientific Liaison Officer, New Zealand*)

Bailey, J. E. C. (*Chairman, British Scientific Instrument Research Association*)
Bedford, L. H., C.B.E., M.A., F.C.G.I. (*Past President*)
Bennett, Dr. R., V.R.D., M.P. (*Chairman, Parliamentary and Scientific Committee*)
Best, Commander K. B., M.V.O., O.B.E.
Booth, A. D., Ph.D., D.Sc. (*Member of Council*)
Bovill, C. B.
Bragg, Sir Lawrence, F.R.S. (*Director, Davy Faraday Research Laboratory*)
Brightmore, Air Commodore A. G. P.
Brinkley, J. R. (*Member, Convention Committee*)
Brockman, Rear Admiral R. V., C.S.I., C.I.E., C.B.E.
Brown, Air Vice-Marshal C. P., C.B., C.B.E., D.F.C. (*Vice-President*)
Brundrett, Sir Frederick, K.C.B., K.B.E. (*Scientific Adviser, Civil Service Commission*)

Caldecote, The Rt. Hon. the Viscount, D.S.C., M.A.
Carroll, Sir John, K.B.E., Ph.D., F.R.S.E. (*Deputy Controller, Research and Development, Admiralty*)
Cawood, W., Ph.D. (*Chief Scientist to the Army Council*)
Chamberlain, Air Vice-Marshal G. P., C.B., O.B.E.
Chapman, Air Vice-Marshal H. H., C.B., C.B.E. (*Director-General, Technical Services, Air Ministry*)
Chapman, S. R., M.Sc.
Clarke, Rear Admiral Sir Philip, K.B.E., C.B., D.S.O. (*Past President*)
Clifford, Graham D. (*General Secretary*)
Cole, Major General E. S., C.B., C.B.E. (*Director of Telecommunications, War Office*)
Cook, Sir William, C.B. (*U.K. Atomic Energy Authority*)
Cooper, V. J., B.Sc. (*Chairman, Television Group*)
Coventry, J. E. C. (*Scientific Liaison Officer, Rhodesia and Nyasaland*)
Cox, Sir Harold Roxbee, D.Sc., Ph.D., D.I.C. (*Chairman, National Council for Technological Awards*)
Cunliffe, The Hon. Geoffrey (*British Standards Institution*)

De Soyza, His Excellency G., O.B.E. (*High Commissioner for Ceylon*)
Diver, F. G., M.B.E. (*Chairman, Technical Committee*)
Drew, H. E. (*Chairman, Membership Committee*)
D'Rozario, Dr. A. M. (*Indian Scientific Adviser*)
Dummer, G. W. A., M.B.E. (*Member, Convention Committee*)
Durlacher, Vice-Admiral Sir Lawrence, K.C.B., O.B.E., D.Sc. (*Deputy Chief of Naval Staff and Fifth Sea Lord*)
Dyson, A. A., O.B.E. (*Member of Council*)

Fletcher, Colonel R. C.
Florey, Sir Howard (*President of the Royal Society*)
Foster, H. G., M.Sc.

Hardman, H., C.B. (*Permanent Secretary, Ministry of Aviation*)
Heightman, D. W.
Hill, C. Gray (*Solicitor to the Institution*)
Hill, Lt. Col. D. V. (*Christ Church, Oxford*)
Holbein, A. M., C.B.E., B.Sc. (*Chairman, Educational Policy, City and Guilds of London Institute*)
Hurren, S. A., O.B.E., M.C. (*Past President*)

Jarrett, Sir Clifford, K.B.E., C.B. (*Secretary of the Admiralty*)
John, Admiral Sir Caspar, G.C.B. (*First Sea Lord and Chief of Naval Staff, Board of Admiralty*)
Jones, Air Marshal Sir Owen (*President, Royal Aeronautical Society*)

King, J. A. (*Scientific Liaison Officer, South Africa*)
Kirkland, Lt. Col. G. W., M.B.E. (*President, The Institution of Structural Engineers*)
Knowles, Air Vice-Marshal E., C.B.E., B.Sc. (*Director of Education, Royal Air Force*)

Lampitt, R. A. (*Chairman, West Midlands Section*)
Leak, H. J. (*Chairman, Electro-Acoustics Group*)
Lindley, A. L. G. (*Chairman, G.E.C. Ltd.*)
Lugg, Group Captain S., C.B.E.

Macdonald, His Excellency T. L. (*High Commissioner for New Zealand*)
Mallaby, Sir George, K.C.M.G., O.B.E. (*First Commissioner, Civil Service Commission*)
Malloch, Dr. J. G. (*Scientific Liaison Officer, Canada*)
Marriott, G. A., B.A. (*Past President*)
Maydon, Lt. Col. S. L. C., M.P.
Melrose, Dr. D. G. (*Department of Experimental Surgery, Postgraduate Medical School*)
Melville, Sir Harry, K.C.B., D.Sc., F.R.S. (*Secretary, D.S.I.R.*)
Miller, W. E., M.A. (*Past President*)

Norman, J.
Norris, Vice-Admiral Sir Charles, K.B.E., C.B., D.S.O. (*Director, British Productivity Council*)

O'Brien, W. J.
Orr, Captain L. P. S., M.P.
Ottway, G. C. (*President, Scientific Instrument Manufacturers' Association*)

Pandit, Her Excellency Mrs. V. L. (*High Commissioner for India*)
Parker, Colonel J. D., M.B.E.
Parsons, D. R.
Pretty, Air Vice-Marshal, W. P. G. (*A.O.C.-in-C., Signals Command*)
Pryor, R. G. (*President, Institution of Production Engineers*)

Radley, Sir Gordon, K.C.B., C.B.E., Ph.D.
Ratcliffe, J. A., C.B.E., F.R.S. (*Director, Radio Research Station*)
Rawson, Sir Stanley (*British Standards Institution*)
Reid, Admiral Sir Peter, K.C.B., C.V.D. (*Third Sea Lord and Controller, Admiralty*)

Schwarz, H. F., B.Sc. (*Member of Council*)
Shirley, Air Vice-Marshal T. U. C., C.B.E. (*Deputy Controller of Electronics, Ministry of Aviation*)
Spreadbury, E. A. W. (*Chairman, Radio Trades Examination Board*)
Spreckley, Air Marshal Sir Herbert, K.B.E., C.B. (*Head of Civil Establishments, Air Ministry*)
Stewart, J. G., C.B., C.B.E. (*Under-Secretary, Ministry of Labour*)
Sutherland, F. N., C.B.E., M.A.
Sutton, Sir Graham, C.B.E., D.Sc., F.R.S. (*Director-General, Meteorological Office*)

Tait, Air Vice-Marshal Sir Victor, K.B.E., C.B.E
Taylor, D., Ph.D. (*Member, Convention Committee*)
Taylor, G. A. (*Hon. Treasurer*)
Taylor, W. J., C.B.E., M.P. (*Parliamentary Under-Secretary of State for Air*)
Thompson, J. L. (*Vice President*)
Truscott, D. N., O.B.E., Ph.D., Sc.D. (*Chairman, Electronic Engineering Association*)
Turner, Captain G. C., A.D.C., R.N.
Tyler, Major-General Sir Leslie, K.B.E., C.B.

Ungoed-Thomas, Sir Lynn, Q.C., M.P.
Uttley, A. M., Ph.D. (*Superintendent, Autonomics Division, N.P.L.*)

Wagner, A. R., C.V.O., D.Litt. (*Richmond Herald*)
Wakefield, Sir Wavell, M.P.
Whiteley, A. H., M.B.E. (*Member of Council*)
Williams, Professor Emrys, Ph.D. (*Vice President*)
Willis, H. F., C.B., M.Sc., Ph.D. (*Chief of Royal Naval Scientific Service*)

Yousuf, His Excellency Lt. Gen. M. (*High Commissioner for Pakistan*)
Ya'acob, His Excellency Y. T. M. Tunku (*High Commissioner for Malaya*)

Zepler, E. E., Ph.D. (*President*)
Zuckerman, Sir Solly, C.B., F.R.S. (*Scientific Adviser, Ministry of Defence*)

U.D.C. 537.531: 621.387.4: 621.397.62

Aspects of the Emission of X-Rays from Television Receivers

By

M. V. CALLENDAR, M.A.†

AND

D. F. WHITE, B.Sc.†

Summary: The principles involved in x-ray dosimetry are reviewed. The characteristics and limitations of ionization chambers, photographic film, Geiger and proportional counters and scintillation detectors when used for low intensity dose-rate measurement are discussed. A proportional counter is used to study the spectra of x-radiation from cathode ray tubes and rectifiers. The energy response characteristics necessary in a counter for dose rate measurements are given.

A proportional counter is used to measure the x-ray dose rates from typical television receiver cathode ray tubes. The radiation level is found to be considerably below the present maximum permissible dose rate. A theoretical treatment of the x-ray spectrum transmitted through a glass envelope shows general agreement with the observed spectra from cathode ray tubes.

1. Introduction

The tremendous increase in the use of artificially produced radioactive isotopes in recent years and the measures taken to ensure the protection of personnel working with these radiations has led to a closer examination of other potential sources of ionizing radiation to which the general public may be exposed, such as x-ray machines for shoe fitting, luminous instrument dials and equipment employing cathode ray tubes and high voltage rectifiers, such as television receivers.

It has been recommended in B.S. 415:1957 and also by the International Commission on Radiological Protection that the x-radiation from equipment containing cathode ray tubes and rectifying valves shall nowhere exceed 0·5 milliröntgens per hour (mr/hr) at the outside surface of the equipment housing.

In cases where the amount of x-radiation is considerable and this 0·5 mr/hr contour occurs at some distance from the source of radiation, its measurement poses no particular problems, being measurable by an ionization chamber. In the particular case of television receivers, however, in which possible sources of x-radiation are localized, and the radiation itself is of low intensity, special techniques are required for accurate measurement, particularly since such measurement must usually be made in the presence of both electric and magnetic fields. X-radiation from such sources is also normally less than 20 keV in energy, so that particular attention must be paid to the response of the radiation detector in the low energy region.

If it is required to set an upper limit of 0·5 mr/hr it is desirable that the instrument for measuring the x-ray dose rate should have a sensitivity of at least 0·1 mr/hr.

Both ionization chambers and Geiger counters have been used in attempting to measure the x-ray dose rate from television receivers, but the results are open to errors which will be considered later.

2. Fundamental Considerations

In dosimetry it is convenient to specify x-rays in terms of their energy in electron volts, E, rather than in terms of their wavelength in angstroms, λ. The relationship between the two units is

$$E = 12396\lambda^{-1} \qquad \text{......(1)}$$

A comprehensive review and bibliography on the subject of radiation dosimetry has been given by Hine and Brownell.[1]

The most widely used unit of radiation dose is the röntgen. The definition of the röntgen as the quantity of x or gamma radiation such that the associated corpuscular emission per 0·001 293 gramme of air (i.e. 1 cm^3 of dry air at 0°C and 76 cm Hg pressure) produces ions carrying 1 e.s.u. of electricity of either sign, implies that the maximum current derivable from air subject to a dose rate of 1 mr/hr will be $9{\cdot}24 \times 10^{-17}$ amps per cm^3 at s.t.p. This fact is the basis for ionization chamber measurements of dose rate.

The same definition, however, may be considered in terms of the energy dissipated by the radiation in its passage through air. If q individual x-ray photons, or quanta, each of energy E electron volts are incident upon an area of 1 cm^2 the total energy flux per cm^2 will be qE electron volts. A certain fraction of

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† Ekco Electronics Ltd., Southend-on-Sea, Essex.

these photons will be absorbed in each centimetre of air path, this fraction being $(1-\exp(\mu_l))$ since the absorption follows the usual exponential law, where μ_l is the true linear absorption coefficient of air for x-rays of energy E. In practice $\mu_l \ll 1$, so that this fraction is equal to μ_l. Thus the total energy lost per cm of air will be $\mu_l\, qE$ electron volts.

The average energy, W, lost by radiation in producing one ion pair is 34 electron volts. This is the value recommended by the International Commission on Radiological Units and Measurements (ICRU) in 1957; previously the value 32·5 had frequently been used. Thus the creation of 1 e.s.u. of charge, corresponding to $2{\cdot}08\times10^9$ ions, involves a total energy absorption of $W\times2{\cdot}08\times10^9$, i.e. $7{\cdot}07\times10^{10}$ electron volts, so that from the definition of the röntgen an energy loss of 1 electron volt per cm^3 corresponds to a dose of $1{\cdot}41\times10^{-11}$ röntgens.

We already have, however, that the energy lost per cm by q photons/cm^2 of energy E is $\mu_l qE$, and hence the dose R in röntgens delivered by this radiation will be

$$R = 1{\cdot}41\times10^{-11}\,\mu_l qE \qquad \ldots\ldots(2)$$

Converting this into the more usual units of dose rate, D, in milliröntgens per hour produced by a radiation flux of n photons/cm^2/s of energy E keV gives

$$D = 6{\cdot}58\times10^{-5}\,n\mu E \qquad \ldots\ldots(3)$$

where μ is the true *mass* absorption coefficient of air ($\mu_l = \mu\times$density of air).

Equation (3) is the basis for all techniques of measuring dose rate by means of photon detectors (e.g. Geiger counters).

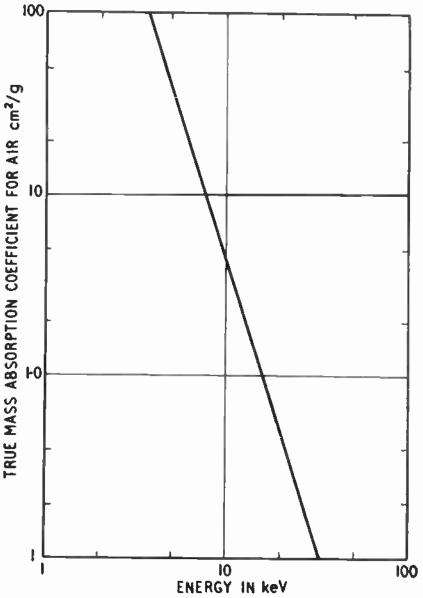

Fig. 1. True mass absorption coefficient of air as a function of x-radiation energy.

The value of μ is accurately known for air both from theory and experiment. Figure 1 shows the variation of μ with the energy of the radiation. Equation (3) may be written as

$$N = \frac{1}{6{\cdot}58\times10^{-5}\mu E} = 1{\cdot}52\times10^4/\mu E \ \ldots\ldots(4)$$

where N is the photon flux required to produce a dose rate of 1 mr/hr. From eqn. (4) and Fig. 1 the value of N may be derived for any value of E, and the result is plotted in Fig. 2.

Figure 2 shows that in the energy range of interest here, i.e. less than 30 keV, the photon flux required to produce a given dose rate increases very rapidly with energy. In this energy range where the absorption of x-radiation energy by air is almost entirely due to the photoelectric effect, μ is varying approximately as $E^{-3{\cdot}1}$ and the curve in Fig. 2 closely follows the law

$$N = 2{\cdot}87\,E^{2{\cdot}1} \qquad \ldots\ldots(5)$$

up to an energy of 30 keV.

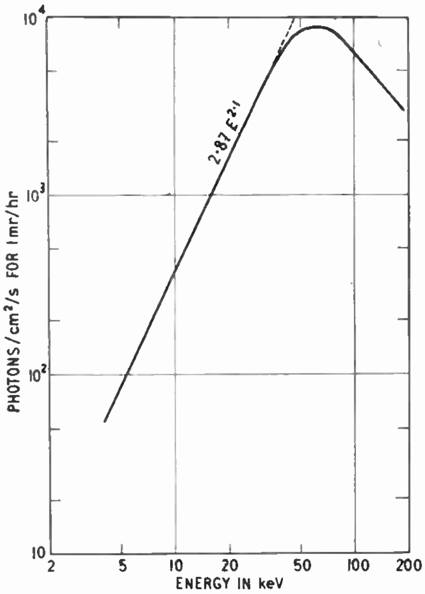

Fig. 2. Photon flux required for a dose rate of 1 mr/hr at various energies.

3. Methods of Measurement

The foregoing considerations show that there are two basically distinct methods of measuring radiation dose. In the first group the amount of ionization produced by the radiation in its passage through matter is measured, whereas in the second group the individual photons are detected and the dose rate derived from the photon flux thus measured.

The various types of radiation detector may therefore be grouped as follows:

Ionization type:	Ionization chambers Photographic film
Photon detectors:	Geiger counter Scintillation counter Proportional counter

3.1. *Ionization Type*

3.1.1. Ionization chamber

Since the röntgen is defined in terms of the ionization produced in air, the saturation current from an ionization chamber of suitable volume and wall material gives a direct measurement of dose rate, provided the dose rate is uniform throughout the chamber, regardless of the energy of the x-radiation being measured.

Since at a radiation level of 0·1 mr/hr this current is only $9{\cdot}24 \times 10^{-18}$ amps per cm^3 a chamber of considerable volume ($\sim$1 litre) is necessary to obtain a measurable current. Because the intensity of radiation decreases as the square of the distance from a point source, such a chamber will only give a true indication of dose rate at distances from the source that are large compared with the dimensions of the chamber, otherwise the dose rate will decrease significantly through the depth of the chamber. In the case of television receivers the dose rate is normally below 0·1 mr/hr at distances of a few inches from the set, so the use of an ionization chamber can give erroneous results.

3.1.2. Photographic film

The blackening of photographic film by x-radiation is frequently used for monitoring purposes and has the advantage that very localized measurements are possible, thus overcoming the difficulties mentioned above for ionization chambers. For a reasonable degree of blackening, however, it is necessary for a total integrated dose of at least 50–100 mr to be received by the film. At the radiation levels here being considered this involves exposure times of 100 hours and upwards for each measurement. A second difficulty is that for low energy radiation, less than 50 keV, the calibration of the film in terms of dose is very dependent upon the energy of the incident radiation.

3.2. *Photon Detectors*

A review of the use of Geiger, proportional and scintillation counters specially adapted for the detection of low energy x-radiation has been given by Long,[2] and also by Taylor and Parrish,[3] and the details of their action will only be reconsidered here in connection with their use for dose rate measurements. Their suitability for this purpose depends on the ability to make their counting efficiency vary with x-ray energy approximately as $E^{-2{\cdot}1}$ (see eqn. (5)) so that the count rate per mr/hr remains reasonably constant over the range of energies of interest. Failing this their output must be such that it is possible to distinguish between radiations of different energies. Each detector in turn will be considered from this point of view.

3.2.1. Geiger counter

The Geiger counter responds to x- and gamma-radiation by virtue of the secondary electrons ejected from the wall material and the gas filling of the counter. These secondary electrons are created, in the region of energies below 50 keV, by photoelectric absorption of the incident x-ray photons. The field strength in a Geiger counter is such that a single free electron in the sensitive volume of the counter will create an "avalanche" resulting in a pulse of charge at the anode of sufficient magnitude that a pulse amplitude of several volts is obtainable. The overall detection efficiency may be in the region of 0·1 – 1%. Thus at, say, 10 keV energy a typical end window Geiger counter with a window diameter of 2 cm may give a count rate of 1–10 c/s for a dose rate of 1 mr/hr. The chief objection to a Geiger counter is that all output pulses are of the same amplitude, regardless of the energy of the incident radiation. Thus, unless the variation in efficiency with energy is the exact inverse of the photon flux required to produce a given dose rate, the results will vary with the energy distribution of the incident radiation.

In particular, an x-ray tube emits a broad spectrum of x-ray energies whose shape is a function not only of the h.t. voltage applied to the tube but also of the target material, window material and thickness. The spectrum will also vary with the nature and degree of any filtration applied to the output from the tube.

In the same way the spectrum of the x-radiation from cathode-ray tubes and rectifiers will be dependent upon the thickness and composition of the envelope and any intervening electrodes. The absorption of radiation in the envelope will be particularly dependent upon the quantities present of any element of high atomic number. In particular the spectrum transmitted by a lead glass envelope will be very different from that transmitted by a soda glass envelope.

Thus even if a Geiger counter has an apparently flat response over a range of energies when calibrated by means of an x-ray tube and using aluminium filtration, a very different response curve may be obtained if different filtration is used. As a typical example, an MX108 Geiger counter, which is one of the few with an apparently flat response up to about 18 keV, was calibrated by means of an x-ray tube using aluminium filtration on the x-ray beam. Lead filtration was then substituted for aluminium and the count rate again noted for the same dose rates. The results are shown in Table 1.

Table 1

keV	Count/min for 1 mr/hr		Ratio Al/Pb
	Al filter	Pb filter	
15	216	196	1·10
18	266	202	1·32
20	297	209	1·42
22	360	284	1·27
25	423	343	1·23
30	490	504	0·97

The L-absorption edges of lead are in the region 15·8–13·0 keV, so that attenuation of x-radiation just above these energy limits will be very heavy. At energies of less than 13 keV the absorption coefficient of lead drops suddenly, and also a certain amount of the radiation of energies greater than that of the absorption edge will generate the characteristic L x-rays of lead from 10–12 keV. The nett result is that although the x-ray tube may be operating at 20 keV, the majority of the radiation is being emitted at approximately 12 keV and as the dose per photon increases as E^{-2} almost the whole of the radiation dose rate will be due to the 12 keV component. More important still, as the voltage is raised, the majority of the radiation will remain at 12 keV rather than increasing in energy. Table 1 shows that differences in calibration of up to 42% were observed even though the degree of filtration was not chosen to give the maximum effect.

This effect must always occur where the radiation passes through a material which has an absorption edge within the range of energies being emitted by the x-ray source and is particularly important in the present type of application where envelopes may be either of soda glass or lead glass.

Even the use of aluminium filters of a thickness which is varied with the voltage applied to the x-ray tube leads in a similar way to a calibration curve which is as much a function of the filtration as of applied voltage. Thus at 20 keV a difference of over 20% in calibration can be obtained with the MX108 simply by alteration of the aluminium thickness, the apparent sensitivity increasing with filtration, as would be expected from the shape of the calibration curves given by Ciuciura[4] and Oosterkamp *et al.*[5] From these calibration curves the sensitivity is clearly rising rapidly with increasing energy above 15 keV. Therefore with higher degrees of aluminium filtration (thereby increasing the mean energy of the transmitted radiation for any given voltage applied to the x-ray tube), the commencement of the steep rise in calibration will be shifted to lower values of applied voltage. With higher applied potential the effect naturally becomes more pronounced. Thus although a Geiger counter may be calibrated empirically by comparison with a standard ionization chamber using an x-ray tube, this calibration may be seriously in error if applied to a source of radiation with a different energy spectrum, unless the dose rate response is truly flat over the range of all emitted x-ray energies.

3.2.2. Scintillation counter

As described by Long,[2] a thin thallium-activated sodium iodide crystal with a beryllium window in conjunction with a suitable photomultiplier will detect x-rays of less than 50 keV energy with approximately 100% efficiency. Although on average the output pulses from such an arrangement are proportional in amplitude to the energy of the incident radiation, there is a large spread in amplitude from pulses due to radiation of a given energy. Due to this poor energy resolution it is difficult to translate the count rate from a scintillation counter into a dose rate.

3.2.3. Proportional counter

As a pulse counter this has the merit that the output pulses are strictly proportional in amplitude to the energy of the ionizing event detected. The gas filling is generally either argon or xenon with a small amount of methane added, the total pressure being approximately equal to atmospheric. Such a counter has good discrimination between radiations of different energies. Thus a 15 keV x-ray will give a resolution (ratio of width of the pulse height distribution at one half its maximum height to the amplitude of the mean pulse height) of 10%.

The sensitivity of such counters is such that count rates of 400 c/s for a dose rate of 1 mr/hr are usual and a sensitivity of 2000 c/s per mr/hr is easily attainable.

The pulses from proportional counters are necessarily very small (of the order of 1 mV), and as pulse height analysers require a pulse of at least 5 volts, a linear amplifier with a gain of at least 10 000 is required. Such a counter was used to plot the actual spectrum of energies being emitted by a source of x-radiation. By suitable choice of gas filling the variation of efficiency with radiation energy may be made such that its response in terms of count rate per unit dose rate does not vary by more than 10% over the range from 7·5 to 20 keV. The high sensitivity means that only short counting times are required in order to obtain a reasonable statistical accuracy.

4. Experimental

4.1. *Equipment*

In view of the above considerations a proportional counter was chosen in an attempt to provide accurate dose rate measurements over the range of energies concerned in commercial television receivers.

The particular counter used in this work was the PX130 (20th Century Electronics Ltd.) which has a 22 mm diameter beryllium side window 0·020 in. thick. The effective window area, after allowing for the window supports which are effectively opaque to x-rays, is 3·58 cm². The pulses from it were fed via a conventional linear amplifier A.E.R.E. type 1430 (or Ekco N568) which, with its associated head amplifier, has an overall maximum gain of 120 dB, to an Ekco N600 ratemeter. The latter incorporates a 0–2500 volt stabilized high voltage supply suitable for operating the PX130 which has a working voltage in the region of 2050 volts. Also incorporated is a pulse height analyser.

More recently a combined scaler/ratemeter has become available (Ekco N683) incorporating not only the necessary h.v. supplies and a pulse height analyser but also an amplifier with a gain of 100 000 which is sufficient to amplify the output pulses from the proportional counter to a suitable level (5–50 volts) to operate the analyser, enabling the A.E.R.E. type 1430 amplifier to be dispensed with, and all the electronics to be contained in a single unit of overall dimensions 17½ in. high × 19 in. wide × 15 in. deep.

Because the pulses from the counter are small, care must be taken with the cleanliness of all insulators and connectors carrying the h.t. voltage to the counter. Also, when operating close to a television receiver, precautions must be taken with the screening of the cable from the counter to the head amplifier, which should preferably be double-screened.

4.2. *Efficiency of Counter*

This can be established from a knowledge of the transmission of the beryllium window and the absorption coefficient of the filling gas at various energies.

The efficiency η will be given by

$$\eta = \exp(-\mu_w \rho_w t)[1-\exp(-\mu_g \rho_g d)] \quad ......(6)$$

since the absorption of radiation follows an exponential law, where μ_g, μ_w are the mass absorption coefficients of gas filling and window respectively (both functions of E), ρ_g and ρ_w are the respective densities, d the diameter of the counter and t the thickness of the window. The term $\exp(-\mu_w \rho_w t)$ in eqn. (6) gives the fraction of the radiation that will pass through the beryllium window, and the term $(1-\exp(-\mu_g \rho_g d))$ the fraction that will be absorbed, and therefore detected, within the counter gas. The mass absorption coefficients of argon, xenon and beryllium are shown in Figs. 3 and 4. With $\rho_w = 1{\cdot}83$ for beryllium and $t = 0{\cdot}05$ cm, the window transmission is 57% at 5 keV, rising to 80% at 7 keV and over 90% for energies above 10 keV. Subsequent observation showed that almost the whole of the x-radiation from television receivers is above 7 keV in energy, so that window absorption is not a serious factor.

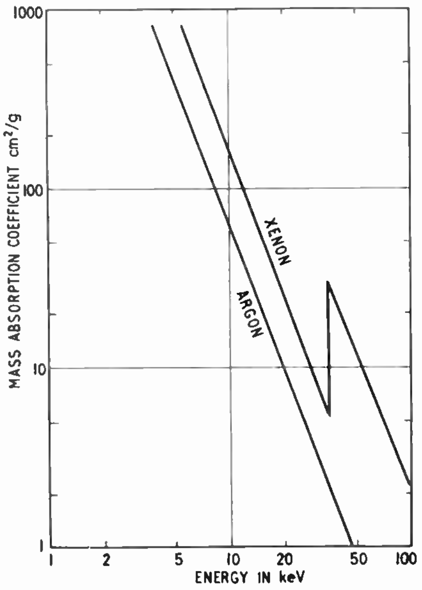

Fig. 3. Mass absorption coefficients of xenon and argon.

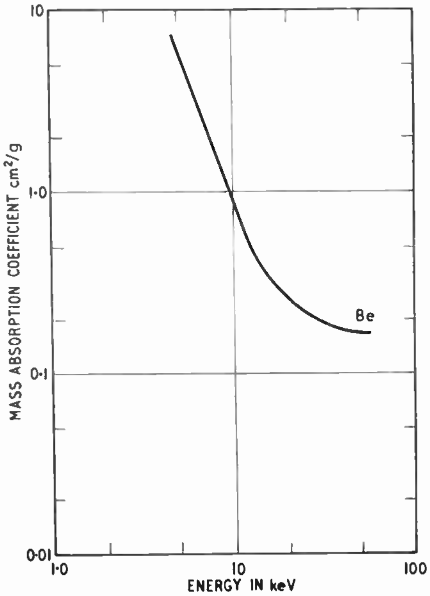

Fig. 4. Mass absorption coefficients of beryllium.

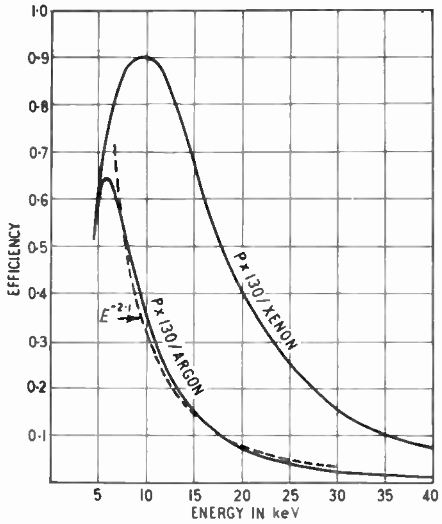

Fig. 5. Variation of counter efficiency with energy of radiation.

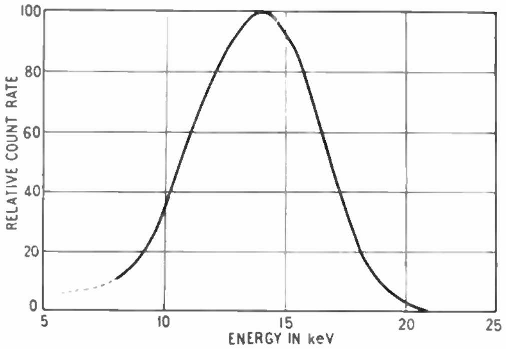

Fig. 6. Typical x-radiation spectrum from an x-ray tube.

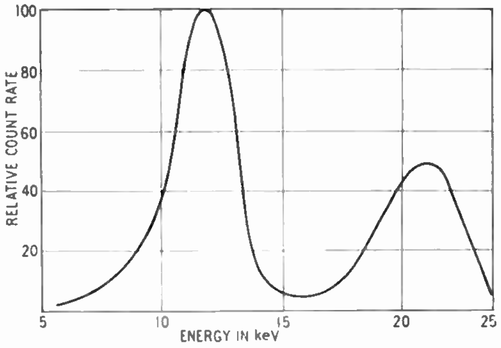

Fig. 8. Spectrum of x-rays from a rectifier with lead glass envelope.

Clearly the general shape of the efficiency characteristic can be modified within limits by alteration of the pressure of the gas filling (affecting ρ_g). In the present experiments the counters were filled to a pressure of 70 cm Hg with argon or xenon and made up to 76 cm Hg by the addition of 6 cm Hg methane. The effect of the methane upon the absorption characteristics of the gas filling is negligible.

Using Figs. 3 and 4 and eqn. (5) the efficiency of the argon or xenon-filled counter may be calculated for any energy of x-radiation. Argon and xenon efficiencies for the present counter which has an effective depth of 4·8 cm and a filling pressure of 70 cm, are shown in Fig. 5.

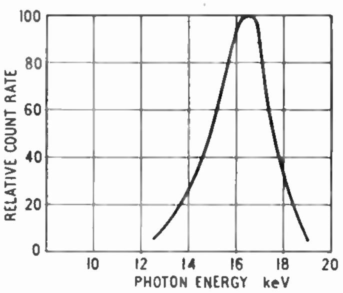

Fig. 7. Spectrum of x-rays from a cathode-ray tube.

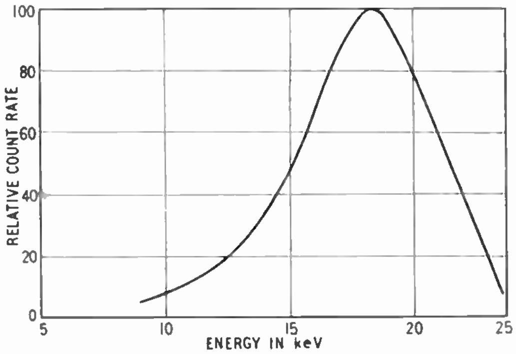

Fig. 9. Spectrum of x-rays from a rectifier with soda glass envelope.

4.3. *Energy Spectra of X-radiation*

The threshold bias of the pulse height analyser was calibrated for a given h.t. voltage applied to the counter in terms of radiation energy by the use of a source of radioactive Fe55. This isotope emits a single x-ray of energy 5·9 keV. By suitable choice of amplifier gain it was arranged that the entire range of threshold voltage (5–50 V) on the pulse height analyser corresponded to 2·5 keV–25 keV in x-ray energy. The gate width of the analyser was set to accept an energy range of 1 keV.

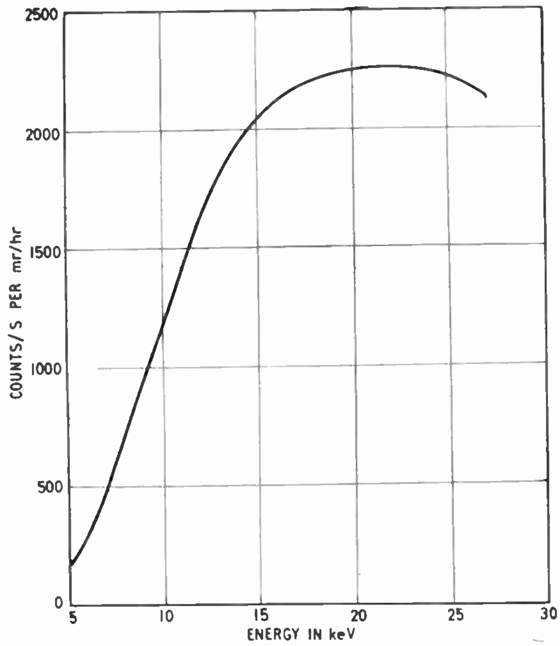

Fig. 10. Variation of count rate per mr/hr with energy from xenon filled PX130. Pressure 70 cm Hg.

By the use of a ratemeter and chart recorder the spectra in Figs. 6–9 were obtained. These curves show that the x-ray energy distribution from an e.h.t. rectifier may be completely different from that of an x-ray tube. The energy spectrum of the radiation from a cathode-ray tube (Fig. 7) is basically similar to that from an x-ray tube (Matchlett Aeromax 12, Fig. 6) but shows a narrower distribution of energies owing to the greater degree of filtration by the walls of the cathode-ray tube. The rectifier (Fig. 8) has a lead glass envelope and the peak in the spectrum in the region of 12 keV arises from the comparatively low absorption by the glass in the region just below the L absorption edge of lead, together with some excitation of the characteristic L-lines in the x-ray spectrum of lead. The spectrum shown in Fig. 9 is for an identical rectifier with a soda glass envelope.

4.4. *Measurement of Dose Rates*

From eqn. (4) and Fig. 5 the count rate per second for a dose rate of 1 mr/hr at any energy was calculated for both the argon filled and xenon filled counter. The results are shown in Figs. 10 and 11. Figure 5 showed that the efficiency of the counter filled with argon to 70 cm Hg pressure follows the desired $E^{-2\cdot 1}$ response characteristic quite closely over a considerable range of energies and in fact from Fig. 11 it can be seen that for this filling the count rate per mr/hr does not vary by more than ± 10% from a mean value of 438 counts per second per mr/hr over the energy range from 7·5 keV to 20 keV. Similarly, over the range 7·5–25 keV there is a maximum variation of ±15% from a value of 417 counts per second per mr/hr.

Both of these error limits can only occur if the whole of the radiation is substantially mono-energetic and occurs at the point of maximum deviation, i.e. 11 keV or the upper end of the range. In practice therefore, for radiation with a distribution of energies, the errors will be less than this.

Increasing the pressure of the argon (or substituting xenon) would raise the high energy end of the response curve but depress the low energy end, i.e. the peak of the curve would move towards higher energies. Reducing the pressure improves the low energy response only slightly because absorption in the window becomes the controlling factor below 7 keV and depresses the high energy response. A series of calculations for various pressures showed that for a counter of the dimensions of the PX130 an argon filling at 70 cm Hg pressure is the most suitable.

In order to check the entire sensitivity curve, an x-ray tube was operated at 30 kVp, the gatewidth of the pulse height analyser being adjusted to accept an energy interval of 1 keV and a spectrum recorded as in Section 3.3. Taking the count rate in each 1 keV interval and using the appropriate count rate to dose rate conversion figure at each energy from Fig. 7, the dose rate was computed, thus using the whole of the calibration curve from 7 keV to 30 keV. At the same time the dose rate was measured with an ionization chamber.

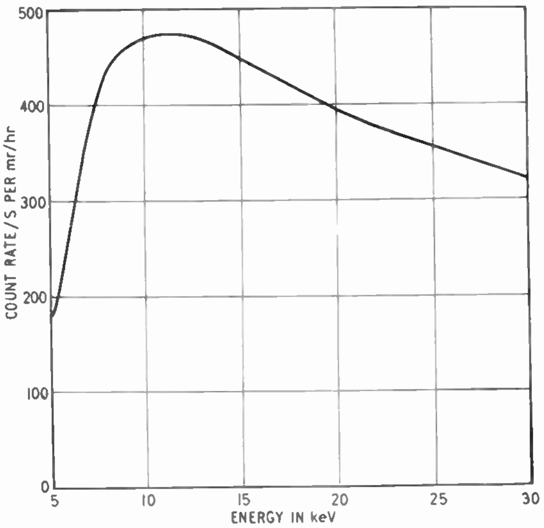

Fig. 11. Variation of count rate per mr/hr with energy for argon filled PX130. Pressure 70 cm Hg.

The output from the x-ray tube was made sufficiently high for the ionization chamber to be placed at a distance from the tube great enough to avoid any geometrical errors. The results of the comparison are shown in Table 2.

Table 2

H.T. kV	Dose rate mr/hr observed	calculated
30	2·1	1·99
30	2·15	2·09
30	1·85	1·78
30	1·2	1·11
30	6·5	6·18
25	1·7	1·59

This scanning technique may always be used when results of the highest accuracy are required, but it is time-consuming as a complete scan may take 45 minutes.

For radiation monitoring work, however, the accuracy given by a single count rate, setting the analyser to accept the whole of the energy spectrum, and applying a single count rate to dose rate conversion figure is normally adequate.

As a compromise for higher energy ranges (e.g. up to 30 keV), or more accurate measurements, it is always possible to split the energy spectrum into two or three intervals, and take count rates using the appropriate mean calibration figure for each interval. For example, if the energy range is 7–30 keV, the pulse height analyser is first set to accept energies between 7 and 20 keV, the count rate measured and converted to a dose rate using the appropriate conversion figure for this range. The analyser is then adjusted to accept energies between 20 keV and 30 keV, and the dose rate computed for this range. As the sensitivity is high, a counting time of one minute in each energy interval is adequate even for dose rates as low as 0·2 mr/hr.

The background count rate with a counter of this type corresponds to less than 0·01 mr/hr.

5. Measurements on Television C.R. Tubes

5.1. *Equipment*

The equipment (using a proportional counter) described in Section 4 was used to examine the radiation from a sample fraction of the c.r. tubes received by the authors' company in January 1960 for incorporation in television receivers. The levels of radiation were, in most cases, so low as to be not reliably measurable by previously available types of radiation measuring equipment: all figures recorded were well below the 0·5 mr/hr limit specified in B.S. 415.

5.2. *Spectrum*

The spectrum (Fig. 6) of the radiation from the tubes showed a fairly sharp peak for photon energies (E), some 2 keV below the maximum possible value $E_{\max}$: here $E_{\max}$ if measured in keV is numerically equal to V_0, the d.c. e.h.t. voltage. This shape agrees in a general way with that estimated from theory in the Appendix.

If I_1 and I_2 are the intensities inside and outside the glass, and t is thickness in cm, we have $I_2/I_1 = \exp(-\mu t)$ for radiation of any one energy (or wavelength).

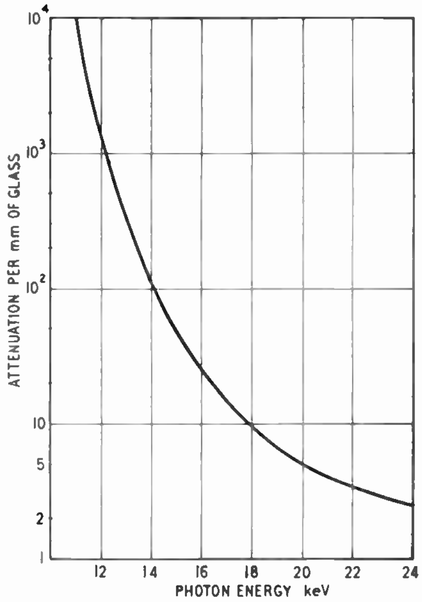

Fig. 12. Estimated attenuation of radiation by the glass of the c.r. tubes.

For the glass here concerned, μ is estimated to be roughly 16 at 20 keV, giving a curve for attenuation against photon energy as shown in Fig. 12. If 19 kV is applied to the tube, the relevant band is 15 keV to 17·5 keV from Fig. 6 and the radiation is thus reduced about 25 times per mm of glass.

The minimum thickness (at the back) is about 5 mm: we should have a variation of some 25/1 in radiation between extreme samples if the thickness were controlled to ± 0·5 mm (precise limits are not available). The front plate thickness is from 6·5 to 8 mm minimum, and thus the radiation from it should be 10^2 to 10^4 times less than the maximum from the back.

5.3. *Distribution of Radiation*

The distribution of the radiation from a typical 21-in. tube was examined in detail. Results are summarized in Fig. 13 where the radiation level is given in terms of the dose rate at the "standard" position, which is the point of maximum radiation on the surface of the smallest rectangular cabinet which could enclose the tube. Attenuation by a wood or plastic cabinet can be assumed negligible.

We see that the maximum dose rate at the surface of the tube is but little greater than the "standard", while that at 1 ft from the cabinet never exceeds about one-tenth of the "standard". Radiation from the front is extremely low, even without the usual glass implosion protection plate.

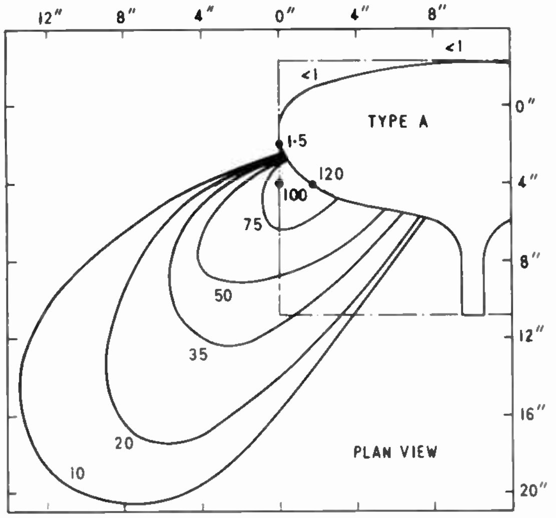

Fig. 13. Distribution of radiation, expressed as a percentage of the maximum at the surface of the cabinet.

5.4. *Variation of Radiation with Current and E.H.T. Voltage*

The variation of radiation with beam current was found to be linear, as expected. A plot of radiation against e.h.t. volts (see Fig. 14) on suitable log-log paper showed a law roughly approximating to $I \propto V^{20}$: this is in general agreement with Ciuciura's results.[4] A useful rough rule for the variation with e.h.t. over the normal range is: dose rate is trebled for each 1 kV increase in e.h.t. voltage.

In view of the critical nature of the relation between radiation and e.h.t. some further investigation was deemed advisable. In the course of this, it was found possible to take a tube (specially selected) up to 26 kV and the results are shown in Fig. 15. A brief theoretical investigation was also made and the results, given in the Appendix, in general agree with the experiments.

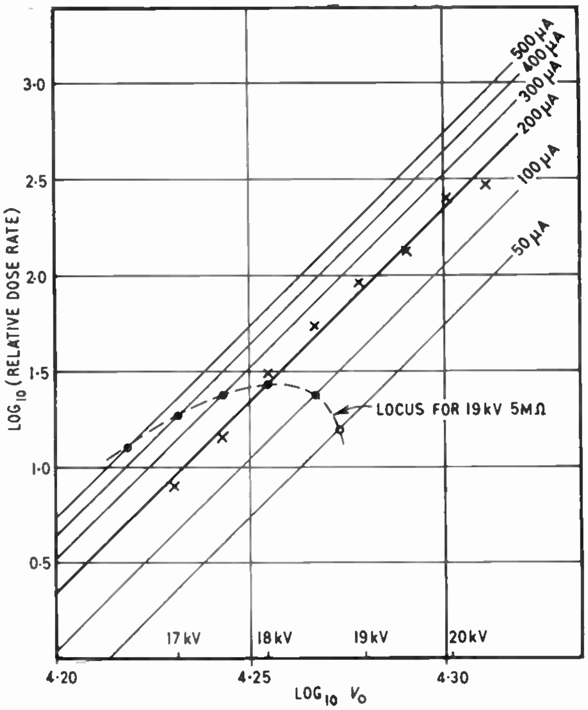

Fig. 14. Experimental results for dose rate versus e.h.t. voltage, are shown by crosses. Full lines show dose rate (approx.) for various currents, and dashed line is locus for 19 kV e.h.t. supply with internal resistance of 5 MΩ.

5.5. *Level of Radiation for a Batch of Tubes*

The radiation at the "standard" position was measured for three batches of 110 deg deflection tubes (as listed in Table 3). It was intended originally to take measurements on larger batches, but this was subsequently not considered worth while in view of the very low levels of radiation encountered, especially with the 17-in. tubes. For the same reason, the statistical method outlined by Ciuciura[4] was not

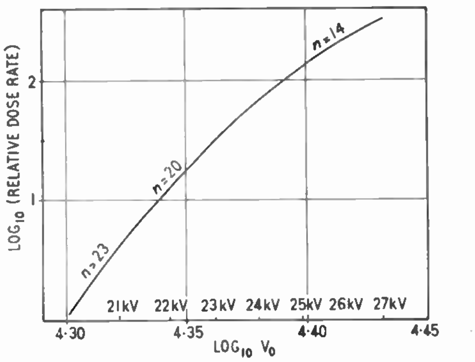

Fig. 15. Measured variation of dose rate with e.h.t. voltage up to 26 kV.

practicable, the e.h.t. required to obtain one "permitted dose" being always well above the maximum permissible for the tube. It is of interest to note that, if a histogram is plotted for count (or dose) rates, a very skew distribution is obtained (see Fig. 16): replotting in terms of log (count rate) restores the symmetry, as anticipated from theory (see Appendix).

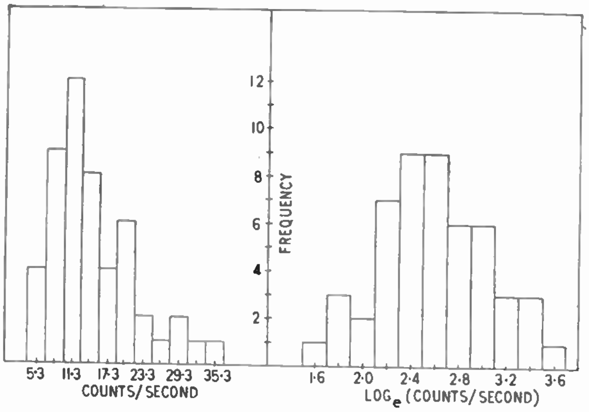

Fig. 16. Histograms for counts/second (N_2) at the standard position for 50 Type "A" tubes, at 19 kV, 200 μA. (440 counts/s = 1 mr/hr).

We require to know the maximum likely voltage of operation. To estimate this, we start from the maker's maximum design centre voltage (18 kV for 21-in. tubes and 16 kV for 17-in. tubes), and then allow for a ± 6% spread of mains supply volts about the nominal, plus a variation of ± 6% in e.h.t. to cover tolerances on components, giving final figures of 20 kV and 18 kV respectively for the two tubes, as measured at zero beam current.

We then consider the variation of radiation with beam current. The receivers in which the tubes were to be used showed a minimum effective resistance of 5 megohms for the e.h.t. supply, and under these conditions the curves for variation of radiation with beam current and with e.h.t. show that the maximum radiation occurs at around 200 μA (Fig. 14). A value between 5 and 10 MΩ is typical for current British receivers: lower values are only encountered with projection or colour tubes, for which special e.h.t. regulation circuits are incorporated.

These considerations show that tests at—

19 kV 200 μA for tubes with 18 kV max. design centre rating

17 kV 200 μA for tubes with 16 kV max. design centre rating

would represent the worst conditions which are likely to be met with in practice. In our case, the 17-in. tubes were actually tested at 19 kV, since the radiation was normally undetectable at 17 kV.

The figures shown in Table 3 were obtained for maximum radiation in mr/hr at the surface of the cabinet with 19 kV e.h.t. and 200 μA beam current:

Table 3

Size	Type	Number tested	Radiation in mr/hr	
			Average	Highest
21 in	"A"	50	0·03	0·08
17 in	"B"	40	<0·01	0·01
17 in	"C"	24	<0·01	<0·01

These are to be compared with the maximum permitted dose rate of 0·5 mr/hr specified in B.S. 415. The background radiation averaged 3 counts per second, corresponding to about 0·006 mr/hr at 15 keV.

It should be noted that these particular tubes were actually used in receivers having a design centre e.h.t. of 16 kV, so that the maximum actual radiation would be only about one tenth of that measured in Table 3.

Tests upon a small number of tubes of other makes, of various dates of manufacture, showed levels of radiation of the same order as those in Table 3, and general characteristics (e.g. the low ratio of front to back radiation) were similar. But quite large variations in level are no doubt possible, arising from variations in the thickness of the glass envelope, and secondarily from differences in the composition of the glass.

5.6. *Conclusions upon the Radiation Level from C.R. Tubes in Television Receivers*

The levels of radiation shown in the above tests are at least one order of magnitude lower than those reported by Ciuciura.[4]

Since the great majority of receivers so far made in Britain use a design centre e.h.t. of 16 kV or lower, it seems likely that the proportion of tubes giving a radiation approaching the B.S. limit in actual receivers in use is vanishingly small. Even on Ciuciura's figures, this proportion would be only of the order of 0·01%. (See his Table 2.)

Nevertheless, it is desirable for manufacturers to maintain a systematic check upon their tubes, especially upon those rated at 18 kV or above.

The level of radiation from the tube at the normal viewing position (about 5 ft from the front) is, of course, extremely low and is likely to lie between 10^{-4} and 10^{-6} mr/hr.

6. Acknowledgments

The writers are indebted to Messrs. E. K. Cole Ltd., for permission to publish this paper. They would also like to thank Messrs. Steele and Suntharalingam for performing most of the tests described.

7. References

1. G. H. Hine and G. L. Brownell, "Radiation Dosimetry", (Academic Press, New York, 1956).
2. A. Long, "Detectors for low energy x-radiation", *J.Brit. I.R.E.*, **19**, pp. 273–87, May 1959.
3. J. Taylor and W. Parrish, "Absorption and counting-efficiency data for x-ray detectors", *Rev. Sci. Instrum.*, **26**, pp. 367–73, April 1955.
4. A. Ciuciura, "Assessment of x-radiation from television receivers", *J.Brit.I.R.E.*, **19**, pp. 469–82, August 1959.
5. W. J. Oosterkamp, J. Proper and J. J. F. de Wijk, "Dosimetry of very weak x-radiation generated in television receivers and x-ray diffraction apparatus", *Philips Tech. Rev.*, **19**, No. 9, pp. 264–7, 1957/58.
6. H. A. Kramers, "On the theory of x-ray absorption and of the continuous x-ray spectrum", *Phil. Mag.*, **46**, p. 836, 1923.

8. Appendix

Calculation of Form of Spectra and of Variation of Radiation with E.H.T. Voltage

We can make an approximate calculation for the shape of the spectrum from a c.r. tube, and for the variation of total x-radiation with e.h.t. (V_0) in the following steps:

(a) We assume (following Kramers[6]) that the spectrum, as generated, for energy flux/cm^2 (I_{1E}) just inside the glass envelope against photon energy E is given by $I_{1E} \propto (E_0 - E)$ where E_0 in keV is numerically equal to V_0.

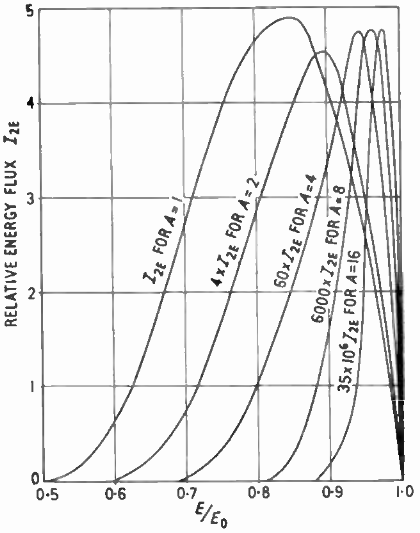

Fig. 17. Calculated x-radiation spectra assuming a normal absorber.

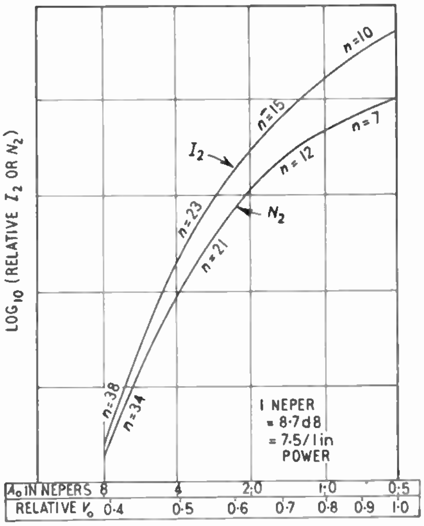

Fig. 18. Calculated curve for variation of x-radiation flux (I_2) with attenuation A_0 and e.h.t. (V_0). Second curve is for counts per second (N_2).

This formula, although apparently the best available, is an idealized approximation: however, the resulting errors are probably unimportant in our case where the subsequent selective attenuation is very large. "Characteristic" radiation is here neglected, since our tests afford no evidence of its presence in any significant quantity: this is fortunate, since little information exists upon its intensity relative to that of the continuous spectrum, or "Bremsstrahlung".

(b) The spectrum for energy flux I_{2E} just outside the glass envelope is then $I_{2E} \propto (E_0 - E) \exp(-\mu t)$ where μ is absorption coefficient and t is thickness of glass. For the case of a "normal" absorber, such that no absorption edge occurs in the relevant range of E, we can put $\mu t = A_0 {E_0}^3/E^3$ where A_0 is the attenuation in nepers for photons of energy $E_0 = V_0$. The form of the spectrum is here dependent solely upon A_0; thus the curves in Fig. 17 apply for any "normal" absorber at that value of V_0 for which the attenuation is A_0.

(c) We now calculate the total radiation

$$I_2 = \int_0^\infty I_{2E} \, . \, \mathrm{d}E$$

from the area under the spectrum curve.

In the general case, we shall have a series of spectra for various values of V_0, and from the area of each we obtain one point upon the required curve for I_2 against V_0. For a "normal" absorber, this process can be conveniently tabulated by taking a series of values of V_0 (each $2^{-\frac{1}{3}}$ times the last) such that A_0 is halved at each step. The results are then plotted on log-log paper as shown in Fig. 18: the slope of this curve gives the value of n in the relation $I_2 \propto V^n$. The value of n is here evidently uniquely dependent upon the attenuation A_0.

(d) We must, however, bear in mind the fact that measurements quoted earlier in this paper are in terms of dose rates (in mr/hr), or more precisely of counts (N_2 per second) and not in energy flux per cm^2. Our calculated spectra for I_{2E} can be transformed, if required, into spectra in terms of N_{2E} by multiplying I_{2E} by a factor η/E, where η is the efficiency of the counter (see Fig. 5 and Sect. 4.2).

(e) The areas under the new spectrum curves for N_{2E} are then found and plotted as before. The resulting curve for N_2 is not greatly different from that for I_2 as can be seen from Fig. 18. The curve for N_2 only applies strictly for our particular counter and for $A_0 = 2$ nepers (or 50 times) at 20 keV, but can be taken without excessive approximation as representing the general relationship between dose rate and A_0 or V_0.

In the case of the Type "A" tubes tested we have a "normal" absorber of high attenuation, the estimated values (for 5 mm thickness) ranging from about 10^{10} at 14 keV to 10^4 at 20 keV and 10^2 at 25 keV. The corresponding laws evaluated from Fig. 17 vary from $n \simeq 28$ for 20 kV to $n \simeq 17$ for 25 kV. The experimental figures are a little lower ($n = 23$ at 21 kV, to 14 at 25 kV, from Fig. 15), but show a similar fall with increasing e.h.t., and the agreement is probably as good as could be expected in view of the lack of exact knowledge of the effective thickness of the glass; this latter is, incidentally, not the same for photons coming from different points on the screen, making an accurate estimate impractical.

The above theory also throws some light upon the probability distribution of radiation from a batch of tubes. We may reasonably assume that variation in glass thickness t is the main cause of variation in radiation, and also that the distribution for t is normal. We have $\log_e I_2 - \log_e I_1 = -\mu t$, and so we should expect a normal distribution for $\log_e I_2$, with a standard deviation μ times that for t. This should also hold roughly for $\log_e N_2$ since the spectrum is quite narrow-band in our case. These assumptions are borne out by Fig. 16 which indicates that the standard deviation in t is of the order of $\pm$ 0·15 mm for the particular batch tested.

If V_x is the e.h.t. voltage at which a specified value (e.g. 0·5 mr/hr) of radiation occurs, we have $\log_e N_2 = n \log_e V_x + \text{constant}$. Thus $\log_e V_x$ (and not V_x) should show a roughly normal distribution with a standard deviation of about μ/n of that for t. But the spread of V_x is so much smaller than that of N_2 that a histogram for V_x (as plotted by Ciuciura[4]) would not necessarily appear noticeably unsymmetrical.

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Component and Valve Reliability in Domestic Radio and Television Receivers

By

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Presented at the Symposium on New Components held in London on 26th–27th October 1960.

Summary: The effect of short and longer term failure of valves and components on the overall reliability of television and radio receivers is considered. Some reasons for failures are reviewed and suggestions for improvement are made. Extracts of statistical information on reliability of relatively large batches of television receivers are presented. The need for information on the predicted reliability of components of new design is emphasized.

1. Introduction

Because it is much more complex than the radio receiver, the television receiver is inherently less reliable and, therefore, this paper, in the main, concentrates on the aspects of television reliability. Nevertheless, many of the observations apply to the radio receiver also.

The average television set has sixteen valves and an assortment of some two hundred and fifty main components, including approximately ninety resistors and ninety capacitors, besides the cathode-ray tube. Whilst this number of individual units contributing to overall reliability (or lack of it) is nothing like as formidable as that of, say, a computer, nevertheless a high degree of individual valve and component quality is called for if the complete television receiver is to give an acceptable standard of reliability. This paper reviews the standard of reliability at present being achieved and the more frequent reasons for failure.

The information given is based on statistics obtained through a quality control scheme concerned with relatively large batches of receivers running in typical field conditions.

2. Failure Rates

A useful yardstick, indicating the general reliability of a television or radio set, is the number of failures per year. Owing to the fact that the number of hours run in any given period varies from household to household, it will be realized that such a yardstick will only be approximate. Over large batches of receivers, however, a fair degree of accuracy is obtained. It can be assumed that the average television receiver runs twenty-five hours per week, or one hundred hours per lunar month. Valve and component failure rate can generally more suitably be stated as a percentage failing in a given period.

2.1. *Early Failure*

Figure 1 illustrates the failure rate on a monthly basis for the first six months of life of a typical batch of fairly modern television receivers. From this figure it will be noted that the rate of failure in the first month is nearly double that after three to four months of use. This characteristic, which is typical of most electronic equipment, is unfortunate because, of course, the purchaser loses confidence when the new television receiver fails soon after installation.

The contribution of faulty valves and components to this overall failure rate curve is shown by the broken lines in Fig. 1. The balance of the "failures" or defects would be due to such items as misalignment, faulty switch contacts, oscillator drift, soldering faults and the like, which are not attributable to particular component failure.

In order to bring about an improvement in the initial reliability of television receivers, there is a

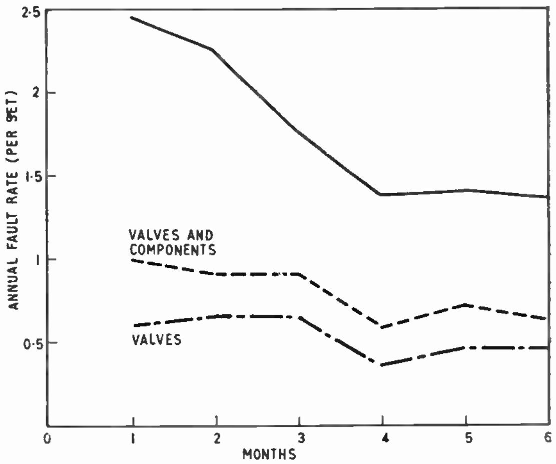

Fig. 1. Failure rate curve for batch of television receivers during first six months of life.

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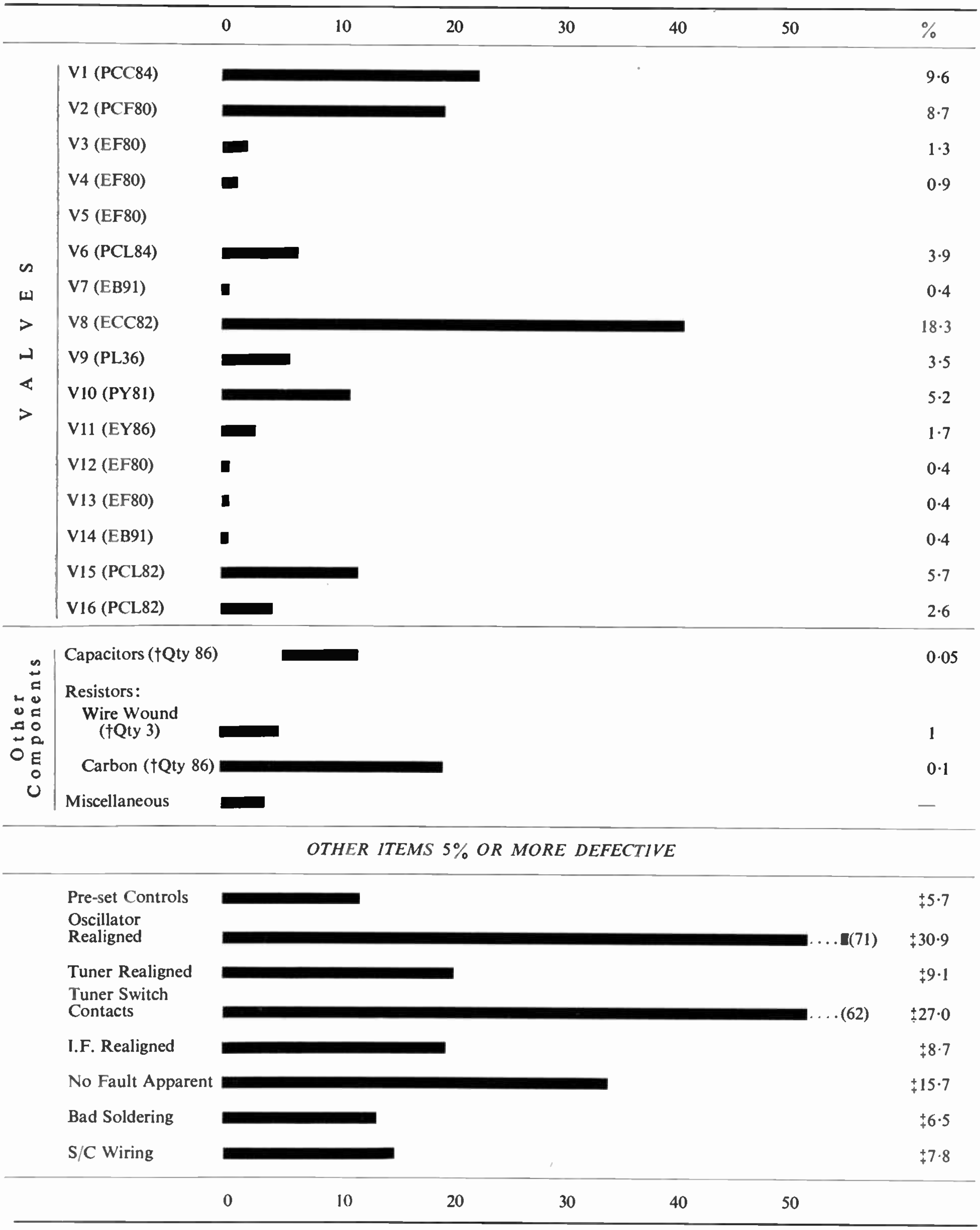
Table 1 : Histogram showing relative failure rate, or defects occurring in the first year of life of a batch of 230 typical receivers.



VALVES	V1 (PCC84)	9·6
	V2 (PCF80)	8·7
	V3 (EF80)	1·3
	V4 (EF80)	0·9
	V5 (EF80)	
	V6 (PCL84)	3·9
	V7 (EB91)	0·4
	V8 (ECC82)	18·3
	V9 (PL36)	3·5
	V10 (PY81)	5·2
	V11 (EY86)	1·7
	V12 (EF80)	0·4
	V13 (EF80)	0·4
	V14 (EB91)	0·4
	V15 (PCL82)	5·7
	V16 (PCL82)	2·6
Other Components	Capacitors (†Qty 86)	0·05
	Resistors:	
	Wire Wound (†Qty 3)	1
	Carbon (†Qty 86)	0·1
	Miscellaneous	—
OTHER ITEMS 5% OR MORE DEFECTIVE		
	Pre-set Controls	‡5·7
	Oscillator Realigned	‡30·9
	Tuner Realigned	‡9·1
	Tuner Switch Contacts	‡27·0
	I.F. Realigned	‡8·7
	No Fault Apparent	‡15·7
	Bad Soldering	‡6·5
	S/C Wiring	‡7·8

† Qty indicates quantity per receiver. ‡ These percentages in right hand column relate to the whole receiver.

field for further investigation on these early "catastrophic" failures in valves and components. Examples of work in this direction with valve makers are quoted later in the paper.

2.2. *Longer Term Failures*

Subsequent to the initial six-months period, the receiver reliability curve settles to a level similar to that at the sixth month, for some two years. After this period longer term failures and defects begin to show up. These are more associated with end of life failures and mechanical items involving wear, such as switches, volume controls and tuning devices.

The histogram of Table 1 depicts the relative failure rate of components and valves, also other defects occurring in the first year of a batch of two hundred and thirty television receivers. The actual percentage failure of the various items is shown in the right-hand column of this figure. For the sake of simplicity, unimportant items causing a few random failures, have been omitted or grouped under "miscellaneous" components. In the following section the "troublesome" items are considered in that order of priority.

3. Reliability of Individual Components and Valves

3.1. *Tuners*

3.1.1. Switches

For reasons of flexibility and best performance, the turret type tuner has been preferred in modern television receivers. It will be noted however from the histogram that tuner switch contacts were, except for oscillator realignment, the most frequent cause of service calls. The contact arrangement used is now well known, i.e. silver plated copper rivet heads as moving contacts and bi-metallic silver and phosphor bronze strip fixed spring contacts. Since these switches are operating on v.h.f. up to 200 Mc/s, any small changes in contact resistance or mechanical alignment have a marked effect on tuner performance, particularly insofar as the oscillator section is concerned. It is particularly desirable to avoid the flow of air, generally contaminated with dust, sulphurous fumes, etc. over the contacts. The formation of silver sulphide on the contacts causes high resistance contacts to develop. Silver oxide formation is not so important as it is not a bad conductor. On the other hand, if the silver plating is not closely controlled and the copper body of the contact head is not properly covered with silver, the formation of copper oxide will soon cause contact troubles.

3.1.2. Wafer switches

Earlier receivers which used the familiar wafer type switches in incremental circuits were found to suffer from switch contact problems to a much higher degree than the turret tuned receivers. One rather expensive cure was the use of solid silver alloy spring contacts and gold plating on the rotors instead of silver plated contacts.

3.1.3. Oscillator stability

A relatively high degree of oscillator frequency stability is called for in the television tuner. Short and long term variations of the frequency, which necessitate realignment of the oscillator inductance, are, it will be noted, the major item for service calls on the batch of receivers under consideration. These particular tuners were fitted with what is known as a "fine tuner", enabling the viewer to make minor adjustments to oscillator frequency, but in 30% of the receivers the variation exceeded the limits over which the correction could be made. With the desirable omission, for reasons of simple operation, of the fine tuner, variation of the order of not more than 50 kc/s at, say, 50 Mc/s (1 part in 1000) over a temperature range of 50 deg C is required (or 1 part in 50 000 per deg C). Where temperature control and other elaborate means are not economically possible, there is an obvious field for improvement in design to reduce existing variation—which may exceed 250 kc/s.

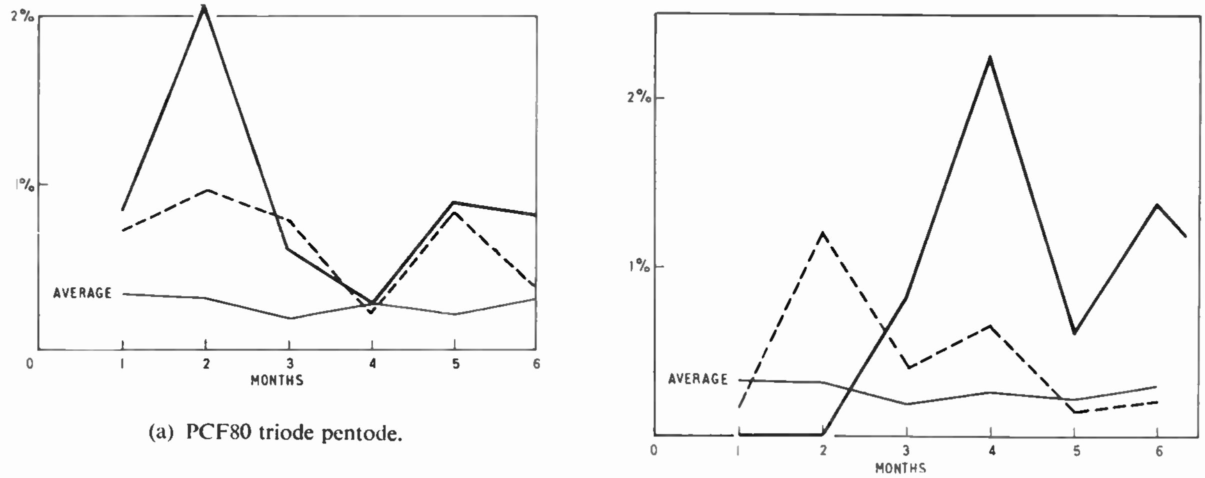
It is normally necessary to use compensating ceramic capacitors to correct for changes in circuit capacitance with temperature. However, it often happens that the production tolerance on the temperature coefficient of the capacitors is such that the frequency drift is still outside the required limits. It is, therefore, desirable in the tuner design to aim at as near as possible overall zero temperature coefficient in the oscillator circuitry without compensation.

In addition to the major items accounting for oscillator frequency variation described above, there can be various other longer term variations due to valve and component changes with time.

3.2. *Valves*

Overall, faulty valves are usually the main reason for service calls, since there are generally fifteen or sixteen used per set, although the unit reliability may not be excessively high. In the receivers considered in Table 1, it will be seen that the particular valves in question had a percentage annual failure varying between 0·4% for EF80 and 18·3% for the ECC82. "Double" valves PCC84, PCF80, PCL84, PCL82 and ECC82 are generally less than half as reliable as single valves—as will be apparent from the histogram.

The graphs given in Fig. 2 show the average failure rate of two particular valve types compared with the average overall valve reliability in receiver batches of total quantity of the order of 600. Following analysis of faulty valves returned to the makers, improved valves were made available for later receiver batches. The degree of improvement resulting can be seen from the broken line curves shown on the same graphs. A comprehensive valve quality control scheme of this nature is being operated with the valve makers—with obviously worthwhile results.



(a) PCF80 triode pentode.

(b) ECC82 double triode.

Fig. 2. Average failure rate for two valve types compared with the average overall valve reliability in a batch of 600 receivers.

3.3. *Valve Pins and Sockets*

A factor calling for a good deal of attention as set life proceeds, particularly in r.f. and i.f. stages, is bad contact between valve pins and sockets. The nickel pins normally used on valves suffer during the heating in manufacture and become oxidized. A test has been conducted in collaboration with the manufacturer in which the valve pins have been chemically cleaned after processing. The nickel then remains bright at normal temperatures and improved results have been obtained in that reduced service calls for faulty contacts have resulted.

In order to obtain overall improvement, it will be necessary for the valve holder makers to study the best plating finishes for the sockets.

3.4. *I.F. Alignment*

Table 1 shows that defects in i.f. alignment accounted for 8·7% of service calls in the batch of receivers under question—this is fairly typical. The reasons for the misalignment have not generally been isolated, but appear to be of a longer term nature rather than short term and are probably associated with changes in valves and components as life proceeds. There is an obvious field for study here. With the coming of printed circuitry and frame grid valves, where circuit inductance and capacitance can be closely controlled, it would appear desirable to design fixed tuned i.f. stages—thus to avoid variations associated with adjustable ferrite cores, etc.

3.5. *Soldering and Wiring Faults*

In the particular batch of receivers under consideration, it will be noted that "bad soldering" and "short circuit" wiring faults between them accounted for 14·3% of faulty receivers. Related to individual joints, of which there are several hundred per receiver, the percentage would, of course, be much lower. However, these faults, which have been with us almost as long as the industry, are still very difficult to eradicate completely—chiefly because of the problem of discovering them in production inspection and test —and are worthy of further detailed study.

3.6. *Printed Panels*

The receivers in question used mainly printed circuitry and probably some 80% of the joints were dip-soldered. It is worth noting that, compared with earlier receivers with conventional wiring, there was a reduction of the order of two to one in the number of soldering and wiring faults. Production inspection of soldered joints on printed panels is found to be simpler and this is probably one reason for the improvement. Components are also now more closely inspected for solderability of end-wires.

Problems associated with burning of printed panels by overheated resistors were covered in an earlier contribution by the author †. Provided the pre-

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† "Printed circuit reliability and flammability", *J.Brit.I.R.E.*, **20**, pp. 281–2, April 1960.

cautions outlined in that note are taken, these very undesirable conflagrations can be avoided.

It is not proposed in this paper to cover principles of good practice in printed panel technique in other respects.

3.7. *Inflammable Impregnants*

This is probably an appropriate point to refer to the desirability of the component industry finding alternative non-inflammable impregnants for such components as capacitors, resistors and inductances. Whilst complete receiver fires do not occur very frequently, such occurrences are to be avoided at all costs. Most of the instances where a complete receiver has been burnt out can be traced to an overheated resistor setting fire to molten wax. Some connecting wire coverings are also sufficiently inflammable to carry the flames from one part of the receiver to another. Once started, such a fire quickly reaches the loudspeaker cone and wooden cabinet.

3.8. *Potentiometers*

There are normally seven or eight potentiometers in a television receiver, all but one or two being of a pre-set variety. The sound volume control is frequently combined with the on/off mains switch both in radio receivers and television receivers. This component generally has the highest failure or defective rate of the several potentiometers used in a receiver.

Apart from early failures, which are generally due to faulty mechanical assembly, the major cause of volume control failure is carbon track wear which results in noisy and erratic operation. From time to time troubles also develop in the mains switch ganged to the volume control. These are mainly (a) weak spring action and causing sluggish operation, which results in burnt contacts, and (b) with double pole switches breakdown of insulation due to narrow clearances and conducting deposits.

3.9. *Resistors*

3.9.1. Carbon

On the whole, and from the histogram of Table 1, it can be considered that carbon solid compound resistors are of satisfactory reliability—provided they are run well within the makers' ratings. These components frequently suffer consequential failure due to short circuits occurring in valves, capacitors or other components. As it was not practical to isolate them in the records under consideration, both primary and consequential failures are combined.

Because these resistors are moisture absorbent it is necessary for them to be covered with a protective coat. This is often a wax impregnant and whilst it stabilizes the resistor characteristics, the associated fire risk is undesirable. Other coatings are available but are generally more costly.

It will also be found that resistors run near maximum rating suffer changes in value over fairly long periods, thus sometimes causing unsuspected changes in receiver performance.

3.9.2. Wire wound

Table 1 shows a failure rate ten times higher for wire wound than carbon resistors. This is fairly typical. Because of their nature, these resistors are inherently less reliable unless there is very close quality control in production. In order to reduce size and cost, extremely fine gauge (generally nickel-chrome) wire is used on ceramic tubes. The making of the end-connections is a critical process. The units are generally coated overall in a binding cement or enamel, vitreous enamel types being the most robust but, unfortunately, also the most costly. Some of the cheaper cement coatings, not often used nowadays, could in the presence of humidity, cause acid action on the fine wire—to the extent that it was severed. This process could occur in storage over a few months, i.e. before the resistor had been used! Most current failures appear to be due to breaks in the windings, the precise cause or causes not having been isolated. Wire impurities and electro-chemical action are probably the main causes.

3.10. *Capacitors*

3.10.1. Electrolytic

Of the various types of capacitor, the electrolytic is the least reliable. There is generally a definite life which is mainly associated with the drying-out of the electrolyte. In normal use this spreads over some three to five years. Earlier failures are generally the result of poor quality control in production. Typical instances are: (a) faulty riveting between electrode connections and terminal lugs; (b) electrolyte seepage due to faulty or blocked vents; (c) corrosive effects on internal connections causing open circuits to develop.

3.10.2. Paper tubular

Provided its use is restricted to relatively low impedance circuits the paper tubular capacitor has reached a satisfactory standard of reliability. Most failures result from moisture ingress and, like resistors, this is a field where the protective coating plays an essential part. The risks associated with wax coatings have already been pointed out. There are nowadays several alternative plastic and other coatings but the author does not have statistics of their reliability at the present time.

3.10.3. Ceramic

Ceramic tubular and disk type capacitors generally have acceptable standards of reliability—provided the peculiar characteristics of these components are taken into account in the receiver design. In particular, the high temperature coefficients of the high permittivity type capacitor should be borne in mind. Over normal temperature ranges capacitance changes of the order of 1·5 to 1 can occur. Some longer term changes in capacitance associated with the silver coating on the ceramic tubes have occasionally been noted. These can be important in tuned circuits. Voltage breakdown occasionally occurs, particularly in relatively cheap and small size feed-through type capacitors.

3.10.4. Silver mica and other types

Silver mica capacitors have generally been found to be most reliable. There are several relatively new types of capacitor using alternative dielectrics such as polystyrene and other films. The indications are that the stability of these capacitors is good, but again the author does not have large scale statistical information on these components.

3.11. *Inductors (Low Frequency)*

Inductors of iron-cored or ferrite types have been generally found to be of acceptable reliability standard—with the possible exception of that critical component the line output transformer. However, even in the case of this component, over the past two or three years, reliability has improved considerably. Most breakdowns were due to insufficient attention by the transformer designers to the simple requirements of high-voltage insulation, particularly of wire coatings, and inter-layer insulation. Since these transformers run with some 15 kV on the outside of the high voltage windings, suitable anti-corona coatings are essential. Many failures have occurred by corona developing through pin-holes in the wax or other "anti-corona tyres" used on these windings. Close quality-control in production is essential.

Deflector coils are also critical components from the point of view of using adequate wire insulation. Peak voltages of the order of 2 kV can exist between wires in fairly close proximity. The use of the newer and thicker wire coatings has brought deflector coils to a good standard of reliability.

Radio-frequency and i.f. inductors present no great reliability problems. Secondary and longer term variations in characteristics with temperature and humidity can cause misalignment problems worthy of future study. F.m. discriminator transformers are particularly critical in this respect.

3.12. *Loudspeakers*

Most troubles associated with loudspeakers result from insufficient clearances between the moving coil and the magnet. Normally the designs are satisfactory but close control in production is necessary in order to maintain centrality between the coil and magnet. With excessively close clearances, the effects of temperature and humidity on the cones are, in service, sufficient to cause slight warping and consequent touching—with objectionable results to the listener!

3.13. *Rectifiers*

Owing to the generally greater reliability, metal (selenium) mains rectifiers have been preferred to valve rectifiers by the author. Provided due precautions are taken in regard to ambient temperatures and cell voltages, these rectifiers can give a high degree of reliability. After an initial and unimportant increase in resistance over the first 500 hours, a steady output may be obtained for several thousand hours.

The silicon rectifier owing to its much smaller size, looks like eventually displacing the selenium type. At present, however, the permissible back voltages of single cells are barely adequate for normal 250 volt a.c. mains use, bearing in mind the superimposed random peaks which can, it is understood, reach 700 or 800 volts occasionally. The use of two cells in series precludes these rectifiers on economic grounds at present.

Germanium diodes, used as detection rectifiers, have been found to give generally good reliability.

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DISCUSSION

Mr. D. H. Busby: The main factors contributing to the reliability of a television receiver are:

(1) The number of components and valves

(2) The maximum dissipation and operating temperatures

(3) Adequate derating for valves and components

(4) Use of circuits tolerant of valve parameters.

The most reliable receiver tends to be the one which has low operating temperatures, uses circuit design tolerant of small changes in valve parameters, and has adequate deratings for components and valves, even though this may be done at the expense of more components.

When one considers the valve complement of a television receiver it is most important to note that the most complex part of the receiver tends to use double valves, for example line and field sweep generators, synch. pulse generators, video amplifiers and tuners. Thus a higher failure rate for these valves may be expected because they are double valves and because they may be used in circuits less tolerant of changes in valve parameters.

Another feature contributing towards field loss in a television receiver is a component failure which often causes excessive valve dissipation. In some receivers for instance if there is a failure in the line oscillator it will result in the line output valve being overloaded and quite probably the final i.f. and video amplifier valves as well. Under some circumstances the set may continue to be used for sound only for some time.

The use of special quality valves is often suggested as a method of reducing valve failure rates in service. Special quality valves however are most often rugged valves which are made mechanically strong and designed to withstand high accelerations, they are not in general better from the point of view of current and dissipation and therefore the use of such valves in an entertainment receiver would not reduce the incidence of valve failures from that point of view.

An average valve failure rate of 0·3% per month is a loss of 3·6% per annum or 3·6 valves per hundred. If we assume that on the average we have 15 valves in a television receiver we see that the average failure rate is about one valve per two receivers per year, or one valve every two years per receiver. Taking into consideration the demands made upon the valves employed in television receivers this does not seem an unreasonably high failure rate.

Early or Short Term Failure Rates

Of recent years considerable work has been undertaken on the investigation of short term failure and they have been found to be mainly attributable to insulation faults or welding faults. A small percentage of welding faults that manage to pass the factory inspection subsequently can come apart on heating and cooling cycles. Insulation faults have always been a problem in valve making as they are due to semi-conductive films being deposited across micas and other insulating parts. The use of a more passive cathode nickel made possible by the more stringent cleaning of components and less risk of contamination during assembly, reduces the risk of barium being deposited on the insulators during the valve life. Insulation faults due to loose particles have been reduced by the adoption of cleaner assembly processes. To reduce the handling and hence the possibility of contamination some success has been achieved with more automated jig assembly and machine assembly of valves. Use of purer materials for the heater/cathode insulation (alundum) has virtually eliminated the possibility of heater-to-cathode shorts. More rigid control of all processing procedures—such as dimensions and tolerances of components has additionally reduced the number of insulation failures.

The work described here has shown a really worthwhile improvement and will no doubt continue to do so into the future, although of course it becomes increasingly difficult to eliminate the last 0·1% of failures for these reasons.

All this work is aimed at achieving a reduction in what is already a reasonably low failure rate and the valve maker is now in a position where his efforts are judged by relatively small changes in an already low figure. It is thus particularly interesting to have Mr. Heightman's valuable observations on the position as he finds it today compared to that of some years ago.

Mr. Heightman (*in reply*): Mr. Busby's interesting contribution fairly explains the valve manufacturer's point of view and gives a good indication of the work which has been done, and is being done, to improve valve reliability.

# of current interest . . .

A Radio Telescope for D.S.I.R.

A fully-steerable radio telescope is to be built at a site near Crowthorne, Berkshire, by the Ministry of Works for the Radio Research Station of the Department of Scientific and Industrial Research. It is expected to be completed and in operation towards the end of 1963 at an estimated overall cost of £250,000 including buildings, equipment and site services.

The radio telescope, which will have a parabolic aerial about 80 ft in diameter, is needed for work in the R.R.S. space research programme and to study certain other aspects of radio science. It will have high accuracy and a faster tracking speed than the 250 ft radio telescope at Jodrell Bank, an essential requirement for following earth satellites and determining orbital data. It will be used to receive weak signals from distant space vehicles transmitting at ultra-high frequencies.

In the field of radio astronomy, one application of the telescope will be a study of radio noise from the sun, including sudden increases in radio emission which are related to disruption in short-wave communications. Radio noise from planets is another phenomenon which may be observed. The telescope will also be generally valuable in extending investigations of the properties of the upper and lower atmospheres which are important in both terrestrial and space communications. This feature will assume even greater significance in the next few years when satellites for communications, as well as for research purposes, are likely to come into use.

Trans-Atlantic Communications Satellite Tests

The British Post Office will co-operate with the U.S. National Aeronautics and Space Administration and the French Centre for Telecommunications Studies in a programme for trans-Atlantic testing of communications satellites. This is jointly announced by the United States, France and the United Kingdom.

Ground stations are to be built in England and France for the reception and transmission of telephone, telegraph and television signals across the Atlantic using satellites to be launched by N.A.S.A. during 1962 and 1963 in Projects RELAY and REBOUND. Stations will be equipped with advanced radio facilities, having extremely accurate tracking and antenna pointing qualities and capable of conducting tests with active and passive satellites at high frequencies and low power.

In Project RELAY, N.A.S.A.'s low altitude active repeater programme the satellite to be launched in 1962 will weigh less than 100 pounds. In addition to equipment for conducting the communications experiments, space-craft will contain instruments to detect radiation damage and other effects of space conditions on critical components. Project REBOUND follows up N.A.S.A.'s first passive reflector communication satellite programme, ECHO. It involves placing several rigidized inflated spheres in orbit by using a single launch vehicle. The first launch to orbit three spheres is scheduled to take place in 1963. RELAY and REBOUND are research and development projects to demonstrate feasibility, basic concepts and technological approaches and to evaluate various systems to be employed in communications satellites.

To an extent permitted by orbital and other technical considerations, it will be possible for other countries so desiring to provide the necessary ground facilities to participate in these co-operative experiments. The British ground station will be at Goonhilly Downs on the Lizard peninsula in Cornwall and will cost about £500,000.

New Television Studios

The Elstree Studio Centre of Associated Television Ltd. was opened recently by the Rt. Hon. Charles Hill, P.C., M.P. This centre will house four studios with a total floor area of 32 000 sq. ft., studio facilities, technical facilities, workshop and administration block. The whole development occupies 340 000 sq. ft. and cost £4,000,000.

The Centre makes use of transistor equipment wherever possible on pulse generation and distribution equipment for video and synchronous circuits and the mixing controls. Much of the equipment has been designed and built by the company's technical departments.

The vision mixer panel provides more facilities than are usually available. The whole mixing operation is effected by the use of germanium diodes and provides inter-frame cutting, i.e. without deformation of succeeding frame. The panel also provides a variety of effects—superimposition, montage, etc.

Provision is made for video tape recording of studio pictures. All the equipment is capable of operation on three line standards: 405, 525 and 625, and conversion from one standard to another can be carried out.

Each studio is provided with five 4½ in. image orthicon cameras. All studio lamps are remotely controlled carried on expandable telescopic supports suspended from the lighting grid. This gives complete freedom of movement in three dimensions and speeds the rigging of lights. Sufficient power capacity has been installed to cover the requirements of colour television.

U.D.C. 621.318.132

High Stability Ferrite Pot Cores

By

W. A. EVERDEN †

Presented at the Symposium on New Components, held in London on 26th–27th October 1960.

Summary: A range of ferrite adjustable pot core assemblies suitable for use in laboratories and for the bulk production of stable inductor units is described. General results on accelerated stability measurements are given.

1. Introduction

When commencing the initial design work on the pot core assemblies to be described in this paper, it was appreciated that many types of assembly were readily available. Few of these, however, were considered to form a complete range designed to give optimum performance when used in filter applications. In addition, existing components did not exploit fully the latest techniques and knowledge within the ferrite industry which were required for modern methods of coil production.

2. The Optimum Pot Core Shape

Neither the scope nor length of the paper allows a complete discussion of the many parameters which influence the introduction of such a range of components. However, as the majority of pot cores used are for applications below 100 kc/s, it was decided to base the calculations leading to the establishment of optimum core proportions on the following:

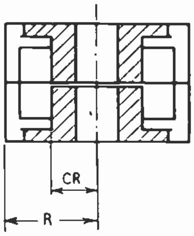
(*a*) The minimum ratio of d.c. resistance R_{cu} to inductance L for a given effective permeability, rationalized to be independent of volume.

(*b*) The maximum ratio of hysteresis volume to overall volume consistent with (*a*).

These assumptions are justified if it is remembered that for low level applications the hysteresis loss is small and the residual core loss at the lower frequencies is also small compared with that due to d.c. resistance R_{cu}. At such frequencies R_{cu} is usually equal to or greater than 50% of the total coil loss.

To maintain the requirements given under (*a*), the ideal core proportions do not include a hole in the centre core suitable for an adjusting mechanism, although from calculations made the diameter is only increased slightly by a hole of typical size. The respective proportions are shown in Fig. 1.

While the requirements under (*b*) indicate that hysteresis loss is of secondary importance, it is probable that the larger cores will be used under conditions of increased level making it desirable to increase the hysteresis volume. A substantial increase can be achieved at little expense to the low level performance by increasing the centre core diameter from optimum until R_{cu}/L has increased by approximately 1%.



	Without centre hole	With typical centre hole
C	0·386	0·445

Fig. 1. Ratio of centre core diameter with and without hole suitable for adjusting mechanism.

Using the above ideas to obtain optimum performance, proportions can be calculated which are general, and any number of sizes within a range can be assumed. If the outside diameter is treated as a basic parameter the theoretical optimum height and internal configuration may be calculated. Dimensions for the Vinkor range are shown in Fig. 2.

3. Effective Permeability

The permeability of an assembly contributes much to its final performance and it is essential that the value be chosen with a complete knowledge of its effect on other parameters. At low frequencies there are advantages in making these values as high as compatible with any temperature coefficient or hysteresis requirements in order to obtain the lowest ratio of resistance R_{cu} to inductance L.

From the general expression for temperature coefficient of inductance $\frac{\mu_e \mathrm{d}\mu_i}{\mu_i^2 \mathrm{d}T}$ per deg C it can be seen that the ultimate value is proportional to μ_e since the

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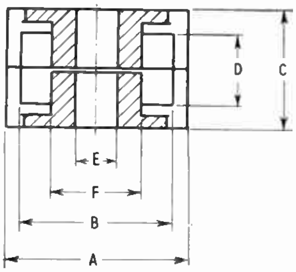
† Mullard Ltd., Torrington Place, London, W.C.1.



Nominal size mm	A mm	B mm	C mm	D mm	E mm	F mm
14/9	14·0	11·6	9·6	3·6	3·6	6·1
16/10	16·0	13·2	10·2	6·6	3·6	6·8
18/11	18·0	15·1	11·2	7·4	4·6	8·0
21/13	21·5	18·0	13·6	8·8	4·6	9·7
25/16	25·4	21·4	16·0	10·4	5·4	11·4
30/18	29·5	24·3	18·8	12·2	5·4	13·8
35/23	35·5	29·2	22·8	14·8	5·4	16·3
45/29	45·0	37·2	29·2	19·0	5·4	20·3

Fig. 2. Dimensions for Vinkor range of pot core assemblies.

term $\frac{d\mu_i}{\mu_i^2 dT}$ is independent of core shape and effective permeability μ_e and hence a constant for any given material.

As polystyrene capacitors which have a temperature coefficient of approximately -60 to -200 parts in 10^6 per deg C, are normally associated with ferrites in tuned circuits it is desirable that a range of effective permeabilities be chosen so that the composite temperature coefficient approaches zero.

Using ferrites from the Mn Zn group, examples of which are shown in Table 1, this requirement is not always compatible with keeping the core loss to a minimum. However, a logarithmic series enables the coil designer to obtain either high Q or a lower value of temperature coefficient dependent upon his application.

3.1. *Choice of Effective Permeability*

The series chosen gives primary standards of 160, 100, 63, 40 and 25 in which the increments are sufficiently small to provide a range of temperature coefficients which cannot economically be improved by intermediate values of permeability.

Actual values of effective permeability are maintained within close limits by mechanically introducing a small air-gap into the cores and controlling the resultant value of turns/mH$-\alpha$. Hence the designer may calculate the number of turns required for a given inductance, and then wind to this number of turns with the knowledge that the inductance will be within the stated accuracy, which in most cases is better than $\pm 3\%$.

4. Adjusting Mechanism

A review of the various forms of adjusting mechanisms available will indicate that a great deal of thought has been given to this subject.[1, 2, 3] However, most of the forms in common use do not constitute the ideal mechanism and usually suffer from one or more of the following imperfections:

(1) Non-linearity of adjustment.
(2) Coarse adjustment.
(3) Non-cyclic instability.
(4) Local high induction resulting in increased losses.
(5) Difficulties of manufacture.

Most of these difficulties are overcome by using a method of shunting the air-gap, the most satisfactory results being obtained with a ferromagnetic slug moving axially and partially filling the hole in the centre core.

Table 1

(*a*) General Material Properties of the two grades of Manganese Zinc Ferrite used for the Pot Cores.

Property	Conditions of test	Limit values	
		Frequency range < 100 kc/s	Frequency range 100 kc/s–2 Mc/s
Initial Permeability μ_i .	$\leqslant$ 0·5 gauss	$\geqslant$ 1150	$\geqslant$ 600
Temperature Factor $\frac{d\mu_i}{\mu_i^2 dT}$	$\leqslant$ 0·5 gauss	$\leqslant 2 \times 10^{-6}$ deg^{-1} C	$\leqslant 2{\cdot}5 \times 10^{-6}$ deg^{-1} C
Curie Point . . .	$\leqslant$ 0·5 gauss	$\geqslant$ 150° C	$\geqslant$ 150° C
Hysteresis C_h 4 kc/s 100 kc/s .	$\leqslant$ 20 gauss	$\leqslant 0{\cdot}97 \times 10^{-6}$	$\leqslant 1{\cdot}16 \times 10^{-6}$
Magnetic Induction . .	H = 10 Oe	$\geqslant$ 3400 gauss	—
D.C. Resistivity . .	—	$\geqslant$ 20 Ω cm	$\geqslant$ 150 Ω cm
Loss Factor $\frac{\tan\delta}{\mu_i}$			
4 kc/s . . .		$\leqslant 2{\cdot}5 \times 10^{-6}$	
30 kc/s . . .	$\leqslant$ 0·5 gauss	$\leqslant 6{\cdot}0 \times 10^{-6}$	
100 kc/s . . .			$\leqslant 10 \times 10^{-6}$
500 kc/s . . .			$\leqslant 14 \times 10^{-6}$
1000 kc/s . . .			$\leqslant 32 \times 10^{-6}$
Disaccommodation . .	—	$\leqslant$ 2·5%	$\leqslant$ 2·5%

All values refer to measurements on standard toroids at 20° C ± 5° C

(*b*) Typical Values of Mechanical Properties

Specific gravity	4·8
Tensile strength	2600 lb/in^2
Crushing strength	10 000 lb/in^2
Young's modulus	24×10^6 lb/in^2
Coefficient of linear expansion	10×10^{-6} deg^{-1} C
Thermal conductivity	0·008 cal/s/cm deg^{-1} C
Specific heat	0·17 cal/gm deg^{-1} C

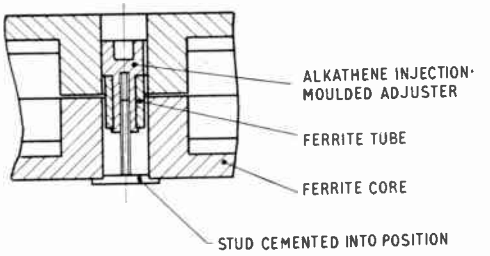


Fig. 3. Sectionalized view of adjusting mechanism.

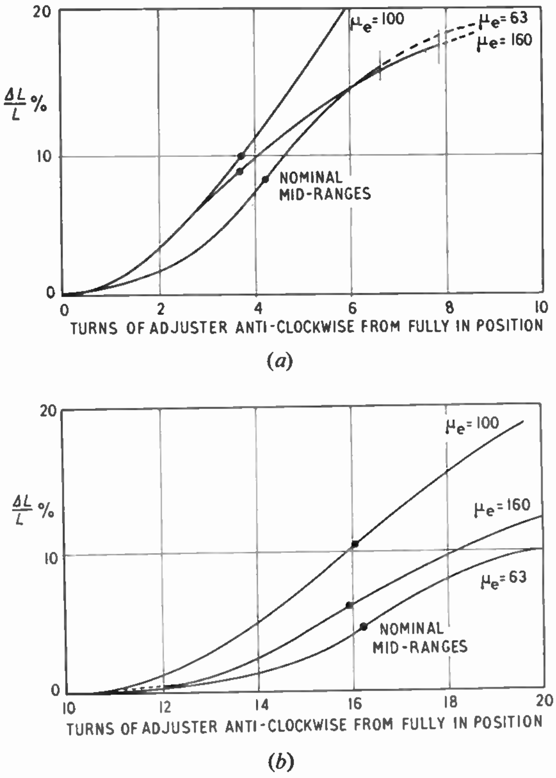


Fig. 4. Typical adjustment curves
(a) 18 mm Vinkors
(b) 35 mm Vinkors

Figure 3 shows the general form chosen. The central stud accurately locates the mechanism whilst the five ribs around the periphery eliminate possible wobble and the resultant non-linearity of adjustment. The fine thread form allows a setting accuracy of better than 0·02% to be obtained whilst the complete adjusting mechanism is substantially contained within the external dimensions of the cup cores and is completely independent of any additional clamping arrangements.

4.1. *Range of Adjustment*

The stability of a complete inductor may be influenced by the amount of adjustment which should therefore be kept to a minimum consistent with the overall requirements of filter construction. A range of approximately 14% irrespective of effective permeability is considered adequate, made up as follows: ±3% or 6% variation for the cancellation of spread in effective permeability of cores, ±2% or 4% for capacitor tolerances, and approximately ±2% or 4% variation for final trimming of the complete network. Two sets of typical adjusting curves are shown in Fig. 4.

5. Housings

An exploded view of a typical housing suitable for either bulk production or Laboratory use is shown in Fig. 5. This type of unit has been designed to overcome many of the disadvantages of previous assemblies and includes such features as the conical spring washer, which when depressed by the fixing bush, applies a predetermined pressure to the cup cores and holds them firmly together without the additional use of an adhesive. The accessibility of the adjusting mechanism makes the assembly equally suitable for impregnation or resin casting prior to a final adjusting operation.

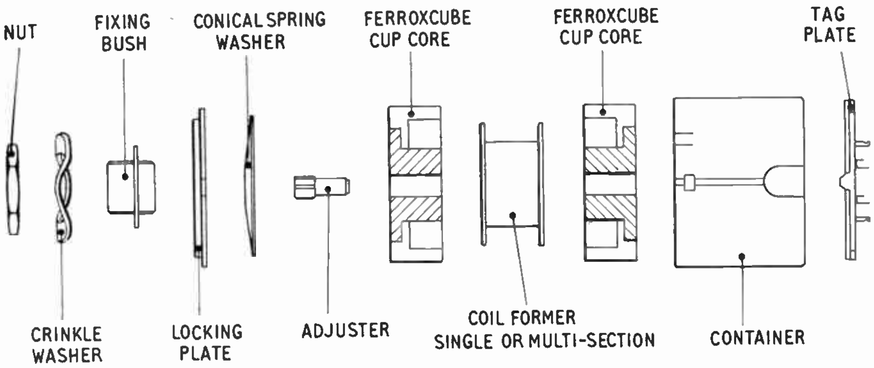


Fig. 5. Exploded view showing complete housing.

6. Evaluation of the Design

The practicable realization of such a range of components produces many difficulties not the least of which are the tight mechanical limits required on the individual cup cores. Ferrites with their inherent high shrinkage during manufacture make it necessary to inspect 100% of cores, while for maximum linearity of adjustment, it is essential to ream the centre hole of a pair of cups to maintain the cylindrical tolerances.

Production-built filters have proved that the features described have been achieved in the Vinkor series. Filter coils have been wound and adjusted in approximately one-third of the time taken for an equivalent coil on the older fixed type of assembly. The bulk of the saving is in the reduction of operator's time when adjusting the final value of inductance.

7. Thermal Stability

The tests carried out were designed to investigate the inductance stability of such components when subjected to daily temperature cycling. Random production samples of 25 mm pot cores with effective permeabilities of 100, were selected for these measurements.

The coils were designed from the information given in Table 2 and after assembly, were adjusted to give their design value of inductance. Vacuum impregnation was carried out on two-thirds of the coils, Okerin 561 wax (m.p. 83° C) being used for the assemblies with Ralsin (low water absorption nylon) coil formers. For those with polystyrene coil formers Okerin 4148 wax (m.p. 75° C) was employed. In the remaining dry assemblies the coil formers were fixed

Table 2(a)

Electrical and Magnetic Data for Pot Cores suitable for use up to frequencies of 200 kc/s

Nominal diameter of core assembly (mm)	μ_e	α	Approx. frequency range (kc/s)	Typical values of $\tan\delta_{r+e}$		Hysteresis Factor measured at 4 kc/s F_H (maximum)	Temperature Coefficient of inductance over 20–50° C (parts in 10^6 per deg C)	μ_e without adjuster	$\Sigma l/A$	$\Sigma l/A^2$	V_{eff}	L_{eff}	A_{eff}
				30 kc/s	100 kc/s								
18	160	52·5	Up to 100	0·576 × 10^{-3}	1·104 × 10^{-3}	35·8	0 to 320	146·5 ± 3%	5·6 cm^{-1}	12·6 cm^{-3}	1·1 cm^3	2·48 cm	0·44 cm^2
	100	66·4	Up to 100	0·360 × 10^{-3}	0·690 × 10^{-3}	17·7	0 to 200	90·0 ± 3%					
	63	83·6	Up to 200	0·226 × 10^{-3}	0·435 × 10^{-3}	8·84	0 to 126	58·0 ± 2%					
21	160	46·1	Up to 70	0·590 × 10^{-3}	1·120 × 10^{-3}	25·1	0 to 320	149·5 ± 3%	4·25 cm^{-1}	5·87 cm^{-3}	2·22 cm^3	3·07 cm	0·72 cm^2
	100	58·3	Up to 70	0·370 × 10^{-3}	0·700 × 10^{-3}	12·4	0 to 200	89·4 ± 3%					
	63	73·4	Up to 200	0·233 × 10^{-3}	0·440 × 10^{-3}	6·2	0 to 126	59·0 ± 2%					
25	160	42·5	Up to 40	0·600 × 10^{-3}	1·230 × 10^{-3}	19·6	0 to 320	146·5 ± 3%	3·64 cm^{-1}	3·64 cm^{-3}	3·64 cm^3	3·64 cm	1·00 cm^2
	100	53·8	Up to 70	0·375 × 10^{-3}	0·770 × 10^{-3}	9·7	0 to 200	90·3 ± 3%					
	63	67·8	Up to 200	0·236 × 10^{-3}	0·485 × 10^{-3}	4·85	0 to 126	59·0 ± 2%					
30	160	37·5	Up to 30	0·800 × 10^{-3}	1·760 × 10^{-3}	14·75	0 to 320	148·3 ± 3%	2·82 cm^{-1}	1·85 cm^{-3}	6·57 cm^3	4·31 cm	1·53 cm^2
	100	47·5	Up to 70	0·500 × 10^{-3}	1·100 × 10^{-3}	7·3	0 to 200	90·0 ± 3%					
	63	59·8	Up to 200	0·315 × 10^{-3}	0·693 × 10^{-3}	3·65	0 to 126	59·5 ± 2%					
35	160	34·3	Up to 20	0·880 × 10^{-3}	2·048 × 10^{-3}	10·92	0 to 320	150·4 ± 3%	2·35 cm^{-1}	1·06 cm^{-3}	11·7 cm^3	5·25 cm	2·23 cm^2
	100	43·4	Up to 50	0·550 × 10^{-3}	1·280 × 10^{-3}	5·4	0 to 200	89·7 ± 3%					
	63	54·7	Up to 200	0·346 × 10^{-3}	0·850 × 10^{-3}	2·7	0 to 126	60·1 ± 2%					

Table 2(b)

Electrical and Magnetic Data for Pot Cores suitable for use over the frequency range 100 kc/s—3 Mc/s

Nominal diameter of core assembly (mm)	μ_e	α	Approx. frequency range (kc/s)	Limit values of $\tan\delta_{r+e}$		Hysteresis Factor measured at 100 kc/s F_H (maximum)	Temperature Coefficient of inductance over 20–50° C (parts in 10^6 per deg C)	μ_e without adjuster	$\Sigma l/A$	$\Sigma l/A^2$	V_{eff}	L_{eff}	A_{eff}
				100 kc/s	1000 kc/s								
18	63	83·6	Up to 700	0·55 × 10^{-3}	1·75 × 10^{-3}	8·9	0 to 158	58 ± 2%	5·6 cm^{-1}	12·6 cm^{-3}	1·1 cm^3	2·48 cm	0·44 cm^2
	40	105	Up to 2000	0·35 × 10^{-3}	1·1 × 10^{-3}	4·5	0 to 100	35·9 ± 2%					
	25	142	Up to 3000	0·23 × 10^{-3}	0·71 × 10^{-3}	2·3	0 to 62·5	22·1 ± 3%					
21	63	73·4	Up to 2000	0·55 × 10^{-3}	1·75 × 10^{-3}	6·2	0 to 158	59 ± 2%	4·25 cm^{-1}	5·87 cm^{-3}	2·22 cm^3	3·07 cm	0·72 cm^2
	40	92·0	Up to 2000	0·35 × 10^{-3}	1·1 × 10^{-3}	3·2	0 to 100	35·9 ± 2%					
	25	123·0	Up to 3000	0·22 × 10^{-3}	0·71 × 10^{-3}	1·6	0 to 62·5	22·6 ± 3%					
25	63	67·8	Up to 700	0·55 × 10^{-3}	1·75 × 10^{-3}	4·9	0 to 158	59 ± 2%	3·64 cm^{-1}	3·64 cm^{-3}	3·64 cm^3	3·64 cm	1·00 cm^2
	40	85·1	Up to 1200	0·35 × 10^{-3}	1·1 × 10^{-3}	2·5	0 to 100	35·7 ± 2%					
	25	115·0	Up to 3000	0·23 × 10^{-3}	0·71 × 10^{-3}	1·3	0 to 62·5	22·1 ± 3%					

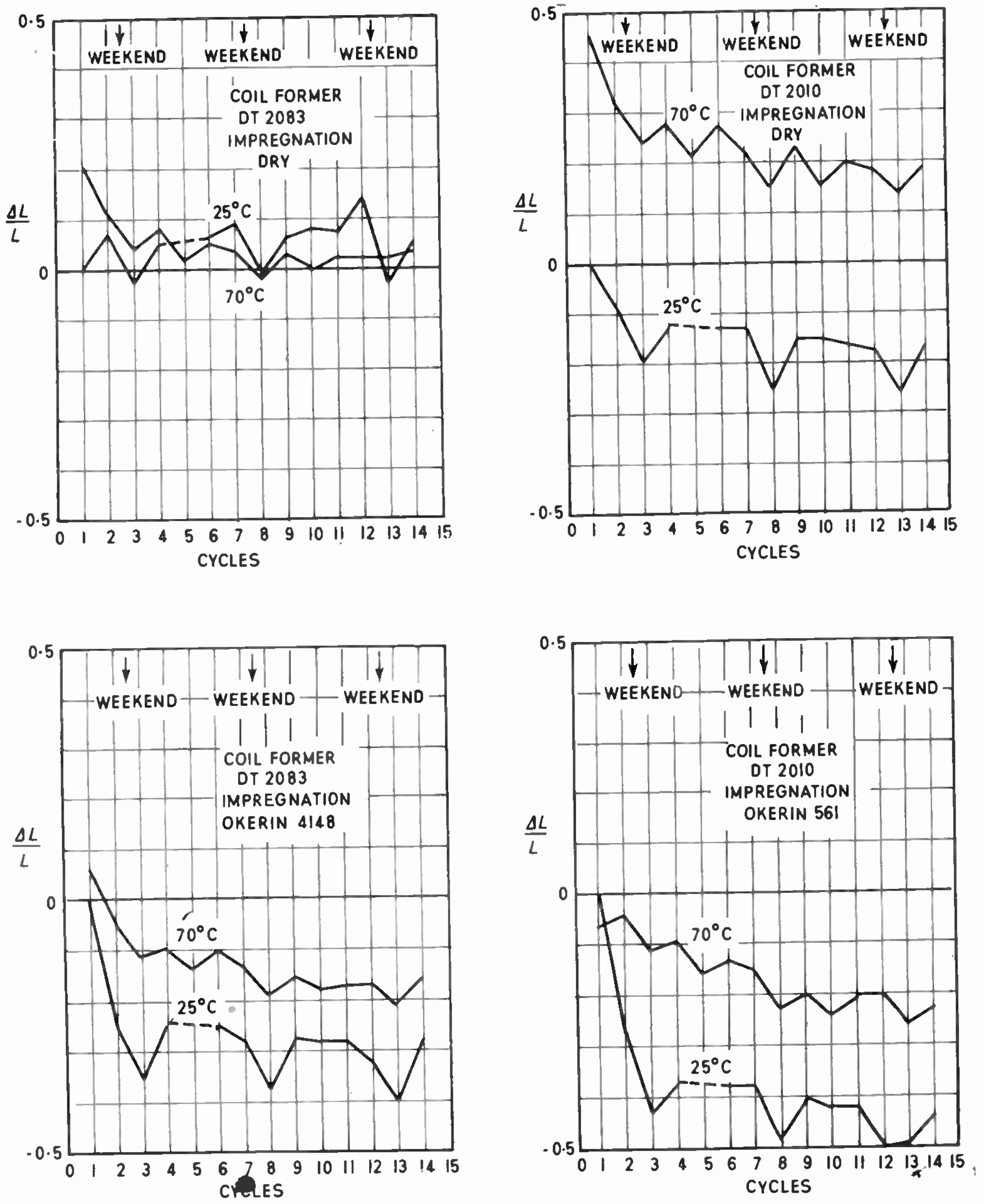


Fig. 6. The effect of temperature on the stability of 25 mm pot cores ($\mu_e = 100$).
Note: Coil former materials are polystyrene (DT 2083) or nylon (DT 2010).

to one half of the core with an epoxy type resin.

The cores were cycled between 25° C and 70° C each day, and allowed to return to 25° C overnight ready for the following day's cycle. During weekends the temperature remained constant at 25° C. Inductance measurements were made at the same times, twice daily, the flux density in the centre core of each assembly being kept well below 5 gauss.

The results of these measurements are shown graphically in Fig. 6. The 25° C curves show clearly the now familiar thermal disaccommodation effect of the core material which arises when there is a sudden change in the rhythm of cycling. In general the curves show no apparent instability which can be attributed to the housing arrangement or adjusting mechanism.

8. Stability when Subjected to Vibration

A set of 25 mm pot cores similar to those used in the thermal stability measurements were subjected to varying modes of vibration. All coils used were first bonded to one half of the ferrite and wax impregnated

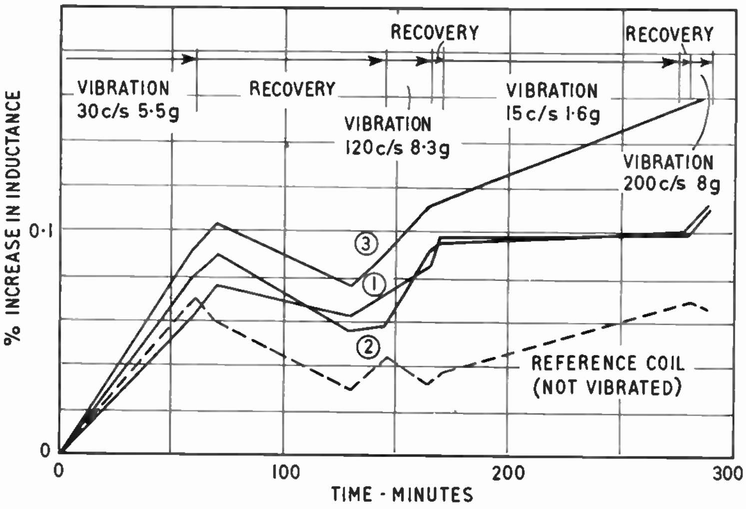


Fig. 7. Stability when subjected to vibration.

to reduce any effects due to turns slipping during the measurements.

Initial tests were carried out in accordance with B.S. 2011. The general result was that the inductance had increased by less than 0·01% indicating a high degree of mechanical stability.

A second series of tests (details shown in Fig. 7) were made on the same assemblies after having disdisturbed the adjuster each time, to simulate initial adjustment conditions. The assemblies were divided into three groups and mounted on the working platform of an electromagnetic vibrator in the manner shown below, and vibrated with a sinusoidal motion:

(1) Axis perpendicular to direction of vibration.
(2) Axis parallel to direction of vibration (tag plate facing down).
(3) Axis parallel to direction of vibration (tag plate uppermost).

The results of these measurements are shown graphically in Fig. 7 and also indicate a similar degree of stability. A resonance search did not reveal any frequency at which the assemblies or adjuster mechanism resonated.

9. Acknowledgments

Acknowledgments are due to Messrs. E. C. Snelling and N. C. O. Jackson who carried out the original design work, and Mr. T. Longhurst who carried out all measurements upon which this paper is based. The author wishes to thank the directors of Mullard Ltd. for permission to present the paper.

10. References

1. R. S. Duncan and H. A. Stone, "A survey of the application of ferrites to inductor design", *Proc. Inst. Radio Engrs*, 44, pp. 4–13, January 1956.
2. H. A. Stone, "Ferrite cored inductors", *Bell Syst. Tech. J.*, 32, pp. 265–91, March 1953.
3. *Matronics Technical Information Bulletin*, No. 6, May 1954, (N. V. Philips Gloeilampenfabrieken).

11. List of Symbols

μ_e	Effective permeability of a gapped core.
μ_i	Initial permeability.
α	Number of turns for one millihenry.
$\tan\delta_{r+e}$	Residual plus eddy current dissipation factor.
$(\tan\delta)/\mu_i$	Total loss factor.
R_{cu}	D.c. resistance of winding.
L	Inductance in henries unless otherwise stated.
Q	Quality factor.
H	Magnetizing force in oersteds.
A_e	Effective core area $= \dfrac{\Sigma l/A}{\Sigma l/A^2}\text{cm}^2$
l_e	Effective core magnetic path length $= \dfrac{(\Sigma l/A)^2}{(\Sigma l/A^2)}$ cm.
V_e	Effective core volume $= \dfrac{(\Sigma l/A)^3}{(\Sigma l/A^2)^2}$ cm^3
A	Cross-sectional area of magnetic core in cm^2.
l	Length of flux path in cm.
F_h	Hysteresis factor.

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An Experimental Magnetic "Talking Book" Machine

By

C. W. ROSS

(*Associate Member*) †

Summary: A slow speed ($1\frac{7}{8}$ in/s) magnetic tape reproducer is described which uses endless loop cassettes in the form of magazines, inserted in a slot in the machine case. The playing time is 96 minutes using four tracks on $\frac{1}{4}$ in. tape. Two-sided tape may be used. The machine is primarily intended for use by blind people and simplicity of operation has been emphasized. This particular machine was designed to reproduce speech only, and the tape transport mechanism and magazine construction are very straightforward and easily manufactured. Methods of producing the loop copies are discussed.

1. Introduction

Although there are now a great many magnetic tape recorders on the market, few manufacturers have produced so far a semi-automatic machine using cassettes of magnetic tape where the actual handling of the tape by hand is not required. Probably the simplest concept of such a device would be a flat square cassette, somewhat similar to a small book, which could be pushed into a "letterbox" opening in the machine case.

For sightless persons, for instance, simplicity of operation is essential, and in the past disc records have been used because of their ease of handling when used with specially adapted record players. The records are fairly bulky, however, and have a comparatively short life compared with magnetic tape which can be entirely handled by the machine during its useful life.

Another aspect to be borne in mind when considering "talking books" is the complexity of the copy-making apparatus which governs the time taken to make a batch of copies. Should a two-spool system be chosen, re-winding all the copies of the "book" periodically prior to re-issue must be taken into account. Re-recording can possibly be simplified by placing the master recording in reverse and "dubbing" in that manner to produce ready re-wound copies to be played in the usual manner. Also, the tape may be made with a very large number of tracks recorded upon it, but such a system, although attractive at first sight, can be costly and cumbersome.

The alternative to the twin spool device is the endless loop cassette ‡ which is a single spool of special dry-lubricated tape, where the inner portion is pulled out and joined to the outermost strip, thus forming in effect a very long loop of tape. No driving mechanism is then needed for the cassette reel, only a simple capstan and roller drive for the portion of the tape between the centre and the outside of the reel (Fig. 1).

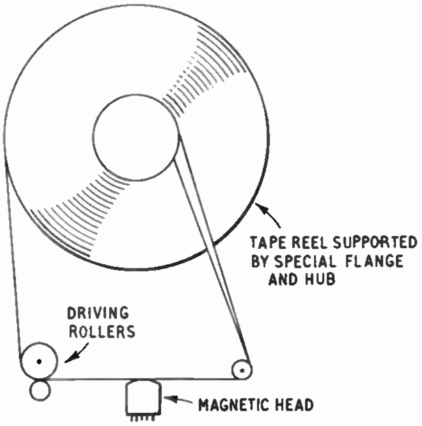

Fig. 1. Endless loop system.

Up to four tracks have been found satisfactory with the standard $\frac{1}{4}$ in. width tape. There is, however, a maximum limit to the length of tape which can be used in this manner, unlike the twin spool system, of the order of 500 ft. Both sides of the tape may be recorded, subject to certain conditions which will be described later. The velocity of the tape may be reduced to a minimum, say, $\frac{15}{16}$ in/s, for a frequency response to cover the speech spectrum satisfactorily. This would give a total playing time per cassette of around 14 hours, but a great deal of further development work will be required in the design of tape and cassette, and recording apparatus to achieve this aim. So far only one side of the tape has been used and the velocity is higher, $1\frac{7}{8}$ in/s, and the cassette is a 225 ft version. These factors therefore reduce the total playing time to around 96 minutes.

The apparatus for making the copies is fairly straightforward and the process is not unduly time-consuming.

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† Kelvin & Hughes Ltd., Hillington, Glasgow, S.W.2.

‡ B. A. Cousino and R. E. Cousino, "A continuous loop magnetic tape cartridge". *J. Audio Engng Soc.*, **6**, pp. 49–57, January 1958.

2. The Basic Requirements

2.1. *Playing Time*

Allowing for the characteristics of the magnetic tape available for use in an endless loop cassette of small size, the period chosen was a total playing time of between two and six hours, or more if possible. The machine became known as the "talking book" but it was suggested that the contents of the "book" should be of magazine nature as a preliminary step. With a view to starting a pilot scheme the subjects first chosen for the pre-recorded tape would be local news, short stories and items of general interest.

The tapes would then be recorded, at set periods at a "central tape library" on special machines built to handle perhaps ten cassettes at a time, and the recordings produced at four times the reproducer speed at least.

The playing time is obviously dependent on the length of tape available and the tape velocity. Since the highest frequency which can be reproduced is a direct function of the tape speed for a given head and tape, the latter is thus fixed by the frequency response considered desirable. This can only be assessed on a subjective basis, being a compromise between intelligibility and listener fatigue. Generally the more pleasant the sound the longer it can be tolerated.

2.2. *Frequency Response*

It was found that a band coverage from 250 c/s to 3500 c/s was the minimum for listening for long periods without undue fatigue. The magnetic head on the reproducer is a four-channel type having high resolution, the effective magnetic gap length being about 0·0002 in. When such a multi-channel head is used, crosstalk is an important factor, and a crosstalk ratio of 30 dB was established as sufficient for speech. Use was made of the loudspeaker resonance to improve the lower part of the reproduced speech spectrum. The replay amplifier was not equalized but the recording contained the necessary boost at high frequencies to maintain a reasonable response on replay up to 3500–4000 c/s.

2.3. *Tape Velocity*

Partly due to the characteristics of the tape the frequency response was too limited at $\frac{15}{16}$ in/s for comfortable listening for long periods and 1·4 in/s was tried. This was found satisfactory, but it is of course a non-standard velocity. The machine was run at $1\frac{7}{8}$ in/s thereafter and the performance was very good.

There are two cassettes which could be used with this machine, a 500 ft cassette and a 225 ft version. At present, the 225 ft type only is readily available in this country and therefore the machine illustrated used this type. This gives the total playing time at the velocity finally chosen of 96 minutes.

The cassette uses a dry-lubricated double-coated tape of the standard $\frac{1}{4}$ in. width and the time quoted is for one side of the tape recorded. Provided the long wavelengths are limited to about 0·005 in. maximum, the tape can be continuously recorded on *both* coated sides, doubling the playing time to 192 minutes per cassette. To do this the loop is cut at the join and reverse spliced to bring each side alternately past the magnetic head (their configuration is known as a "Mobius loop"). The crosstalk due to this arrangement is proportional to the wavelength existing on the tape, and taking into consideration the variation in sensitivity of the ear to sounds of differing frequencies but similar intensity,[†] the figure of 0·005 in. maximum wavelength was found acceptable. At this level the output did not produce any worse effects than the crosstalk already existing in the head due to the close proximity of the magnetic circuits.

An alternative course would be for a single channel head to be moved mechanically across the width of the tape to overcome the crosstalk within the present head.

3. General Description

Figure 2 shows the form of the prototype machine. This uses valves and a mains-operated motor. The magazine or "cassette" is also shown. Figure 3 is a view of the rear of the machine with the cover removed, in which the magazine is shown in position.

To operate the machine, the magazine is pushed home and clicks into place, where it is retained by the clip lever visible just above the aperture which takes the magazine. This operates a micro-switch in the

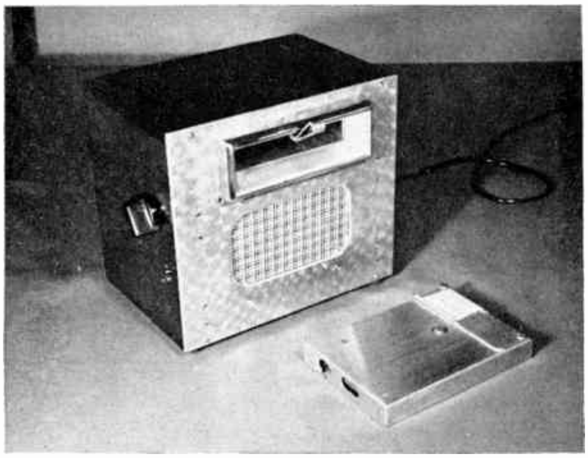

Fig. 2. General view of the machine.

[†] H. Fletcher and W. A. Munson, "Loudness, its definition, measurement and calculation" *Bell Syst. Tech. J.*, **12**, No. 4, p. 377–430, October 1933.

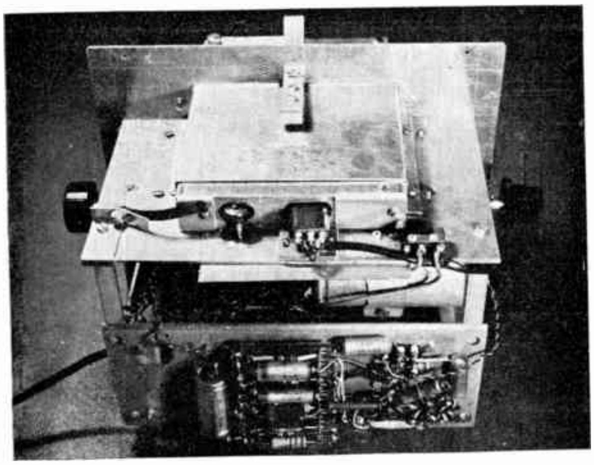

Fig. 3. Rear view of the machine.

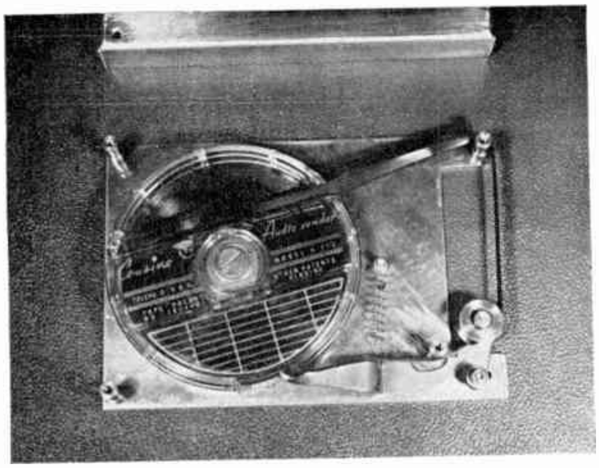

Fig. 4. Cassette removed from its case.

motor circuit and the machine starts normally. The pressure roller and capstan are brought into contact by this operation and the "auto-brake" on the tape is released by the movement of the pressure roller arm. This can be seen clearly in Fig. 4 which shows the magazine with its cover removed.

The tape registration is governed by the guide preceding the magnetic head and the flanged pressure roller which is attached to the spring-loaded brake arm. There are two apertures at the end of the cassette, one for the head—$\frac{3}{8}$ in. × $\frac{7}{8}$ in.—and the other to accommodate the partial entry of the capstan—$\frac{3}{8}$ in. × $\frac{5}{8}$ in.

The capstan is directly driven through a gear train from a small split-phase 6V induction motor supplied from the heater winding on the transformer. At first glance the prospect of direct gear drive would appear to lead to poor replay quality, but by using a skewed rotor and fibre gears, tooth flutter, etc. is hardly noticeable at all on speech. This system provides a really reliable drive of long life, without maintenance, whereas the rubber-tyred reduction drive with flywheel is rather cumbersome for this type of machine and requires periodical maintenance.

The machine is stopped by lifting the clip lever which releases the cassette or magazine from the drive, and the brake is applied inside the magazine. Operation of the machine therefore is very simple, merely pushing the magazine home to start and pressing the clip to stop the machine. There are two controls, shown at the sides of the machine, respectively for channel selection and volume. The usual procedure when operating the machine is to select the next channel after each convolution of the tape to provide one long programme. Alternatively, any channel may be selected at any time if there are separate items of interest or one channel can be repeated continuously, etc. Admittedly channel selection could be made fully automatic, but the slight advantage gained thereby hardly justifies the considerable added complication. Below the volume control there is an output socket for headphones.

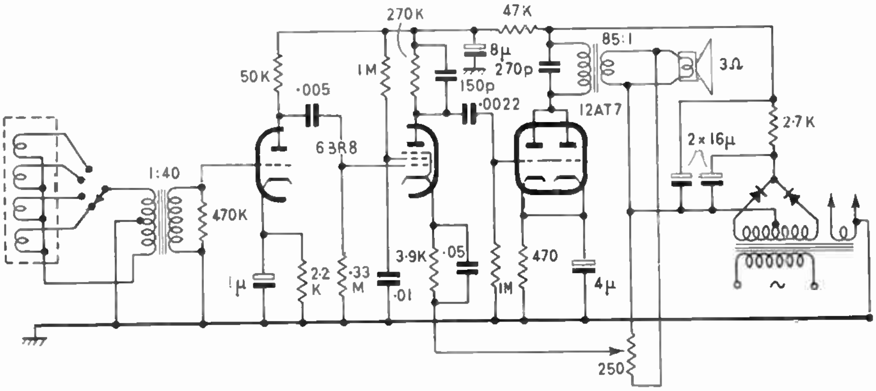

Fig. 5. Circuit diagram of the amplifier.

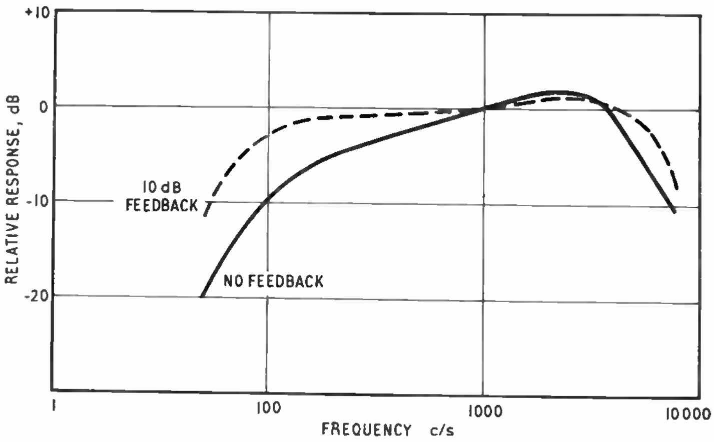

Fig. 6. Static frequency response of "talking book" amplifier injecting across 1 Ω resistor in series with head.

The four-section, in-line magnetic head (9 mH per section) is transformer coupled to a two-stage high gain amplifier shown in Fig. 5. The high gain of a resistance-capacitance coupled 6BR8 triode-pentode is utilized to bring the signal voltage at the transformer secondary up to a value large enough to drive a parallel-connected 12AT7 double-triode acting as an output stage to feed the loudspeaker. Overall gain is controlled by varying the feedback from the output transformer secondary, by means of the 250 Ω potentiometer shown. This arrangement ensures that the output of the machine cannot be reduced to zero. Therefore the machine is unlikely to be left running unnoticed with volume control at minimum. The amplifier is fully loaded with 4 mV on the first stage control grid, and the maximum output is 200 mW into a 4 in. × 7 in. elliptical speaker with a 10 500 gauss magnet which gives a high acoustic level for 200 mW electrical power output. The static response of the amplifier for two conditions is shown in Fig. 6; the response includes the magnetic head and transformer combination and the signal was injected by inserting a 1-ohm resistor in the head circuit. The response is shown for two conditions, zero negative feedback and 10 dB negative feedback. Figure 7 shows the input/output curves of the amplifier for the two conditions specified, and the start of the non-linearity shows clearly the overload point of the output stage.

The power supply for the amplifier is derived from a small "C" core mains transformer which has the advantage of a fairly low external 50 c/s field, reducing hum pick-up in the multi-section head. Small metal rectifiers connected "push–pull" and smoothed by an R-C filter supply the h.t. at 300V to the amplifier stages.

The final drive speed to the capstan is 68 rev/min and the period taken from "switch on" to full speed is quite short due to the drive system presenting a very light load to the motor. This ensures that the possibility of a word being missed during starting or stopping time is remote.

All the recorded tapes that were used with the machine during its development were actually re-corded on the same machine. Therefore the flutter and wow on replay was at worst the sum of the record and replay irregularities in speed, and as previously mentioned was hardly noticed on speech. There should be less flutter and wow in a complete system where the tapes are recorded on a more conventional drive running at some multiple of the replay machine.

The complete prototype was presented for test to various blind persons, and their comments were most favourable. The only criticism made was that the clip

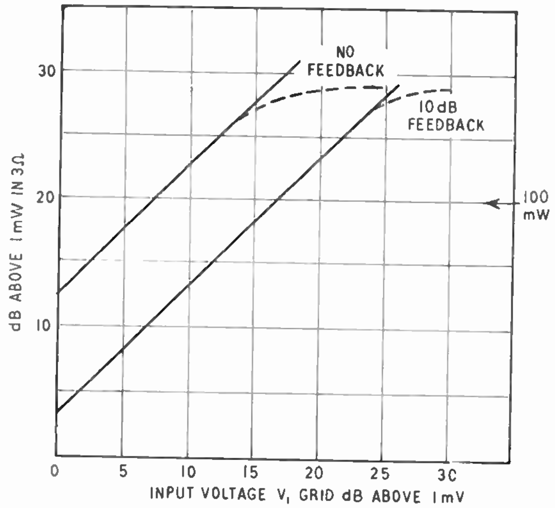

Fig. 7. Input/output curves for "talking book" amplifier.

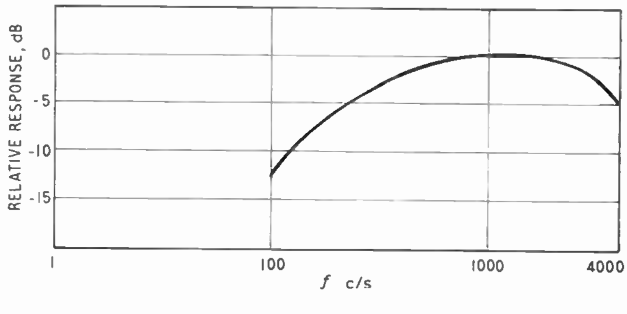

Fig. 8. Overall response of recording and reproducing system using external record amplifier with pre-emphasis. (Tape speed $1\frac{7}{8}$ in./s.)

lever could be extended each side so as to form a "bar" across the top of the "letterbox" aperture rather than the central version, shown in the photographs. This would enable the lever to be located and operated more easily.

The overall frequency response of the system using an external record amplifier with the usual boost at high frequencies is shown in Fig. 8. This represents the performance electrically from the input of the microphone to the output at the loudspeaker terminals of the reproducer.

The average book takes about 16 hours of continuous speech, reading at normal speed, and this can be accommodated in five magnetic tape magazines using both sides of the tape and limiting the wavelength as suggested. This could be reduced to two cassettes of improved design (500 ft capacity).

The weight of the prototype magazine of aluminium construction was 0·6 lb. This figure could be reduced by using a moulded plastic (styrene, etc.) case for the magazine, or one of glass fibre construction giving high strength and low weight. The prototype reproducer illustrated measured 5½ in. × 7½ in. × 8½ in. and being a mains-operated type, weighed 8 lb. A transistorized version was also developed, using a four-stage amplifier for driving lightweight headphones. This version was, of course, smaller and lighter. The case measured 4½ in. × 5 in. × 6½ in., and an extra micro-switch was incorporated in the battery circuit so that the amplifier was brought into operation only when the magazine was pushed into position.

The apertures in the end of the magazine are small enough to prevent curious fingers pulling the tape out of position and damaging it, and in a magazine-loaded device of this nature the tape must be protected as much as possible. The internal brake lever prevents dislocation of the tape due to rough handling, etc.

4. Production of Multi-track Loop Copies

The original recorded material could be in the form of an edited single track master, recorded at any convenient tape speed. This is then replayed and the signal fed to a multi-channel machine which will record a multi-channel master at "book speed" (i.e. the speed of the "talking book" playback machine).

This second master can be in two spool, or endless loop cassette form, the latter probably being the most convenient. This is then played back at a fixed multiple (say 4) of "book speed" by a machine using multiple replay amplifiers, simultaneously, their outputs being fed to the final copy-making machine also running at the same speed. This machine would have multiple record amplifiers which are designed to cover increased signal frequencies due to this higher tape speed. The system can probably be more easily visualized by examination of the block diagram shown in Fig. 9. As a final refinement the copy-making machine could be made to handle several cassettes at once, as suggested earlier.

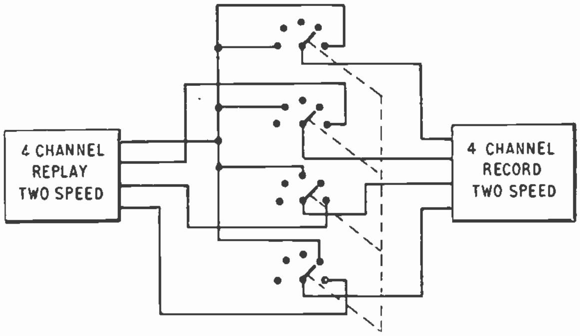

Fig. 9. Block diagram of method of producing loop copies.

The minimum amount of equipment would therefore be two machines, A and B. Machine A is a four-channel two-speed reproducer only, whilst B is a four-channel record only machine, again having two speeds like the reproducer, for instance 1⅞ in/s and 7½ in/s.

The edited master in single channel form is first played at its correct speed on machine A, one channel only being used and its output fed via a channel selector switch to machine B running at "book speed" (e.g. 1⅞ in/s). The original master should have four parts whose running time approximately equals the time taken for the cassette to complete one cycle, so that at the end of each cycle the channel selector switch is operated. The fully recorded four-channel endless loop master thus made is then transferred to machine A, the channel switch moved to the "straight through" position, and the speed switches on both machines turned to 7½ in/s. (It is assumed that the speed switches on the machines also alter the amplifier characteristics to suit the relevant speed.) In this way the copy cassette is recorded in one cycle and the time taken thereafter to produce each copy would be of the order of six minutes per copy. The recording machine B could be designed to take a number of cassettes at one time. Alternatively a number of identical machines may be used with their inputs paralleled which is the system commonly used for producing pre-recorded tapes.

5. Acknowledgments

The Author wishes to thank the Directors of Messrs. Kelvin & Hughes Limited for their permission to publish the information obtained during the development of this experimental machine.

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News from the Sections

Montreal Section

On 22nd February last, the Section held a meeting in the McConnell Engineering Building of McGill University at which a paper on "Thermo-Electricity" was read by Mr. J. Keane.

The history of the phenomenon was first discussed, leading up to the important Seebeck and Peltier equations. The latest developments arose from Ioffe's work in Russia, in which he synthesized calculations of thermo-electrical problems. Mr. Keane then discussed the significance of the Z-quality factor and its relationship to the basic thermo-electric equation. Data were presented in the form of curves showing Δt versus current and also Δt versus QC, and examples were given of their application to design. Future uses of thermo-electric elements were discussed and several demonstrations were given.

K. N. C.

North Western Section

The Section held its fourth meeting of the session at the Manchester College of Science and Technology on 2nd February when Mr. P. Funnell of the Mullard Simonstone Works discussed "The Manufacture of Television Tubes".

Mr. Funnell gave a brief history of the development of the television tube including its design and the methods of manufacture. He divided the periods of advances in technique into three main phases:

1946–54—The introduction of larger rectangular tubes,

1954–58—Mechanization when new factories were built which were solely concerned with television tube production, and

1958–60—Introduction of the 110 degree tube.

Some of the more important advances which occurred in the stages of development were then reviewed. Stage 1 was, said Mr. Funnell, primarily concerned with the development of pressed glass techniques and the tetrode electron gun. Stage 2 brought in advances in manufacturing techniques, a typical example being improvements in the deposition of the fluorescent screen. One of the biggest changes was the introduction of aluminizing as a mass production process.

With the introduction of the 110 degree tube, stage 3 was concerned with overcoming the design problems which included:

(*a*) The design of bulbs to avoid "corner cutting".

(*b*) A new electron gun and modified processing equipment following the advent of the smaller neck.

(*c*) Special aluminizing technique to avoid the "dark centre" on the screen and "damping" effect on deflection coils. (A demonstration of this effect was given.)

(*d*) Cleanliness and accurate spacing of gun parts to avoid the dangers of "flash over".

Finally, Mr. Funnell discussed the latest type of tube, namely the 19 in, where the major and minor axes are slightly bigger than the 17 in, but the corners are squarer. There is greater electron beam deflection in the corners and the neck/cone area is shaped on the inside so that a section of the glass tends to be rectangular.

F. J. G. P.

West Midlands Section

At a well-attended meeting held at Wolverhampton on 15th February, Mr. R. Wooldridge read a paper on "Computers and Mathematics". His theme was that, with the increasing use of computers, engineers required a knowledge of several mathematical techniques which at present are not included in most engineering syllabuses: as an example of this, binary arithmetic was quoted.

The evaluation of a polynomial of high degree was then discussed. It was shown, using numerical analysis, that a method could be evolved which was particularly suited for use with a machine.

The final part of the lecture was taken up by a discussion of Boolean algebra. This algebra was shown to be particularly relevant to the solution of problems of the AND/OR type; this point was illustrated by examples of switching circuits.

The meeting on 22nd February, was held in Birmingham and Mr. Peter Styles described the development work carried out by the electronics group at St. Thomas' Hospital. This work consists of the development of specialized equipment which is not available commercially.

Among the items mentioned were systems for measuring muscle potentials, of the order of microvolts, and for measuring the velocity with which an impulse is propagated along a nerve fibre.

One of the most interesting of the systems described was one for the measurement of the rate of blood flow along a blood vessel, without severing the vessel; this was done by measuring the time taken for an ultrasonic pulse to travel along a fixed length of the vessel.

Finally, a short account was given of very small pulse generators, designed to be placed inside the chest cavity, which would provide a triggering impulse to the heart of a patient whose heart would otherwise fail to beat.

D. H. A.

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Video Integration in Radar and Sonar Systems

By

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AND

J. W. R. GRIFFITHS, Ph.D.,
(Associate Member) †

Summary: The performance of some basic integration systems is analysed, and the effect of beam pattern on their performance is considered. A new system is described employing two delay loops which, in general, will give a small improvement in threshold detection and furthermore has a number of practical advantages.

List of Symbols

β	voltage or current gain around delay loop.
θ	parameter proportional to time.
θ_1	value of θ which corresponds to the output time for the single loop integrator.
θ_2	value of θ which corresponds to the output time for the double loop integrator.
θ_U	integration limit of θ which applies to uniform weighting.
σ	r.m.s. of noise voltage or current.
ϕ	spacing between adjacent values of θ defining signal amplitudes.
F_1	signal/noise improvement factor for single loop integration.
F_2	signal/noise improvement factor for double loop integration.
F_I	signal/noise improvement factor for ideal weighting.
F_U	signal/noise improvement factor for uniform weighting.
$f_1 = F_1/F_I$	
$f_2 = F_2/F_I$	
k	weighting factor applied to waveform samples before addition.
$k(\theta)$	envelope of weighting function in terms of θ.
M	number of waveforms which contain signal.
N	number of waveform samples integrated.
n	number of repetition periods a waveform sample is stored.
$q = \frac{\log \beta}{\phi}$	– parameter which determines integrator performance.
s	signal measured as change in mean level (voltage or current).
$s(\theta)$	signal envelope as a function of θ.
t	time.
t_d	delay time of delay line.
$x = \frac{\theta_1 - \theta}{\phi}$	
$y = \frac{\theta_2 - \theta}{\phi}$	

1. Introduction

In all echo-ranging systems the repetitive received signals from any one target are limited to some finite interval of time which is determined by the system. It is usually convenient to subdivide this interval into sub-intervals which correspond exactly to the repetition cycles of the system and then define the duration of the signals from one target by the number of sub-intervals or cycles in which the signal occurs. It is well known that, as far as marginal detection is concerned, the performance of the equipment can be much improved by the storage and subsequent collation of the information from these successive cycles. Elementary forms of storage were inherent in even the earliest radar equipments and without them the performance would have been much inferior. A typical example is the 'A' scan; here the envelope of the received waveform is displayed on a cathode-ray tube against a time-base synchronized to the transmitter repetition rate and thus successive traces are superimposed. A combination of the afterglow properties of the phosphor of the tube and the eye and mind of the observer serve as the storage mechanism and it has been shown[1] that substantial improvements in detection are obtained with such a display. Intensity modulated displays such as the chemical recorder,[2] used extensively in sonar equipment, are also very

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† Electrical Engineering Department, University of Birmingham.

effective. Here the traces may be displayed side-by-side and the eye detects the presence of a target as a pattern in a random background. Attempts have been made to develop a more sophisticated form of video storage and recently practical systems have been described which have made use of mercury delay lines[3] in one case and digital registers[4] in another.

One method of using a mercury delay line (or for that matter any other delay line of suitable length) is to recirculate the old information in the same delay line. The length of the delay line is made exactly equal to the recurrence period of the radar and the old information coming out of the line is added directly to the new information being received, the resulting summated waveform being reinjected into the line. This system, whilst reducing the total storage capacity required and having the merit of simplicity, imposes a number of restrictions on the system. In theory, of course, it would be possible to recirculate without adding the traces by using a frequency multiplex system, but, quite apart from the complexity of the terminal equipment, the bandwidth of the delay line would restrict the number of traces that could be stored, and furthermore, it can be shown[5, e.g.] that the addition of traces or, to give the process its more common name, integration, does not degrade the information. However, since the dynamic range of a delay loop is limited, the process of adding the old to the new cannot be carried out indefinitely unless some weighting factor is applied which restricts the summation to a finite total. The weighting factor may be such as to allow a fixed number of traces to be integrated, the store then being cleared and a new series started. This system is generally known as group integration. Continuous integration may be obtained by weighting the old information by a factor less than unity, say β, since the sum of such a series is finite. The larger the dynamic range available the more nearly β can approach unity and hence the more traces that are integrated, but, of course, there are other difficulties besides dynamic range which limit the value of β. Both the continuous and the group integration schemes have disadvantages which will be discussed later. An incidental advantage of integration systems is their ability to discriminate against coherent interference such as may originate from adjacent echo-location equipments.[6]

In spite of this wide knowledge of the benefits of storage systems little quantitative practical information is available on the improvement resulting from the application of such methods. On the theoretical side the material is a little more promising but of course it is obvious that the theoretical measure of performance will depend upon the criterion of detection chosen. This problem has been discussed fairly fully elsewhere.[7, 8] However, it is fairly obvious that whatever the criterion chosen the actual improvement of performance due to integration is best measured in terms of the threshold signal/noise ratio, i.e. the value of the input signal/noise ratio necessary for a particular value of the accepted criterion.

One widely accepted criterion relates the change in mean voltage level which occurs when the signal is present to the r.m.s. of the background noise when the signal is absent. Adding N traces will increase the change in mean by a factor of N but the noise, adding on a power basis because of the lack of correlation from trace to trace, will increase by a factor of only $\sqrt{N}$. Hence the numerical value of the criterion increases by a factor of $\sqrt{N}$. For small signal/noise ratios, however, the change in mean in the output of an envelope detector is proportional to the square of the input signal/noise ratio. Thus if we reduce the input signal/noise ratio so that the value of the criterion for the N traces is the same as for only one trace the depression of the threshold is only $\sqrt[4]{N}$. This is the well known fourth root law of improvement and means that for each doubling of the number of traces integrated the input threshold signal/noise ratio is reduced by 1·5 dB.

On the other hand, experimental evidence[8, 9] indicates that a better return can be obtained using the side-by-side storage systems where a figure of 2·2/2·4 dB per doubling was obtained, and suggest that such systems may be superior to the integration methods. Unfortunately, to the authors' knowledge little or no experimental evidence has been published on integration systems, but we now have some results which indicate that integration systems are not inferior and that it is the criterion which is in error. These experimental results tie up with some of the points made in the discussion[8a] on reference 8, in which the work of Marcum[10] was mentioned. Marcum's work, which has been only recently published,[10a] is based on a quite different criterion of detection. He assumes that the integrated trace is applied to a circuit which will respond only when the voltage reaches a predetermined bias level, and if the voltage exceeds this level it is assumed that a target is present. The bias level is fixed by the "false alarm" rate which can be tolerated, a false alarm being due to the crossing of the bias level when noise alone is present in the input. Given the bias level, the probability of the voltage exceeding this level when the signal is present, i.e. the probability of detection, can be determined. A practical realization of this method lends itself to automatic detection systems. Automatic detection systems are becoming of increasing importance in complex situations such as exist at an aircraft terminal. The operator (if he still exists) can be presented with a filtered display, resulting in less fatigue and hence less deterioration in performance.[11]

2. The Delay Line Integrator

Although it has been suggested that the criterion given by the ratio

$$\frac{\text{change in mean level when signal applied}}{\text{r.m.s. of noise in the absence of signal}}$$

cannot always be linearly related to detectability, the optimization of this ratio will normally result in optimum signal detection. In this paper we will be concerned with the optimization of integrator performance and hence the use of this criterion is permissible. It has the advantage of simplifying the mathematics considerably!

The use of a delay line and feedback loop as a video integrator in an echo ranging equipment implies that some method is devised to limit the time any waveform is stored to something of the order of the interval in which the input signals are known to occur, since longer overall storage times cause waveforms containing noise alone to be incorporated in the integrator output thus degrading its performance.

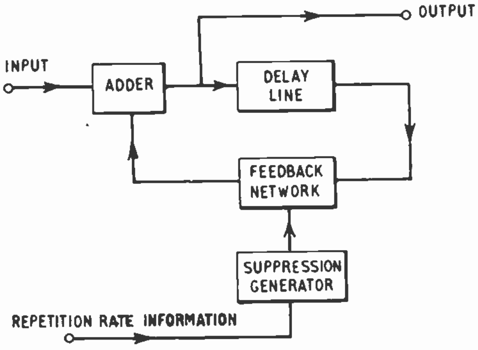

Fig. 1. Block diagram of a loop iterator.

The block diagram of Fig. 1 shows one way in which a delay line may be arranged to integrate a limited number of repetitive waveforms, the arrangement being termed an iterator or group integrator. In this arrangement the delay of the line is made equal to the repetition period of the echo ranging system and the output of the delay line is fed back and added to the input. The loop gain is arranged to be unity for a time corresponding to the required integration time and is then rapidly reduced to zero and kept there for one repetition period of the input waveform. This process is repeated continuously.

The simple iterator which we have described carries out the addition of a given number of input waveforms only, and therefore will not operate on waveforms containing no signal provided that the suppression can be synchronized to the start of the set of signal waveforms. However, an iterator is usually used to obtain some improvement in the detectability of the signal waveforms and under such conditions we cannot presume any prior knowledge of the waveform in which the signal will first be present. It is therefore highly probable that the input signals will be divided between two integration periods resulting in an inferior integrated output for each.

This phenomenon is called signal splitting and the degraded performance caused thereby can only be avoided completely by the use of a number of iterators equal to the number of repetitive waveforms in which the signal is known to appear. In this case the suppression of the iterators is staggered and occurs at intervals corresponding to the repetition period of the input waveforms so that one integrated output is available during each repetition period. It is obvious that this solution to the problem is very costly in terms of equipment when there are a large number of waveforms in which the signal is expected to exist.

The improvement in the simple signal/noise criterion produced by the iterator is easily determined. We will assume that the change in mean voltage or current produced by the presence of a signal in any input waveform is denoted by s and that the r.m.s. of the background noise is σ.

If the iterator operates on N waveforms, all of which contain the signal, we have a resultant change of mean of Ns, and a resultant r.m.s. background noise of $(N\sigma^2)^{\frac{1}{2}}$. It should be noted that these results apply for any statistical distributions of the noise and noise-plus-signal waveforms, provided there is no correlation between successive noise contributions.

Thus the final value of our criterion is

$$\frac{Ns}{(N\sigma^2)^{\frac{1}{2}}} = \sqrt{N} \cdot \frac{s}{\sigma} \qquad \ldots\ldots(1)$$

giving an improvement ratio of $\sqrt{N}$.

When only M of the waveforms contain signal the change in mean of the resultant waveform will be reduced to Ms and the final value of the criterion will be

$$\frac{M}{\sqrt{N}} \cdot \frac{s}{\sigma} \qquad \ldots\ldots(2)$$

In this case the improvement ratio can be written in the form

$$\frac{M}{N} \cdot \sqrt{N} = \frac{M}{N} \times \text{optimum improvement ratio} \qquad \ldots\ldots(3)$$

The worst case which can occur with a single iterator is that in which the group of signals is equally divided between two integrated groups and M/N is equal to 0·5 for each. A loss of 6 dB in video signal/noise ratio occurs in this case, when the performance is compared with that obtained with the integration group corresponding exactly to the signal group.

Another method for limiting the time of storage of a delay line loop is to operate continuously, but with a loop gain β which is less than unity. This latter restriction is also, of course, a stability requirement for continuous systems. In this arrangement we still obtain a maximum output at the termination of a set of input signal waveforms but the integrated output is the weighted sum of all preceding waveforms, the weighting factor for each at any particular instant being a function of the time which has elapsed since its arrival.

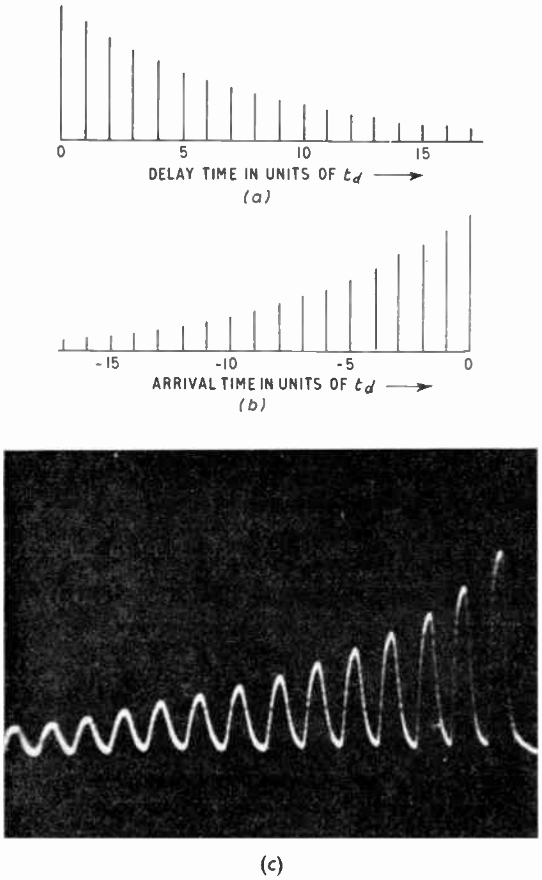

Fig. 2. (*a*) Impulse response of the loop integrator.
(*b*) Weighting function of the loop integrator.
(*c*) Display of integrator weighting function.

The values of the weighting factors may readily be obtained by considering the response of this integrator to an input pulse of unit amplitude. If the pulse is applied to the integrator at time $t = 0$ and the delay of the line is t_d, the output at time $t = t_d$ will be a pulse of amplitude β, and at time $t = 2t_d$ the output pulse will have amplitude β^2. In other words the pulse amplitude is modified by the factor β every time a transit around the loop is completed and at time $t = nt_d$ the output amplitude will be β^n.

Figure 2(a) is a sketch which shows the decaying form of this output and it is not difficult to see that if the response is reversed with respect to time $t = 0$, as is shown in Fig. 2(b), the weighting as a function of time since arrival of signal is obtained. The time at which the output of the integrator is observed is now $t = 0$.

The weighting function may also be obtained experimentally by applying a continuous repetitive pulse to the input of the system and adjusting the repetition period of the input to be slightly greater than the delay of the loop. In this case each of the input pulses produces a set of responses such as is shown in Fig. 2(a) and individual pulses from the responses of all the input pulses are grouped closely together just before each pulse of the input train. A photograph of an oscilloscope display of the integrator output showing one group in such circumstances is shown in Fig. 2(c).

If the integrator is to operate on a signal which is known to appear in N successive waveforms the optimum output will be available at the same time as the arrival of the last waveform containing signal and the value of β chosen must be a compromise between the requirements of a high value of the weighting function for waveforms containing signal and a low weighting factor for waveforms containing noise only. The optimum performance will not be as good as that obtained with the iterator under ideal synchronization conditions but the system has the advantage of being simple and continuous in its operation. Old information is lost in a gradual manner and no signal splitting occurs.

The video signal/noise ratio for N signal-plus-noise input waveforms can now be found. This analysis has been published before [e.g. 3] but it is included here for the sake of completeness.

Thus output signal/noise ratio

$$= \frac{s+\beta s+\beta^2 s+\ldots\beta^{N-1}s}{(\sigma^2+\beta^2\sigma^2+\beta^4\sigma^2+\ldots)^{\frac{1}{2}}}$$

$$= \frac{s}{\sigma}\left\{\frac{1-\beta^N}{1-\beta}\right\}[1-\beta^2]^{\frac{1}{2}}$$

$$= \frac{s}{\sigma}(1-\beta^N)\left\{\frac{1+\beta}{1-\beta}\right\}^{\frac{1}{2}}$$

and the *video* signal/noise improvement ratio

$$= (1-\beta^N)\left\{\frac{1+\beta}{1-\beta}\right\}^{\frac{1}{2}} \qquad \ldots\ldots(4)$$

Figure 3 shows the way in which this improvement ratio varies with N for a number of values of β. Remembering that the maximum improvement ratio obtained with an iterator is $\sqrt{N}$ we find that the ungated loop integrator is about 1 dB inferior in performance when β has been optimized. This is not a particularly high price to pay for a much simpler continuous system and there are many radar equipments in existence which incorporate this arrangement.

However, the ungated loop integrator is not a very satisfactory solution to the problem of integrating waveforms when the number of these waveforms is large, for example one hundred or more, since this requires the use of a loop gain very close to unity. Urkowitz[12] has described a system using frequency modulation which maintains accurately a loop gain very close to unity, but stability requirements still limit the use of the system to the integration of less than about one hundred signal-plus-noise waveforms.

3. The Effect of Beam Pattern

The finite duration of the signals received by an echo-ranging system is often the result of the search role of the equipment. Thus the sweeping of a narrow beam of energy through the bearing of the target produces a set of input waveforms which contain signals in proportions determined by the beam pattern. We are therefore presented with the problem of finding the best way in which to integrate a set of waveforms in which the signal/noise ratio increases and decreases in sympathy with the beam pattern.

3.1. *Optimum Weighting*

Let us consider the problem of optimizing the signal/noise criterion of the waveform resulting from the weighted addition of N waveforms where the change in mean due to the application of signal in each waveform is given by $s_1, s_2 \ldots s_N$, and the r.m.s. background noise is σ for all waveforms.

Note that we are dealing with the signal/noise ratio after detection but this is a function of the input signal/noise ratio. The signal/noise ratio after detection (as defined in Section 2) is proportional to the square of the input signal/noise ratio for values of the latter below unity.

We assume a set of weighting factors $k_1, k_2 \ldots k_N$ and now find that the signal/noise ratio for the weighted sum is given by

$$\frac{k_1 s_1 + k_2 s_2 + \ldots k_N s_N}{[k_1^2\sigma^2 + k_2^2\sigma^2 + \ldots k_N^2\sigma^2]^{\frac{1}{2}}}$$

$$= \frac{\sum\limits_{r=1}^{N} k_r s_r}{\sigma\left[\sum\limits_{r=1}^{N} k_r^2\right]^{\frac{1}{2}}} = \frac{A}{\sigma} \qquad \ldots\ldots(5)$$

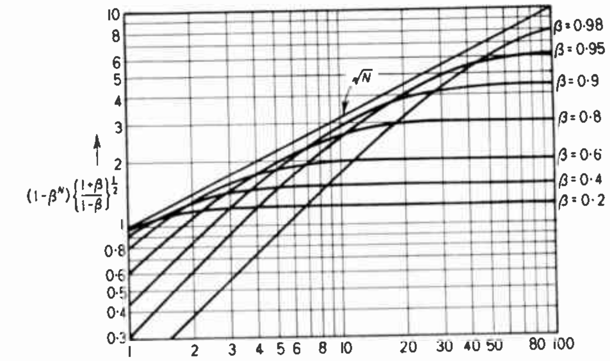

Fig. 3. Improvement of signal/noise ratio as a function of N.

In order to find turning values for this expression we require that all the partial derivatives with respect to the k_r's are zero.

Hence we obtain

$$\frac{\partial A}{\partial k_i} = \frac{s_i}{[\sum k_r^2]^{\frac{1}{2}}} - \frac{k_i \sum k_r s_r}{[\sum k_r^2]^{3/2}} = 0 \text{ for all values of } i. \qquad \ldots\ldots(6)$$

We therefore write

$$s_i \sum k_r^2 = k_i \sum k_r s_r$$

or

$$\frac{s_i}{k_i} = \frac{\sum k_r s_r}{\sum k_r^2} = \text{constant for any particular set of } k_r\text{'s and } s_r\text{'s} \qquad \ldots\ldots(7)$$

It is not difficult to show that this condition gives us a maximum value of A and we therefore have the result that in order to obtain the optimum performance from an integrator the individual waveforms must be weighted by a factor which is directly proportional to the change in mean level produced by the signal present in that waveform.

This automatically implies that no account is taken of waveforms containing noise only and in addition indicates that all waveforms containing signal plus noise should be used. Perhaps this is only too obvious, but we now know that uniform weighting of all signal-plus-noise waveforms is not always the best procedure. Although we have assumed that σ is constant in the above argument it is not necessary to do so, and the presence of noise which depends upon beam pattern can also be taken into account. This gives the result that $\frac{s_i}{k_i \sigma_i^2}$ = constant for all values of i. For the remainder of the analysis we will, however, continue with our original assumption of constant noise background.

In order that we may estimate what improvement in signal/noise ratio is to be obtained by the use of ideal weighting as opposed to uniform weighting we must assume some particular form for the distribution of the signal amplitudes. The problem of uniform weighting has been considered by Blake[13] who has assumed that the signal (interpreted as change in mean voltage throughout the remainder of the paper) is distributed among the waveforms with an envelope of Gaussian form. The Gaussian form is a close approximation to the beam pattern of a narrow beam antenna and we will adopt this form in the following discussion. We shall also follow the approach taken by Blake in assuming that there are a large number of individual signal waveforms under the assumed envelope. This permits us to use integral calculus instead of algebraic summations and the results appear in a parametric form which is of general use. Blake has shown that in certain cases the results of the calculus treatment for the high pulse density case correspond to the mean of all possible results in the low pulse density case and he points out that a mean value is then the only useful measure of performance.

The envelope of the signal amplitude is therefore defined by the curve $\exp(-\theta^2)$ where θ is some parameter which is directly proportional to time. We will first obtain an expression for the improvement ratio obtained with optimum weighting and in order to do this we assume that the maximum signal at $\theta = 0$ is s_0 and the r.m.s. noise in all the waveforms is σ. The values of θ which determine the individual signal amplitudes will be assumed to be separated by the small increment ϕ.

The ideally weighted integrator will therefore produce an output signal/noise ratio given by

$$\frac{\sum k_r s_r}{\sigma[\sum k_r^2]^{\frac{1}{2}}} = \frac{s_0 \sum\limits_{r=-\infty}^{r=+\infty} \exp(-2r^2\phi^2)}{\sigma\left[\sum\limits_{r=-\infty}^{r=+\infty} \exp(-2r^2\phi^2)\right]^{\frac{1}{2}}} \qquad \text{......(8)}$$

since we write $k_r = \frac{s_r}{s_0} = \exp(-r^2\phi^2)$.

If we write $k(\theta)$ and $s(\theta)$ for the functions which define the values of the weighting and the signal amplitudes, the general expression for the output signal/noise ratio can be written in integral form.

Thus our expression becomes

$$\frac{\frac{1}{\phi}\int\limits_{-\infty}^{+\infty} k(\theta)\,s(\theta)\,\mathrm{d}\theta}{\sigma\left[\frac{1}{\phi}\int\limits_{-\infty}^{+\infty}\{k(\theta)\}^2\,\mathrm{d}\theta\right]^{\frac{1}{2}}} \qquad \text{......(9)}$$

Putting $k(\theta) = \frac{s(\theta)}{s_0} = \exp(-\theta^2)$ we obtain

$$\frac{s_0}{\sigma}\cdot\frac{1}{\sqrt{\phi}}\left[\int\limits_{-\infty}^{+\infty}\exp(-2\theta^2)\,\mathrm{d}\theta\right]^{\frac{1}{2}} = \frac{s_0}{\sigma}\cdot\frac{1}{\sqrt{\phi}}\cdot\sqrt[4]{\frac{\pi}{2}}$$

The improvement factor for ideal weighting is therefore

$$F_I = \frac{1}{\sqrt{\phi}}\cdot\sqrt[4]{\frac{\pi}{2}} = \frac{1{\cdot}119}{\sqrt{\phi}} \qquad \text{......(10)}$$

It should be noted that the form of signal envelope we have assumed implies that for ideal weighting an infinite number of waveforms must be used. This is obviously not possible in any practical situation but an approach to the ideal improvement can be made with a finite number of waveforms since the function $\exp(\theta^2)$ falls off very rapidly for large values of θ. We will not pursue this any further at the moment but will use the ideal improvement expression as a standard of comparison.

3.2. *Uniform Weighting*

Returning to the case of uniform weighting it is obvious that only a restricted number of waveforms should be used and we wish to find the number that give optimum performance. This optimum performance will also correspond to the integration of an equal number of waveforms on each side of that containing the maximum signal and we assume that these waveforms are bounded by values of θ such that $-\theta_U < \theta < \theta_U$.

The output signal/noise ratio is now

$$\frac{\frac{1}{\phi}\int\limits_{-\theta_U}^{+\theta_U} s_0\exp(-\theta^2)\,\mathrm{d}\theta}{\sigma\left[\frac{1}{\phi}\int\limits_{-\theta_U}^{+\theta_U}\mathrm{d}\theta\right]^{\frac{1}{2}}} \quad \text{since } k = 1 \text{ for all } \theta \text{ within the assumed range} \qquad \text{......(11)}$$

The improvement ratio is now

$$F_U = \frac{1}{\sqrt{\phi}}\,\frac{1}{\sqrt{2\theta_U}}\int\limits_{-\theta_U}^{+\theta_U}\exp(-\theta^2)\,\mathrm{d}\theta = \frac{1}{\sqrt{\phi}}\cdot\sqrt{\frac{\pi}{2}}\cdot\frac{\operatorname{erf}\theta_U}{\sqrt{\theta_U}} \qquad \text{......(12)}$$

The maximum value of the expression $\frac{\operatorname{erf}\theta_U}{\sqrt{\theta_U}}$ is $0{\cdot}8427$ and occurs when $\theta_U = 0{\cdot}994$, therefore the optimum improvement factor in the unweighted case is

$$F_{U\,\max} = \frac{1}{\sqrt{\phi}}\sqrt{\frac{\pi}{2}}\,.\,0{\cdot}8427 = \frac{1{\cdot}0559}{\sqrt{\phi}} \qquad \text{......(13)}$$

This is only 0·5 dB less than that obtained using ideal weighting and thus shows that weighting can produce only a small improvement in performance and from this point of view would not be worth using unless it could be achieved without increasing the complexity of the equipment.

It is worth noting that the optimum value found for θ_U in the unweighted case indicates that unweighted integration should only be performed on waveforms with video signal/noise ratio greater than 8·66 dB below the maximum. However, performance is not critically dependent on the number of waveforms integrated.

4. The Video Delay Loop Integrator operating on a Gaussian Input Signal Train

We have just shown that an integrator which is capable of unweighted operation on a group of input waveforms does not fall far short of the ideal performance, but we must remember that in order to obtain the best performance from the group integrator it must operate in a continuous fashion and so avoid signal splitting. The requirement of continuous operation can only be met by using a large number of basic storage systems such as the iterator, and MacFarlane[3] has considered some of the problems involved.

Another approach to the problem has been described by Hinckley,[4] and this method involves the quantization of the amplitude of the individual waveforms into two regions, one above and one below some threshold level. After quantization the waveforms appear as a series of marks and spaces which are handled in digital fashion, and integration is achieved by comparing with a second threshold the number of digits occurring at each range position in a fixed number of waveforms.

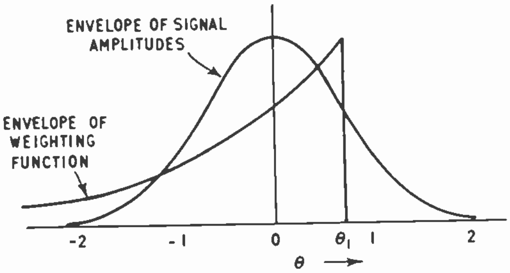

Fig. 4. Relative positions of signal envelope and weighting function of loop integrator.

The use of this double threshold method does simplify the apparatus to some extent, although it is still quite complicated and therefore costly, and in addition some loss in performance must result from the quantization. In fact Harrington[14] has shown that between 1 and 2 dB loss in performance results if the double threshold method is compared with the single threshold method treated by Marcum and the input signals are assumed to have constant amplitude. Therefore the loss in video signal/noise performance when compared with an ideally weighted integrator operating on a set of signals with a Gaussian envelope will probably be between 1·5 and 2·5 dB.

In view of the complication introduced if the group type integrator is to be made to operate in a continuous fashion it is of interest to determine what improvement can be obtained if a continuous delay loop integrator operates on a set of signals with amplitudes distributed according to the Gaussian law.

We shall consider two forms of delay loop integrator, the single loop type which we have already described and a double loop system which will be shown to possess certain advantages. In order that we may use the integral form of the expression for video signal/noise improvement ratio we must express the envelope of the weighting function for each arrangement as a function of the parameter θ.

The single loop integrator has been shown to have a weighting function given by the values 1, β, β^2... etc. at times which correspond to values of θ of θ_1, $\theta_1-\phi, \theta_1-2\phi,$... etc., where θ_1 is the value of θ which corresponds to the time at which the output of the integrator is being considered.

We may therefore express the envelope of the weighting function in the form

$$\begin{aligned} k_1(\theta) &= \beta^{(\theta_1-\theta)/\phi}, \qquad \theta_1 \geqslant \theta \\ &= 0, \qquad \theta_1 < \theta \qquad \text{......(14)} \end{aligned}$$

Figure 4 shows this weighting envelope positioned with respect to the envelope of the signal set in the manner we assume for analysis.

The video signal/noise improvement ratio can now be found by using eqn. (9). Thus we obtain

$$F_1 = \frac{1}{\sqrt{\phi}}\left\{\frac{\int_{-\infty}^{+\theta_1} \beta^{(\theta_1-\theta)/\phi}\, e^{-\theta^2}\, d\theta}{\left[\int_{-\infty}^{\theta_1} \beta^{2(\theta_1-\theta)/\phi}\, d\theta\right]^{\frac{1}{2}}}\right\} \qquad \text{......(15)}$$

The part of eqn. (15) in brackets is evaluated in Appendix 9.1 eqn. (34), and hence

$$F_1 = \frac{1}{\sqrt{\phi}}\sqrt{\frac{\pi}{2}}\exp(-\theta_1^2)\sqrt{-q}\,.\,\exp(\tfrac{1}{2}q+\theta_1)^2 \times$$
$$\times[1+\text{erf}(\tfrac{1}{2}q+\theta_1)] \text{......(16)}$$

where $$q = \frac{\log\beta}{\phi}.$$

and noting that the ideal improvement factor (eqn. (10)) is given by

$$F_I = \frac{1}{\sqrt{\phi}}\sqrt[4]{\frac{\pi}{2}}$$

we can write

$$F_1 = F_I \sqrt[4]{\frac{\pi}{2}}\exp(-\theta_1^2)\sqrt{-q}\,.\,\exp(\tfrac{1}{2}q+\theta_1)^2 \times$$
$$\times[1+\operatorname{erf}(\tfrac{1}{2}q+\theta_1)] \quad \ldots\ldots(17)$$

or

$$F_1 = F_I f_1(q,\theta_1) \quad \ldots\ldots(18)$$

Since the value of θ_1 is continuously changing as the output of the integrator is observed, and q is fixed for any given loop gain, the output signal/noise ratio will vary with time in a manner defined by the way in which the above function f_1 (q, θ_1) varies with θ_1. The variation will always correspond to a curve which rises to a peak value at some value of θ_1 and we are interested in finding the value of q which maximizes this peak.

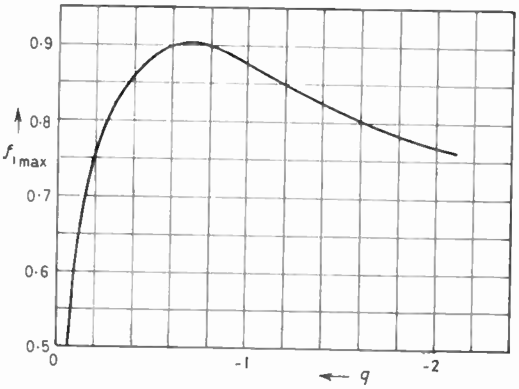

Fig. 5. Maximum of f_1 $(q, \theta_1)_{(q = \text{const.})}$ as a function of q.

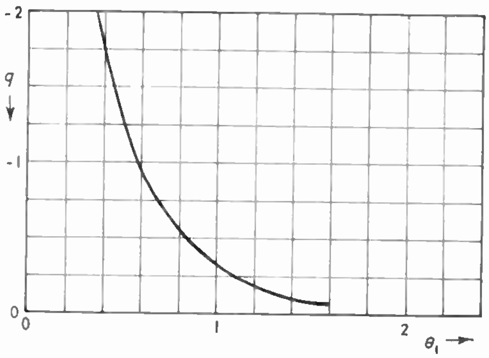

Fig. 6. θ_1 as a function of q for maximum values of f_1 $(q, \theta_1)_{(q = \text{const.})}$.

An analytical approach to this problem did not appear tractable† and values of the function have been tabulated for suitable ranges of q and θ_1. The peak values of the function for given values of q can then be obtained and Fig. 5 shows these peak values as a function of the parameter q.

Figure 6 shows the relation between θ_1 and q for a maximum of the function f_1 $(q,\theta_1)_{(q = \text{const.})}$. The optimum value of θ_1 can be regarded as a measure of the time which must elapse after the arrival of the waveform containing the maximum signal, before the output signal/noise ratio reaches its peak value. The rate of change of this time with respect to a change in q (or β) is of importance because it fixes to the accuracy with which one can compensate for the time delay of the peak output. Incorrect compensation will usually appear as an angular error in the output of an echo-ranging system.

Returning to Fig. 6 we see that optimum performance is obtained with a value of q of $-0{\cdot}7$ and the maximum value of the function is then 0·895. This indicates that the single integrator optimum performance is within 0·95 dB of the ideal.

What is rather surprising is the fact that the single integrator performance is only 0·45 dB inferior to that obtained by unweighted group integration.

If we specify a value for the number of video input signals which are not more than 6 dB below the peak level then we may obtain some idea of the value of β required for optimum performance by expressing ϕ in terms of this number and thence log β in terms of ϕ and the optimum value of q. The figure of 6 dB has

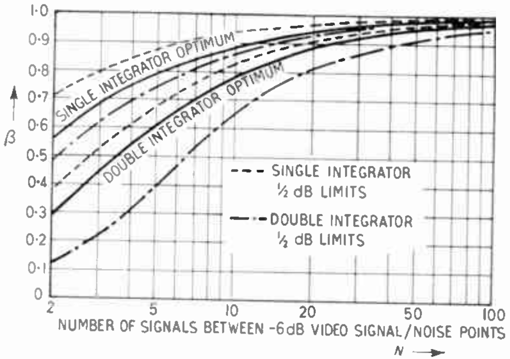

Fig. 7. Optimum and limit values of loop gain β for single and double integrators.

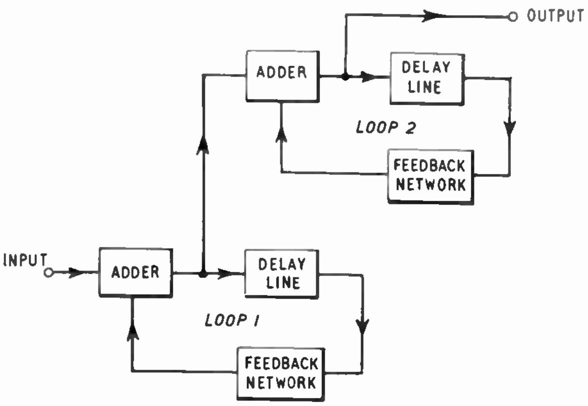

Fig. 8. Block diagram of double loop integrator.

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† It has since been brought to our notice that it is possible to solve this problem by an iterative method which is rapidly convergent.

been chosen here because it corresponds to the 3 dB range of the pre-detector r.f. signal/noise ratios.

The above range of signal/noise ratio corresponds to a range of θ from $-0{\cdot}835$ to $+0{\cdot}835$ and if there are N signal waveforms in this range we have

$$\phi = \frac{1{\cdot}67}{N} \qquad \ldots\ldots(19)$$

Hence

$$\log \beta_{\text{OPT}} = \phi \, q_{\text{OPT}} = \frac{-1{\cdot}67 \times 0{\cdot}7}{N} = \frac{-1{\cdot}17}{N} \qquad \ldots\ldots(20)$$

or $$\beta_{\text{OPT}} = \exp\left(\frac{-1{\cdot}17}{N}\right) \qquad \ldots\ldots(21)$$

Optimum values of β for a range of values of N are shown in Fig. 7 and also included are the limiting values for β to maintain the performance within 0·5 dB of the optimum. These 0·5 dB loss curves correspond to values of q of $-0{\cdot}4$ and $-1{\cdot}16$, and the values of q for other limit losses may be obtained, if required, from Fig. 5.

5. The Double Integrator

The double integrator is an arrangement in which the output of the first loop is operated on by a second loop. Identical delays are required and for the purposes of the analysis the gain of each loop will be assumed to have the same value β.

Figure 8 shows the arrangement of the double integrator system, and we will consider the response of this system to a single input pulse of unit amplitude applied at time $t = 0$.

A simple method of obtaining the response is to tabulate two rows of pulse amplitudes in the time sequence in which they occur. The top row is the response of the first loop to the pulse and this we already know. Below this row we form the second row corresponding to the input of the second delay line and also the output of the system. The amplitudes of the second row are obtained by adding β times the amplitude immediately to its left, that is the previous input to the second loop, to the amplitude immediately above in the first row.

It is not difficult to see that this method corresponds to the way in which the impulse response is formed and the table which results is shown below.

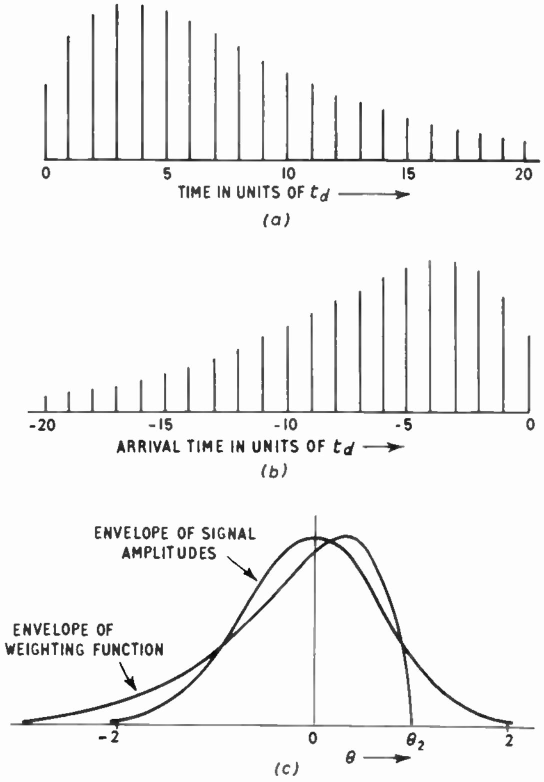

Fig. 9. (*a*) Impulse response of double loop integrator.
(*b*) Weighting function of double loop integrator.
(*c*) Relative position of signal envelope and weighting function of the double loop integrator.

Some idea of the form of this response is shown in Fig. 9(a) and as before we obtain the weighting function by reversing the impulse response. This weighting function is shown in Fig. 9(b), and it will be observed that it is a better approach to the ideal weighting, assumed to be Gaussian, than that given by the single integrator. The envelope of the weighting function for the double integrator is the product of the weighting function of the single integrator and a linearly rising function. In Fig. 9(c) the envelope of the weighting function is compared with the signal envelope in the relative position which will be shown later to give the optimum performance.

Table 1

	Time t							
	0	t_d	$2t_d$	$3t_d$	$4t_d$	...	nt_d	...
1st loop: output .	1	β	β^2	β^3	β^4	...	β^n	...
2nd loop: output .	1	2β	$3\beta^2$	$4\beta^3$	$5\beta^4$	...	$(n+1)\beta^n$	...

We now obtain the expression for the envelope of the weighting function in the manner already adopted. Thus we have

$$k_2(\theta) = \frac{\theta_2-\theta}{\beta\phi}\beta^{(\theta_2-\theta)/\phi} \qquad \text{......(22)}$$

but in this case θ_2 corresponds to the time one period after the time at which we are observing the integrated output, this slight change of reference being made to simplify the expression.

The expression for the improvement ratio is now

$$F_2 = \frac{1}{\sqrt{\phi}}\left\{\frac{\int\limits_{-\infty}^{\theta_2}\frac{\theta_2-\theta}{\beta\phi}\beta^{(\theta_2-\theta)/\phi}\exp(-\theta^2)\,d\theta}{\left[\int\limits_{-\infty}^{\theta_2}\left(\frac{\theta_2-\theta}{\beta\phi}\right)^2\beta^{2(\theta_2-\theta)/\phi}\,d\theta\right]^{\frac{1}{2}}}\right\} \qquad \text{...(23)}$$

The expression in brackets is evaluated in Appendix 9.2 (equation (43)) and thus

$$F_2 = \frac{1}{\sqrt{\phi}}\exp(-\theta_2^2)\sqrt{-(q)^3}\times$$
$$\times\left[1+\sqrt{\pi}\{\tfrac{1}{2}q+\theta_2\}\exp(\tfrac{1}{2}q+\theta_2)^2\{1+\text{erf}(\tfrac{1}{2}q+\theta_2)\}\right] \qquad \text{......(24)}$$

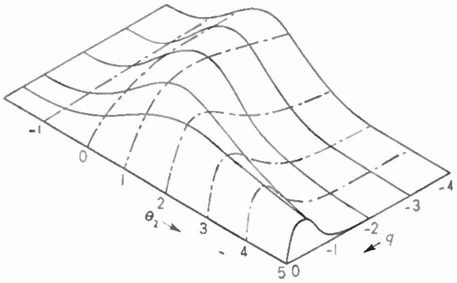

Fig. 10. Relief graph of the function $f_2(q, \theta_2)$.

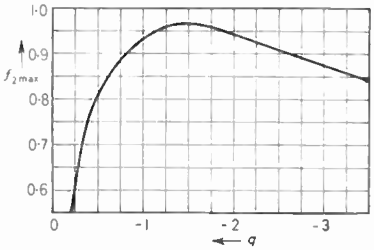

Fig. 11. Maximum of $f_2(q, \theta_2)_{(q\,=\,\text{const.})}$ as a function of q.

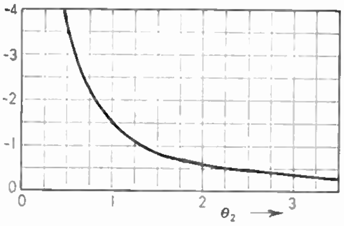

Fig. 12. θ_2 as a function of q for maximum values of $f_2(q, \theta_2)_{(q\,=\,\text{const.})}$.

which expressed in terms of the ideal improvement factor becomes

$$F_2 = F_I\sqrt[4]{\frac{2}{\pi}}\exp(-\theta_2^2)\sqrt{-(q)^3}\times$$
$$\times\left[1+\sqrt{\pi}\{\tfrac{1}{2}q+\theta_2\}\exp(\tfrac{1}{2}q+\theta_2)^2\{1+\text{erf}(\tfrac{1}{2}q+\theta_2)\}\right] \qquad \text{......(25)}$$

or

$$F_2 = F_I f_2(q,\theta_2) \qquad \text{......(26)}$$

The function $f_2(q, \theta_2)$ has been treated in the same manner as the function $f_1(q, \theta_1)$ and is found to behave in a similar fashion. Some idea of the general behaviour of both of these functions can be obtained from the relief of $f_2(q, \theta_2)$ shown in Fig. 10.

Figure 11 shows the maximum value attained by $f_2(q, \theta_2)_{(q\,=\,\text{const.})}$ as a function of q, and this corresponds to the view which is seen looking in a direction parallel to the θ_2 axis of Fig. 10. It will be seen that Fig. 11 has the same form as Fig. 5 but the maximum is shifted to a smaller value of q.

The curve relating θ_2 and q for maximum values of $f_2(q, \theta_2)_{(q\,=\,\text{const.})}$ is shown in Fig. 12 and we see that it is again of similar form to Fig. 6.

In the operation of the double integrator we find that the optimum value of q is $-1{\cdot}5$ and the 0·5 dB performance loss occurs for $q = -0{\cdot}87$ and $q = -2{\cdot}53$.

For N signals in the range previously taken we now find that the optimum value of β is given by

$$\beta_{\text{OPT}} = \exp\left(\frac{-1{\cdot}67\times1{\cdot}5}{N}\right) = \exp\left(\frac{-2{\cdot}5}{N}\right) \text{......(27)}$$

A number of values of β_{OPT} and also the values of β for 0·5 dB loss, have been calculated and in order that a comparison may be made with the values required for the single integrator these later results are also shown in Fig. 7. It will be seen that for the optimum integration of a given number of input signals the double integrator requires a smaller loop gain and is less sensitive to changes in loop gain. These are considerable virtues in many practical situations.

The remaining point in favour of the double integration system is the magnitude of the improvement obtained with optimum setting of the loop gains. The function $f_2(q, \theta_2)$ has an absolute maximum value of 0·9663 and thus the performance of the system is approximately only 0·3 dB below the ideal.

We have shown therefore that the double integration system will give a slightly better performance than the unweighted group integrator and is inherently continuous in its operation. As we have previously pointed out a *continuous* group integrator, even for a relatively small number of signals in the group, will be a complex and costly piece of apparatus, and although quantization may simplify the system somewhat it still results in an inferior performance. In comparison the double loop integrator is simpler and more efficient.

A double integrator does not necessarily need two delay lines but may be achieved using only one by utilizing two carrier frequencies, one for each delay. This method has the advantage of ensuring that the delays are equal but of course requires double the bandwidth of the single delay arrangement.

Although we have assumed identical loop gains in the two loops of the double integrator the analysis can be shown to be valid for loop gains which are similar but not identical. In this case it can be shown that the geometric mean of the two loop gains is the quantity which must correspond to the optimum obtained in the above analysis.

6. Conclusions

The performance of ideal and practical integrating systems for pulse echo-ranging has been examined on the basis of optimizing the ratio:

$$\frac{\text{change in mean level when signal applied}}{\text{r.m.s. of noise in the absence of signal}}$$

Since the number and amplitude distribution of the successive pulse signals available from a given target is usually determined by the beamwidth and angular response of the antenna beam, the discussion has been based on this, and for convenience the angular response—in terms of video signal/noise ratio—has been assumed to be of Gaussian form. The maximum improvement of the ratio shown above due to integration has been determined for the following integration systems:

(*a*) ideal weighting of old (re-circulated) information. (This is, of course, not a realizable system, but forms a standard for comparison).

(*b*) uniform weighting with the optimum number of signal pulses, the system input being cut off before and after. (This system is 0·5 dB poorer than (*a*)).

(*c*) single delay-loop integrator—a thoroughly practical system. (This system is 0·95 dB poorer than (*a*)).

(*d*) a proposed new double delay-loop integrator. (This system is only 0·3 dB poorer than (*a*)).

It is thus clear that the new system (*d*) is not only a practical one, but is also the best, giving only a very small loss compared with the ideal integrator. The actual improvement of the system over other practical systems is not perhaps very significant in itself, but the new system is shown to have some practical advantages. Thus a greater margin of loop stability is achieved by the use of a lower loop gain, and the more rapid final fall-off of the weighting function, when compared with that of the single loop system, should result in less target "smearing" when the integrator output is displayed on a plan position indicator.

It is obvious that the use of more than two loop integrators may be of value in certain practical situations but the signal/noise performance of the double loop system cannot be improved upon very much and it is suggested that more complex arrangements will be of interest when a very large number of waveforms is to be integrated, or when a very rapid fall off of the weighting function "tail" is desired.

7. Acknowledgments

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8. References

1. J. L. Lawson and G. E. Uhlenbeck, "Threshold Signals", M.I.T. Radiation Laboratory Series, Vol. 24. (McGraw-Hill, New York, 1950.)
2. J. W. R. Griffiths and I. G. Morgan, "The chemical recorder and its use in detecting pulse signals in noise", *Trans. Soc. Instrum. Tech.*, **8**, No. 2, p. 62, June 1956.
3. A. G. J. MacFarlane, "An analysis of a type of comb filter", *J. Instn Elect. Engrs*, **107**, Part B, No. 31, p. 39, January 1960.
4. G. L. F. Hinckley, "An experimental system for automatic radar target detection" (I.E.E. Paper No. 3248). Read at Symposium on Data Handling and Data Storage for Air Traffic Control, April 1960.
5. C. W. Helstrom, "Statistical Theory of Signal Detection". (Pergamon Press, London, 1960.)
6. R. E. Ellis and W. C. Rohlman, "Amplitude modulated video integrator", *Convention Record Inst. Radio Engrs*, **7**, Part 5, p. 263, March 1959.
7. D. G. Tucker and J. W. R. Griffiths, "Detection of pulse signals in noise", *Wireless Engineer*, **30**, p. 264, November 1953.
8. D. G. Tucker, "Detection of pulse signals in noise: trace-to-trace correlation in visual displays", *J.Brit.I.R.E.*, **17**, p. 319, June 1957.

8a. Discussion on reference 8, *J.Brit.I.R.E.*, **17**, p. 705, December 1957.

9. P. MacGregor, "A note on trace-to-trace correlation in visual displays", *J.Brit.I.R.E.*, **15**, p. 329, June 1955.

10. J. I. Marcum, "A Statistical Theory of Target Detection by Pulsed Radar". U.S.A.F. Project Rand Research Memoranda RM-753 and RM-754. July 1948.

10a. J. I. Marcum and P. Swerling, "Studies of target detection by pulsed radar", *Trans. Inst. Radio Engrs (Information Theory)*, **IT-6**, No. 2, April 1960 (Special Monograph Issue).

11. C. H. Baker, "Factors affecting radar operator efficiency", *J. Inst. Navigation*, **13**, No. 2, p. 148, April 1960.

12. H. Urkowitz, "Delay-line secondary responses in a.m. and f.m. sweep integrators", *J. Franklin Inst.*, **269**, p. 1, January 1960.

13. L. V. Blake, "The effective number of pulses per beamwidth for a scanning radar", *Proc. Inst. Radio Engrs*, **41**, p. 770, June 1953.

14. J. V. Harrington, "An analysis of the detection of repeated signals in noise by binary integration", *Trans. Inst. Radio Engrs (Information Theory)*, **IT-1**, No. 1, 1955.

9. Appendices

9.1. *Evaluation of the Signal/Noise Improvement Expression for the Single Integrator*

Evaluation of the expression

$$\frac{\int\limits_{-\infty}^{\theta_1} \beta^{(\theta_1-\theta)/\phi} \exp(-\theta^2)\,\mathrm{d}\theta}{\left[\int\limits_{-\infty}^{\theta_1} \beta^{2(\theta_1-\theta)/\phi}\,\mathrm{d}\theta\right]^{\frac{1}{2}}} \quad \text{for } \beta < 1 \qquad \text{......(28)}$$

We will evaluate the numerator first. Putting $(\theta_1 - \theta)/\phi = x$ we obtain a simpler form thus,

$$\int\limits_{-\infty}^{\theta_1} \beta^{(\theta_1-\theta)/\phi} \exp(-\theta^2)\,\mathrm{d}\theta$$

$$= -\phi \int\limits_{+\infty}^{0_1} \exp[x \log\beta - (\theta_1 - \phi x)^2]\,\mathrm{d}x$$

$$= \phi \exp(-\theta_1^2) \int\limits_{0}^{\infty} \exp[(\log\beta + 2\phi\,\theta_1)x - \phi^2 x^2]\,\mathrm{d}x$$

$$= \phi \exp(-\theta_1^2) \int\limits_{0}^{\infty} \exp(cx - \phi^2 x^2)\,\mathrm{d}x$$

where $\qquad c = \log\beta + 2\phi\,\theta_1 \qquad$ ......(29)

We can rewrite this as

$$\phi \exp(-\theta_1^2) \exp(c^2/4\phi^2) \int\limits_{0}^{\infty} \exp[-(\phi x - c/2\phi)^2]\,\mathrm{d}x \qquad \text{......(30)}$$

and putting $u = \phi x - c/2\phi$ we obtain

$$\exp(-\theta_1^2) \exp\left(\frac{c^2}{4\phi^2}\right) \int\limits_{-c/2\phi}^{\infty} \exp(-u^2)\,\mathrm{d}u$$

$$= \exp(-\theta_1^2) \exp\left(\frac{c^2}{4\phi^2}\right) \frac{\sqrt{\pi}}{2} \left[1 + \operatorname{erf}\frac{c}{2\phi}\right] \qquad \text{......(31)}$$

Now $\dfrac{c}{2\phi} = \dfrac{\log\beta}{2\phi} + \theta_1 = \dfrac{q}{2} + \theta_1$ where $q = \dfrac{\log\beta}{\phi}$. Thus we write the numerator of the expression in the form

$$\frac{\sqrt{\pi}}{2} \exp(-\theta_1^2) \exp(\tfrac{1}{2}q + \theta_1)^2 [1 + \operatorname{erf}(\tfrac{1}{2}q + \theta_1)] \qquad \text{......(32)}$$

The integral in the denominator can also be simplified by putting $x = (\theta_1 - \theta)/\phi$ and we obtain

$$\int\limits_{-\infty}^{\theta_1} \beta^{2(\theta_1-\theta)/\phi}\,\mathrm{d}\theta = -\phi \int\limits_{\infty}^{0} \beta^{2x}\,\mathrm{d}x$$

$$= -\frac{\phi}{2\log\beta} = \tfrac{1}{2} \cdot \frac{1}{(-q)} \qquad \text{......(33)}$$

where q is again $= \dfrac{\log\beta}{\phi}$.

Hence we finally obtain

$$\frac{\int\limits_{-\infty}^{\theta_1} \beta^{(\theta_1-\theta)/\phi} \exp(-\theta^2)\,\mathrm{d}\theta}{\left[\int\limits_{-\infty}^{\theta_1} \beta^{2(\theta_1-\theta)/\phi}\,\mathrm{d}\theta\right]^{\frac{1}{2}}}$$

$$= \sqrt{\frac{\pi}{2}} \exp(-\theta_1^2) \sqrt{-q} \exp(\tfrac{1}{2}q + \theta_1)^2 [1 + \operatorname{erf}(\tfrac{1}{2}q + \theta_1)] \qquad \text{......(34)}$$

9.2. *Evaluation of the Signal/Noise Improvement Expression for the Double Integrator*

Evaluation of the expression

$$\frac{\int\limits_{-\infty}^{\theta_2} (\theta_2 - \theta)/\phi\; \beta^{(\theta_2-\theta)/\phi} \exp(-\theta^2)\,\mathrm{d}\theta}{\left[\int\limits_{-\infty}^{\theta_2} \left(\frac{\theta_2 - \theta}{\phi}\right)^2 \beta^{2(\theta_2-\theta)/\phi}\,\mathrm{d}\theta\right]^{\frac{1}{2}}} \qquad \text{......(35)}$$

In order to simplify the numerator we put $y = (\theta_2 - \theta)/\phi$ and obtain

$$\int\limits_{-\infty}^{\theta_2} (\theta_2 - \theta)/\phi\; \beta^{(\theta_2-\theta)/\phi} \exp(-\theta^2)\,\mathrm{d}\theta$$

$$= -\phi \int\limits_{\infty}^{0} y \exp[y \log\beta - (\theta_2 - \phi y)^2]\,\mathrm{d}y$$

$$= \phi \exp(-\theta_2^2) \int\limits_{0}^{\infty} y \exp(gy - \phi^2 y^2)\,\mathrm{d}y$$

where $\qquad g = \log\beta + 2\phi\,\theta_2 \qquad$ ......(36)

Integrating by parts we find that

$$\int\limits_{0}^{\infty} y \exp(gy - \phi^2 y^2)\,\mathrm{d}y$$

$$= \frac{1}{2\phi^2} \left[1 + g \int\limits_{0}^{\infty} \exp(gy - \phi^2 y^2)\,\mathrm{d}y\right] \qquad \text{......(37)}$$

The integral in this expression is of the same form as that treated in appendix 9.1, and we can write

$$\int_0^\infty \exp(gy-\phi^2 y^2)\,\mathrm{d}y = \frac{1}{\phi}\exp\left(\frac{g^2}{4\phi^2}\right)\frac{\sqrt{\pi}}{2}\left(1+\operatorname{erf}\frac{g}{2\phi}\right) \qquad \text{......(38)}$$

The numerator of our expression therefore reduces to

$$\frac{1}{2\phi}\exp(-\theta_2^2)\left[1+\frac{g}{\phi}\exp\left(\frac{g^2}{4\phi^2}\right)\frac{\sqrt{\pi}}{2}\left(1+\operatorname{erf}\frac{g}{2\phi}\right)\right]$$

and putting $q = \dfrac{\log\beta}{\phi}$ ......(39)

and therefore $\dfrac{g}{2\phi} = \tfrac{1}{2}q+\theta_2$ we obtain

$$\frac{1}{2\phi}\exp(-\theta_2^2)\times$$
$$\times[1+(\tfrac{1}{2}q+\theta_2)\exp(\tfrac{1}{2}q+\theta_2)^2\sqrt{\pi}\,\{1+\operatorname{erf}(\tfrac{1}{2}q+\theta_2)\}] \qquad \text{......(40)}$$

In order to simplify the integral of the denominator we also put $y = (\theta_2 - \theta)/\phi$ and obtain

$$\int_{-\infty}^{\theta_2}\left(\frac{\theta_2-\theta}{\phi}\right)^2\beta^{2(\theta_2-\theta)/\phi}\,\mathrm{d}\theta = \phi\int_0^\infty y^2\exp(2y\log\beta)\,\mathrm{d}y \qquad \text{......(41)}$$

Repeated integration by parts gives us the result

$$\phi\int_0^\infty y^2\exp(2y\log\beta)\,\mathrm{d}y = -\frac{\phi}{4(\log\beta)^3} = \left(\frac{1}{2\phi}\right)^2\frac{1}{-(q^3)} \qquad \text{......(42)}$$

Hence we finally obtain

$$\frac{\displaystyle\int_{-\infty}^{\theta_2}(\theta_2-\theta)/\phi\;\beta^{(\theta_2-\theta)/\phi}\exp(-\theta^2)\,\mathrm{d}\theta}{\left[\displaystyle\int_{-\infty}^{\theta_2}\left(\frac{\theta_2-\theta}{\phi}\right)^2\beta^{2(\theta_2-\theta)/\phi}\,\mathrm{d}\theta\right]^{\frac{1}{2}}}$$
$$= \exp(-\theta_2^2)\sqrt{-(q^3)}\times$$
$$[1+\sqrt{\pi}(\tfrac{1}{2}q+\theta_2)\exp(\tfrac{1}{2}q+\theta_2)^2\{1+\operatorname{erf}(\tfrac{1}{2}q+\theta_2)\}] \qquad \text{......(43)}$$

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1961 Convention

"RADIO TECHNIQUES AND SPACE RESEARCH"—OXFORD, 5th–9th JULY

Synopses of Papers to be presented at the Convention

This is a further selection of some of the papers accepted for presentation during the Convention.

INTRODUCTORY SESSION: "RADIO TECHNIQUES IN SPACE RESEARCH"

Radio Tracking of Satellites

B. G. PRESSEY, PH.D. (*Radio Research Station, Department of Scientific and Industrial Research.*)

The introduction to the subject will include a brief discussion of the accuracy of position determination required for various purposes and of the optical and radio methods in use.

Radar and Doppler techniques will be described, but the main emphasis will be on the radio interferometer type of tracking system, which is the one most generally used at present. The design, operation and performance of such a system, with particular reference to the Minitrack station installed at the Radio Research Station, will be discussed in detail.

The Scientific Uses of Earth Satellites

J. H. BLYTHE, M.A., PH.D. (*Marconi's Research Laboratory.*)

A survey is given of the following topics, which have been selected for the extent to which they have been advanced by satellite research: the gravitational field and atmosphere of the earth; the ionosphere above the F2 layer maximum; magnetic fields in interplanetary space; and the zones of trapped radiation. The survey shows the value of satellite research, but limitations due to the nature of orbits are also pointed out.

SESSIONS ON "SATELLITE ENGINEERING"

An Economical and Timely Technique for Conducting Radio Research in Space

JOHN D. NICOLAIDES, M.S.E. (*U.S. Bureau of Naval Weapons.*)

Experimental radio investigation in space is now limited because of the exorbitant cost and the unavailability of satellite and probe launching vehicles. An uninformed efficiency criteria of dollars/pound in orbit serves only to accelerate vehicle development towards even greater size and cost.

Over the last three years, however, the U.S. Navy has developed a simple and inexpensive launching technique for economical satellite radio research. This technique employs an aircraft and rocket combination which can place a 20-pound satellite into orbit. The purpose of this paper will be to review the progress to date and to indicate performance potential.

Some Thermal Considerations on the Use of Silicon Solar Cells in Earth Satellites

R. P. HOWSON, B.SC., D. H. ROBERTS, B.SC., AND B. L. H. WILSON, M.A. (*Plessey Company, Caswell Research Laboratories, Towcester.*)

An important problem in the design of a solar cell power unit is concerned with the possibility of providing a suitable surface treatment to alter the operating temperature of the device. Equations for the heat balance of a plate are used to derive expressions for the maximum and minimum temperatures reached during an orbit round the earth. It is shown that excessive mass would be needed to increase the thermal time constant if specific heat alone is used to act as a sink for the thermal energy. However, it is possible to use a paddle construction in the form of a hollow aluminium canister containing a material of high latent heat (e.g. a fatty acid) which melts at a temperature slightly above the mean temperature that would be reached if the cells were mounted on solid aluminium. The temperature of the complete paddle will then remain sensibly constant at the melting temperature over a large part of the orbiting period.

Solar Cells for Communication Satellites in the Van Allen Belt

F. M. SMITS, PH.D., K. D. SMITH AND W. L. BROWN. (*Bell Telephone Laboratories.*)

The paper describes the effects of the radiation environment on the expected life of solar cells and the various considerations dictating the design of solar cells for use in the Van Allen belt. The design of solar cells especially developed for an active communication satellite is described and bombardment studies of these and alternate cells are reported. It is shown that long life satellites are feasible using *n*–on–*p* solar cells now available.

SESSION ON "EXTRA-TERRESTRIAL MEASUREMENTS"

The Use of Probing Electrodes in the Study of the Ionosphere

R. L. F. BOYD, PH.D. (*University College, London*)

The Langmuir probe and its developments provide a means whereby such important ionospheric quantities as electron concentration and temperature, and ion mass spectrum, concentration and temperature may be measured. These quantities are of fundamental importance in any consideration of the ion and electron equilibrium in the ionosphere.

The problems involved arise largely because the vehicle is isolated from earth and because of the varying aspect of the vehicle. These problems require very careful study before any instrumentation is built and thus the solution involves the use of special techniques which are not needed in the laboratory.

The ways in which the problems have been tackled will be discussed and special reference will be made to the instrumentation planned for the first international *Scout* satellite.

Cosmic Ray Measurements in the U.S. Scout I Satellite

PROF. H. ELLIOTT, M.SC., PH.D., J. J. QUENBY, PH.D. AND A. C. DURNEY, B.SC. (*Imperial College, Department of Physics*) AND D. W. MAYNE (*McMichael Radio Limited.*)

The main purpose of this experiment is to make accurate measurements of the primary cosmic ray energy spectrum and of the way in which this spectrum changes as a result of modulation by the interplanetary magnetic field. The cosmic ray intensity will be measured as a function of latitude and longitude using Geiger-Muller counters, which respond to the total flux of primary particles, and the Cerenkov counter which responds only to the heavy primaries with a charge of six or greater.

This paper will give an outline of the physical background to the experiment, together with a description of the sensor and its associated electronics.

Measurements of Solar X-Radiation

K. A. POUNDS, PH.D. (*University of Leicester, Department of Physics.*)

The considerable flux of X-radiation, radiated from the solar corona, is absorbed in the earth's atmosphere between 80–130 km. Accurate measurement of this flux is important for the further understanding of both the emitting regions of the Sun and of the earth's upper atmosphere. Of particular interest in both respects is the strength *and* spectral distribution near the short-wavelength limit of the solar X-ray spectrum. Two methods developed for the study of this emission with the aid of rocket or satellite vehicles are described. In the first method, a simple photographic detection method is used and an equipment recovery technique is employed to collect flight data. This experiment is an example of the work that may be done with very little ancillary rocket equipment. The second experiment, however, goes to the other extreme and involves a high degree of data processing in the actual vehicle. The equipment is a Proportional Counter X-Ray Spectrometer and rocket and satellite models are described and their experimental scope examined. Practical limitations, imposed in part by the detector employed, partly by the electronics system and partly by the type of vehicle available are discussed.

The Defence Research Board Topside Sounder Satellite

J. C. W. SCOTT. (*Defence Research Board of Canada.*)

The paper will describe the satellite now being constructed at the Telecommunications Establishment for launching early in 1962. The purpose of this satellite is to measure the regions of the ionosphere above the F region maximum with the use of an ionosonde carried in the satellite.

Some Synchronous Observations of Satellite Transmissions at Spaced Receiving Stations

F. A. KITCHEN, B.SC., R. N. GOULD, W. R. R. JOY, B.SC.(ENG.) AND W. R. CARTER. (*Admiralty Surface Weapons Establishment.*)

Synchronous measurements of the Doppler frequency shifts associated with the 20·005 Mc/s transmission from the Russian artificial earth satellite 1958 Delta 2 were made at Banbury, Portsdown and Great Baddow. Analysis of the results obtained of a key observing transit (revolution No. 3404) was carried out in detail. Comparison was made between the frequency/time curve calculated from the observed data and a theoretical computation of the Doppler shift. The irregularities were found to be of two kinds. The first was a rapid irregular fluctuation of a few cycles which was superimposed on a second variation of larger amplitude with a period of over a minute. These effects are discussed and their interpretation is considered.

SESSIONS ON "COMMUNICATION SATELLITES"

Optimum System Engineering for Satellite Communication Links

W. L. WRIGHT, B.A., and S. A. W. JOLLIFFE. (*Marconi's Research Laboratory.*)

The conservation of signal and the reduction of noise and interference dominate the satellite communication problem. For this reason it is important to design so that maximum use is made of the available power and bandwidth, with a view to obtaining a system with an economically acceptable information capacity. Factors which determine system capacity are discussed. Special circuit techniques are described including low-noise pre-amplifier design, frequency following and phase-locked receivers associated with the ground terminal and special repeater equipment for the satellite.

Consideration of modulation methods and information processing techniques, including s.s.b. wideband f.m. and pulse code methods, leads to the conclusion that the latter possesses distinct advantages and more nearly approaches the ideal expressed by the Hartley-Shannon law. Examples are given of a 600 telephone channel satellite system and of bandwidth compression techniques studied in connection with an astronomical satellite.

A Proposal for an Active Communication Satellite System based on Inclined Elliptic Orbits

B. BUSS AND J. R. MILLBURN, B.SC., GRADUATE. (*Hawker Siddeley Aviation Limited, Advanced Projects Group.*)

For a "first generation" communication satellite system, primarily intended to give a service to the Northern Hemisphere in general and the North Atlantic link in particular, there is much to recommend the use of elliptic orbits in the 63° slot. This paper describes the design of a multichannel telephony satellite which is suitable for use in this type of orbit.

In the absence of specific frequency allocations for earth-space-earth communications, it is necessary to examine the question of operating frequency not only with regard to optimum performance, but also with regard to minimum interference to or from other services. It is shown that interference considerations lead to the use of frequencies in the 2–6 kMc/s region. Consideration of various types of repeaters and solar cell arrays leads to a design based on a semi-sun-seeking concept, in which the satellite is stabilized in two axes so that its aerials point towards the earth while rotation about the third axis is used to maximize the output from the solar cells. It is shown that a satellite of this type, weighing 550 lb, would have a capacity of up to 1000 duplex telephone circuits of good quality.

Long-Distance Communications via the Moon

P. A. WEBSTER. (*Pye Telecommunications Ltd.*)

The paper begins by outlining the history of a series of tests of signals reflected from the moon using the Jodrell Bank radio-telescope carried out in collaboration with the University of Manchester. Three experiments made using transmitters and receivers in the v.h.f. band are described. The first test was with an f.m. transmitter of 1 kW and a standard commercial receiver. Intelligible speech signals were received back at Jodrell Bank and c.w. signals were received in the U.S.A. A second test using amplitude modulation yielded results with improved intelligibility and allowed further measurements of the moon's effective bandwidth and other factors to be made. The third test also employed the a.m. transmitter, but used a teleprinter keyed frequency shift sub-carrier of 1620 c/s. On this occasion some of the transmission was received on a 60-ft. diameter aerial in Sydney, Australia.

The paper then discusses the nature of the reflected signals and the limitations this imposes on the utilization of the path. The economics involved in the use of the moon as a reflector are dealt with, together with consideration of the optimum frequency, transmitter power and aerial size. To conclude the paper, some examples of the extent of useful operating hours between different parts of the world are given.

SESSION ON "TECHNIQUES IN RADIO ASTRONOMY"

A Satellite Technique for Performing a High Resolution Survey of the Radio Sky at Medium Wavelengths

R. C. JENNISON, PH.D. (*University of Manchester, Nuffield Radio Astronomy Laboratories, Jodrell Bank.*)

Highly directional aerials at frequencies of the order of 1 Mc/s cannot at present be carried in space vehicles. This paper draws attention to techniques whereby the properties of the ionosphere combined with a simple Hertzian dipole on the vehicle may be used to achieve very high angular resolutions.

The Sessions to which the above papers are allocated are provisional and will be confirmed in the final programme which will appear in the "Convention Preview Issue" of the Journal *to be published next month. A further selection of synopses will also be given in the June* Journal.

U.D.C. 517.945.2:621.382.3

The Calculus of Deviations Applied to Transistor Circuit and Network Analysis

By

T. R. NISBET (*Associate Member*)†

AND

W. W. HAPP, Ph.D.†

Summary: The Calculus of Deviations is introduced, and its nature and basis are briefly discussed. A deviation is described as "a measure of the lack of complete functional interdependence between two quantities". Rules are given for using the calculus in solving various transistor and network problems, including the analysis of distributed-parameter networks. The complexity of network analysis is very greatly reduced by the use of this powerful, easy-to-use, mathematical tool.

1. Introduction

Historically, transistors were first described by the "device parameters", r_b, r_e and α; these quantities were attractive in that they tied in fairly well with the physicist's model of the transistor. They could not be measured directly, however, and the well-established methods of network analysis[1] were borrowed to enable their values to be computed. In the preface to his 1953 book,[2] R. F. Shea comments: "Currently there is considerable debate over the relative merits of specifying the equivalent circuits in terms of the open-circuit impedances, short-circuit admittances, or other parameters, such as the *h*-factors which involve impedance, admittance, and current and voltage ratios".

The next step was gradually to accept the network theorist's equivalent circuit as well as his terminology. The circuit engineer was not handicapped by using an equivalent circuit unrelated to the physics of transistor action, and the *z*-parameters (sometimes referred to as the *R*-parameters) gained popularity.

The convenience of the hybrid or *h*-parameters, both for measurement and description purposes, gave them a distinct advantage, and transistor data sheets of the 1955–57 period show their increasing acceptance by circuit engineers.

Nowadays, we have six conventional sets of circuit parameters, known by the symbols *a*, *b*, *g*, *h*, *y*, and *z*. The network analyst's victory is complete, and the transistor circuit designer has the advantage of more than two decades of work in network analysis.

Minor improvements have come along with the major ones. Suffixes *i*, *r*, *f* and *o* have replaced the more cumbersome 11, 12, 21 and 22 notation. Capital letters have come to indicate large-signal or d.c. parameters, while small letters indicate small-signal parameters.

The subject of this paper may be considered as another such minor improvement. Superficially, such may be its effect. One may, on the other hand, regard the technique of using deviations as a revolutionary innovation which streamlines all network and transistor calculations by placing a powerful new mathematical tool in the hands of the circuit designer. Deviations‡ however are not altogether new; they are akin to the mean and standard deviations of statistics, to the influence coefficients of aerodynamics and to the Jacobians of thermodynamics.[3] Their application to electrical network theory is a logical, if belated, step towards flexibility in the use of circuit mathematics.

2. Definitions

Informal as well as formal definitions can be given. A deviation can be likened to the two or the three in the fraction 2/3: by itself, neither number has any significance, but when they are combined in the form of a ratio, a very specific meaning emerges. The meaning is the same if the 2 is replaced by $2/t$ and the 3 by $3/t$.

The numerator and denominator of a partial derivative can be separated by the introduction of a new parameter, *t*, *which need never be specified.*

$$\left(\frac{\partial x}{\partial u}\right)_y = \frac{(\partial x/\partial t)_y}{(\partial u/\partial t)_y} \underline{\Delta} \frac{(x, y)}{(u, y)} \qquad \text{......(1)}$$

(symbol $\underline{\Delta}$ denotes "by definition"). If all the partial derivatives of a system are referred to the same variable, *t*, and if only ratios are considered, then *t* can be dispensed with and the numerator written as (x, y) and the denominator as (u, y). The symbol (x, y) is called a *deviation* of *x* and *y*, and the rules governing deviations—known as the calculus of deviations—are outlined below.

An alternative definition of the deviation of two quantities results from consideration of the *Jacobian*

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‡ In previous work by the authors, the words "polynomials" and "Jacobians" have been used for what are here described as "deviations", the latter now being thought to be the most appropriate term.

determinant

$$\begin{vmatrix} \left(\frac{\partial x}{\partial u}\right)_v & \left(\frac{\partial x}{\partial v}\right)_u \\ \left(\frac{\partial y}{\partial u}\right)_v & \left(\frac{\partial y}{\partial v}\right)_u \end{vmatrix} = \frac{\partial(x,y)}{\partial(u,v)} \triangleq \frac{(x,y)}{(u,v)} \qquad \ldots\ldots(2)$$

If $v = y$, then eqn. (2) simplifies to eqn. (1) correlating the two definitions. If (u, v) is non-zero, interchange of x and y in eqn. (2) yields the important *anti-commutation rule*,

$$(x, y) = -(y, x) \qquad \ldots\ldots(3)$$

and the corollary

$$(x, x) = 0 \qquad \ldots\ldots(4)$$

which has an interesting interpretation in that *x does not deviate from x.*

Further, for any function $f = f(x)$ which depends only on x,

$$(f, x) = 0 \quad \text{since} \left(\frac{\partial f}{\partial t}\right)_x = 0 \qquad \ldots\ldots(5)$$

Finally, if (1) is substituted into (2), we have

$$\begin{vmatrix} (x,v) & (x,u) \\ (y,v) & (y,u) \end{vmatrix} = (x,y)(u,v) \qquad \ldots\ldots(6)$$

which becomes

$$(x,y)(u,v)+(x,u)(v,y)+(x,v)(y,u) = 0 \qquad \ldots\ldots(7)$$

This basic relationship relates four variables of a system, say i_i, i_o, v_i and v_o, the voltages and currents of a two-port as in Fig. 1. Any two of the variables may be independent variables while the remaining two are dependent, e.g.:

$$i_i = f(v_i, v_o) \quad \text{and} \quad v_o = f(v_i, i_o) \qquad \ldots\ldots(8)$$

Differentiating i_i with respect to t and keeping i_o constant yields:

$$(\partial i_i/\partial t)_{i_o} = (\partial i_i/\partial v_i)_{v_o}(\partial v_i/\partial t)_{i_o} + (\partial i_i/\partial vo)_{v_i}(\partial v_o/\partial t)_{i_o}$$

or

$$(i_i, i_o)(v_i, v_o) = (i_i, v_o)(v_i, i_o) - (i_i, v_i)(v_o, i_o) \ldots\ldots(9)$$

Thus eqn. (9) provides a physical interpretation of eqn. (2), namely: If two variables of a two-port are given (say i_i and v_o) the other two variables (v_i and i_o) are thereby uniquely determined.

3. Application to Network Theory

Rules are given below, which govern the manipulation of deviations in network problems. All are applicable to both active and passive networks.

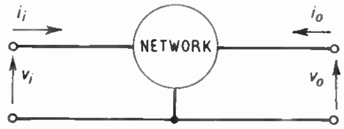
3.1. *Terminology*

The representation of network voltages and currents is as in Fig. 1 so that for example, the standard parameter, h_i, conventionally defined as $(\partial v_i/\partial i_i)v_o$, becomes in deviation notation

$$h_i = \frac{(v_i, v_o)}{(i_i, v_o)} \qquad \ldots\ldots(10)$$

It is convenient to choose a single letter to represent the different deviations, and for reasons which will appear, (see Section 4), the single letters are the same as those used in the standard parameters, bracketed to signify that they are deviations.

Fig. 1.



The single-letter representation is as follows:

Deviation	Denoted by
(v_o, i_o)	(a)
(i_i, v_i)	(b)
(v_i, i_o)	(g)
(i_i, v_o)	(h)
(v_i, v_o)	(y)
(i_i, i_o)	(z)

......(11)

An additional symbol is added to this list for later convenience

$$\begin{aligned}(v_o - v_i, i_o + i_i) &= (f) \\ &= (a)+(b)-(g)-(h)\end{aligned}$$

3.2. *Rules*

3.2.1. Uniqueness condition

If, when the network and two variables are known, the remaining two variables are uniquely determined, then the six deviations resulting from these four variables are interrelated by the uniqueness condition:

$$(a)(b)-(g)(h)+(y)(z) = 0 \qquad \ldots\ldots(12)$$

3.2.2. Associative law

Individual variables in a deviation can combine or be combined to form new deviations, thus

$$(i_i, v_i)+(i_i, v_o) = (i_i, v_i + v_o) \qquad \ldots\ldots(13)$$

3.2.3. Anti-commutation law

Reversal of the sequence of two variables reverses the sign of the deviation, thus

$$(v_i, v_o) = -(v_o, v_i) \qquad \ldots\ldots(14)$$

3.2.4. Multiplier rule

If one of the variables in a deviation is multiplied by a constant, the deviation itself is multiplied by the same constant. Thus,

$$(v_i, c\,v_o) = (c\,v_i, v_o) = c(v_i, v_o) \qquad \ldots\ldots(15)$$

$$(v_i, -v_o) = (-v_i, v_o) = -(v_i, v_o) \qquad \ldots\ldots(16)$$

3.2.5. Degree of freedom

The six deviations which make up a set occur only as ratios, never singly. Under this limitation, any *one* deviation of a set can be given an arbitrary numerical value, frequently 1, and this determines numerically the values of all the deviations of the set.

3.2.6. Deviation zero

The deviation of a quantity with respect to itself or to a constant is zero.

$$(v_i, v_i) = 0 \qquad \ldots\ldots(17)$$

$$(v_i, c) = 0 \qquad \ldots\ldots(18)$$

Conceptual importance of deviation zero. It was shown earlier that the deviation of two variables is zero if they are functionally completely dependent on each other (eqn. (5)). In statistics, such dependence is described by mean and standard deviations, and it is therefore reasonable to speak of a quantity such as (x,y) as the deviation of x and y, and of the rules of calculation as the *calculus of deviations.*†

There is a striking correspondence between the calculus concepts portrayed by focusing attention respectively on deviations and derivatives. In the calculus of derivatives $(\partial x/\partial y)_z$ becomes $\mathrm{d}x/\mathrm{d}y$ if all the derivatives of the system have the same z. In the calculus of deviations $(\partial x/\partial y)_z$ becomes (x,z) if all deviations of the system have the same y. From this viewpoint, the fact that $(z,z)=0$ is obviously true, and eqns. (13), (14) and (16) follow at once from this equality.

If $z=x+y$, then by expanding the deviation,

$$(x+y, x+y) = (x, y)+(y, x) = 0 \quad \ldots\ldots(19)$$

Clearly, eqn. (14) is a corollary of eqn. (17).

Another condition where deviation zero is encountered is when one term is a constant. Since $\partial(x+k)=\partial x$, it follows that $(x+k,y)=(x,y)$, whence

$$(k, y) = 0 \qquad \ldots\ldots(20)$$

In a manner analogous to eqn. (17) it can be shown that $(k,y)=-(k,y)$, and we can deduce a rule that *any deviation which includes a constant as one of its terms is zero.*

Conceptually, "complete functional interdependence" exists between any two quantities which can be plotted against each other on a graph, and since a constant and a variable can be so plotted, it is quite reasonable to postulate the existence of "complete functional interdependence" between them—giving $(k,x)=0$ as a special case of $(y,x)=0$ where $y=f(x)$. Conversely, a "lack of complete functional interdependence" is present when two quantities cannot be plotted against each other, *and the measure of this lack of functional interdependence is the deviation.*

4. Application to Transistor Parameters

4.1. *Conversion of Transistor Parameters*

The conventional transistor parameters are given in Table 1, with the deviation notation. For example,

$$h_i = \frac{\partial v_i}{\partial i_i}\,v_o = \frac{(v_i, i_o)}{(i_i, v_o)} = \frac{(y)}{(h)} \qquad \ldots\ldots(21)$$

All the h-parameters have the deviation (h) as their denominator (Table 1), and in a numerical evaluation it is convenient to let $(h)=1$, so that numerical values are immediately available for the complete set of deviations.

To convert to the z-parameters, the denominator (z) is required. Thus,

$$z_i = \frac{(g)}{(z)} = \frac{(g)/(h)}{(z)/(h)} = \frac{\Delta^h}{h_o} \qquad \ldots\ldots(22)$$

The use of Table 1 therefore enables any parameter to be written down at a glance, in terms of the parameters which are known.[6]

4.2. *Rotation and Interchange of Terminals*

Conversion from one circuit orientation to another is a simple manipulation of deviations. If it is desired to convert the common base parameters to common emitter, it is necessary only to manipulate the devia-

Table 1

Network Parameters in Deviation Notation

Network parameters	Subscripts				Determinant
	11 or *i*	12 or *r*	21 or *f*	22 or *o*	
a	$\frac{(g)}{(a)}$	$\frac{(y)}{(a)}$	$\frac{(z)}{(a)}$	$\frac{(h)}{(a)}$	$\Delta^a = \frac{(b)}{(a)}$
b	$\frac{(h)}{(b)}$	$\frac{(y)}{(b)}$	$\frac{(z)}{(b)}$	$\frac{(g)}{(b)}$	$\Delta^b = \frac{(a)}{(b)}$
g	$\frac{(z)}{(g)}$	$\frac{(b)}{(g)}$	$\frac{(a)}{(g)}$	$\frac{(y)}{(g)}$	$\Delta^g = \frac{(h)}{(g)}$
h	$\frac{(y)}{(h)}$	$\frac{(b)}{(h)}$	$-\frac{(a)}{(h)}$	$\frac{(z)}{(h)}$	$\Delta^h = \frac{(g)}{(h)}$
y	$\frac{(h)}{(y)}$	$-\frac{(b)}{(y)}$	$-\frac{(a)}{(y)}$	$\frac{(g)}{(y)}$	$\Delta^y = \frac{(z)}{(y)}$
z	$\frac{(g)}{(z)}$	$\frac{(b)}{(z)}$	$\frac{(a)}{(z)}$	$\frac{(h)}{(z)}$	$\Delta^z = \frac{(y)}{(z)}$

---

† Second order deviations, such as $((x,y), (x,z))$ can also be found[4] and in fact the process of taking the deviation can be repeated as often as required, leading to a whole series of mathematical methods which together form a calculus.

(a) (b)

Fig. 2.

tions, one at a time, from eqn. (11). Using a subscript to denote the circuit configuration, the deviation (y_E) is seen to be (v_{iE}, v_{oE}). Now v_{iE} is the voltage between base and emitter, or $-v_{iB}$, and v_{oE} is $v_{oB}-v_{iB}$. Using the rules of Section 3, we have

$$(y_E) = (-v_{iB}, v_{oB}-v_{iB}) = (-v_{iB}, v_{oB}) + (-v_{iB}, -v_{iB})$$
$$= -(y_B)$$

Similar manipulations yield simple conversion formulas for the remaining CE deviations in terms of CB deviations. The results for all possible conversions are given in Table 2.

With six sets of parameters in each of three circuit configurations, it would require 72 individual conversion tables to provide the same information as is contained in Tables 1 and 2. The tables can be used for active or passive networks, with real and imaginary components (or magnitude and phase) in all terms.

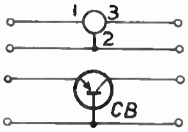
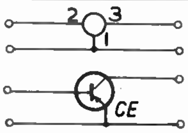
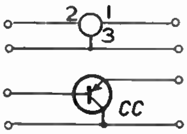
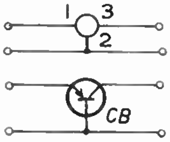
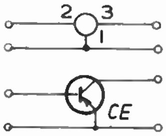
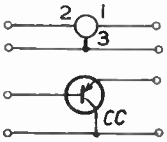
5. Asymptotes of Design Formulae

5.1. *Basic Concepts*

Deviations find a useful application in computing the asymptotes of most of the design graphs used in

Table 2

Deviations of Rotated Networks

Orientation described ↓	IN TERMS OF		
			
	a_B b_B g_B h_B y_B z_B f_B	a_E-g_E b_E-g_E $-g_E$ f_E $-y_E$ $-z_E$ h_E	$-b_C+g_C$ $-a_C+g_C$ $-f_C$ g_C y_C z_C $-h_C$
	a_B-g_B b_B-g_B $-g_B$ f_B $-y_B$ $-z_B$ h_B	a_E b_E g_E h_E y_E z_E f_E	a_C-h_C b_C-h_C f_C $-h_C$ $-y_C$ $-z_C$ $-g_C$
	$-b_B+h_B$ $-a_B+h_B$ h_B $-f_B$ y_B z_B $-g_B$	a_E-h_E b_E-h_E f_E $-h_E$ $-y_E$ $-z_E$ g_E	a_C b_C g_C h_C y_C z_C f_C
Note: $f = a+b-g-h$			

All symbols are deviations—brackets omitted.

Supplement to

"Library Services and Technical Information for The Radio and Electronic Engineer"*

ACCESSIONS TO THE LIBRARY

030.8 Encyclopedias, Dictionaries, etc.

CARTER, Harley. Dictionary of Electronics. *George Newnes Ltd.*, 1960, 377 pp.

COOKE, Nelson M., MARKUS, John. Electronics and Nucleonics Dictionary. *McGraw Hill*, 1960, 543 pp.

PAGET MOFFATT, C. W. Science French Course. *University Tutorial Press Ltd.*, 4th edn, 1951, 332 pp.

PAGET MOFFATT, C. W. Science German Course. *University Tutorial Press Ltd.*, 5th edn, 1952, 328 pp.

371.048 Vocational Guidance

COCHRANE, Louise. Anne in Electronics. *Chatto & Windus*, 1960, 140 pp. (A technical career novel for girls.)

The Directory of Opportunities for School Leavers, 1960. *Cornmarket Press Ltd.*, 212 pp.

51 Mathematics

ALGER, Philip L. Mathematics for Science & Engineering. *McGraw Hill*, 1957, 360 pp.

517 Calculus

DAY, W. D. Introduction to Laplace Transforms for Radio and Electronic Engineers. *Iliffe*, 1960, 183 pp.

GOURSAT, Edouard. A Course in Mathematical Analysis. Vol. 1.—Applications to Geometry. *Dover*, 1959, 584 pp. Translated by E. R. Hedrick.

GOURSAT, Edouard. A Course in Mathematical Analysis. Vol. 2. Part 1. Functions of a Complex Variable. *Dover*, 1959, 259 pp. Translated by E. R. Hedrick and Otto Dunkel.

GOURSAT, Edouard. A Course in Mathematical Analysis. Vol. 2. Part 2. Differential Equations. *Dover*, 1959, 300 pp. Translated by E. R. Hedrick and Otto Dunkel.

PIERPONT, J. The Theory of Functions of Real Variables. Vol. 1. *Dover*, 1959, 560 pp.

PIERPONT, J. The Theory of Functions of Real Variables. Vol. 2. *Dover*, 1959, 645 pp.

PIERPONT, J. Functions of a Complex Variable. *Dover*, 1959, 583 pp.

WEBB, H. A., ASHWELL, D. E. Mathematical Tool Kit for Engineers. *Longmans*, 2nd edn, 1959, 116 pp.

525.7 Terrestrial Atmosphere

BUDDEN, K. G. Radio Waves in the Ionosphere. *Cambridge University Press*, 1961, 542 pp.

53 Physics

BLANCHARD, C. H., and others. Introduction to Modern Physics. *Pitman*, 1960, 420 pp.

530.145 Quantum Theory

HOUSTON, William V. Principles of Quantum Mechanics. *Dover*, 1959, 288 pp.

534 Acoustics

BICKLEY, W. G., TALBOT, A. An Introduction to the Theory of Vibrating Systems. *Oxford at the Clarendon Press*, 1961, 238 pp.

FRANSSEN, N. V. Some Considerations on the Mechanism of Directional Hearing. *Technical High School, Delft*, 1960, 115 pp.

536 Heat

PLANCK, Max. The Theory of Heat Radiation. *Dover*, 1959, 244 pp.

PORTER, A. W. Thermodynamics. *Methuen Monographs*, 4th edn, 1951, 124 pp.

537 Electricity

MOON, P., SPENCER, D. E. Foundations of Electrodynamics. *Van Nostrand*, 1961, 314 pp.

MUTER, Ivan L. The "*Advance*" Science Masters Handbook. *Advance Components Ltd.*, 1960, 124 pp.

612 Physiology

SMYTH, C. N. (editor) Medical Electronics. *Iliffe*, 1960, 614 pp. Proceedings of the 2nd International Conference, Paris 1959.

STACY, Ralph W. Biological and Medical Electronics. *McGraw Hill*, 1960, 308 pp.

621-52 Control

SOLODOVNIKOV, V. V. Introduction to the Statistical Dynamics of Automatic Control Systems. *Dover*, 1960, 307 pp. Translated from the Russian.

WEST, John C. Analytical Techniques for Non-Linear Control Systems. *E.U.P.*, 1960, 223 pp.

WILTS, C. H. Principles of Feedback Control. *Addison-Wesley Publishing Co. Inc.*, 1960, 271 pp.

621 Mechanical Engineering

HOGARTH, C. A., BLITZ, J. (editors) Techniques on Non-Destructive Testing. *Butterworths*, 1960, 216 pp.

621.3 Electrical Engineering

Basic Electricity. *Technical Press*, 1959, issued in 5 parts.

CORNELIUS, P. Electrical Theory. *Cleaver-Hume Press Ltd.*, 1961, 187 pp.

621.3.012.1 Vector Analysis

KRON, Gabriel. Tensors for Circuits. *Dover*, 1959, 250 pp.

---

* Published by The British Institution of Radio Engineers, 9 Bedford Square, London, W.C.1. Price 2s. 6d.

621.314.7 Transistors and Semiconductors

AMOS, S. W. Principles of Transistor Circuits. *Iliffe Books Ltd.*, 2nd edn, 1961, 210 pp.

CARROLL, John M. Modern Transistor Circuits. *McGraw Hill*, 1959, 268 pp.

CATTERMOLE, K. W. Transistor Circuits. *Heywood & Co.*, 1959, 442 pp.

FITCHEN, F. C. Transistor Circuit Analysis and Design. *Van Nostrand*, 1960, 356 pp.

GARTNER, Wolfgang W. Transistors: Principles, Design and Applications. *Van Nostrand*, 1960, 675 pp.

Mullard Reference Manual of Transistor Circuits. *Mullard Ltd.*, 1960, 308 pp.

Semiconductor Application Report. Vol. 1. No. 4, D.C. Power Supply Circuits Using Silicon Rectifiers. *Texas Instruments Ltd.*, 1959, 35 pp.

SJOBBEMA, D. J. W. Using Transistors. *Philips' Technical Library*, 1960, 110 pp.

621.315.2 Cables

DUMMER, G. W. A., BLACKBAND, W. T. Wires and R.F. Cables. Radio and Electronic Components. Vol. 5 *Pitman*, 1961, 240 pp.

621.317.755 Oscilloscopes

CARTER, Harley. An Introduction to the Cathode Ray Oscilloscope. *Philips' Technical Library*, 1960, 2nd edn, 121 pp.

621.318.1 Magnetic Materials

ROBERTS, J. High Frequency Applications of Ferrites. *English Universities Press*, 1960, 166 pp.

621.319.4 Capacitors

MARK, David. R-L-C Components Handbook. *Rider*, 1959, 146 pp.

621.351 Batteries

Proceedings of the Second International Symposium on Batteries, Bournemouth, 1960. *Inter-Departmental Committee on Batteries*, 314 pp.

621.365.5 Induction Heating

SIMPSON, P. G. Coil and System Design. *McGraw Hill*, 1960, 295 pp.

621.37 Circuits

HEIJN, H. J. Representations of Switching Functions and their Application to Computers. *Technical High School, Delft*, 1960, 79 pp.

SCOTT, R. E. Linear Circuits. Part 1. Time Domain Analysis. *Addison-Wesley Publishing Co.*, 1960, 510 pp.

SCOTT, R. E. Linear Circuits. Part 2. Frequency-Domain Analysis. *Addison-Wesley Publishing Co.*, 1960, 418 pp.

621.371 Transmission

MOORE, Richard K. Travelling-Wave Engineering. *McGraw Hill*, 1960, 360 pp.

SLATER, J. C. Microwave Transmission. *Dover*, 1959, 309 pp.

621.372.8 Waveguides

YOUNG, Victor, J. Understanding Microwaves. *John F. Rider*, 1960, 292 pp.

621.374 Pulses

MILLMAN, Jacob, TAUB, Herbert. Pulse and Digital Circuits. *McGraw Hill*, 1956, 687 pp.

NEETESON, P. A. Analysis of Bistable Multivibrator Operation. *Technical High School, Delft*, 1956, 82 pp.

STRAUSS, Leonard. Wave Generation and Shaping. *McGraw Hill*, 1960, 520 pp.

621.375 Vacuum Tube Amplifiers

RODENHUIS, E. Hi-Fi Amplifier Circuits. *Philips' Technical Library*, 1960, 105 pp.

VUYLSTEKE, Arthur A. Elements of Maser Theory. *Van Nostrand*, 1960, 362 pp.

621.376 Modulation

KNUDTZON, Nic. Theoretical Comparison of Fluctuation Noise Characteristics for Various Modulation Methods with Special View to Multi-channel System Planning. *Technical High School, Delft*, 1957, 261 pp.

621.38 Electronics

DUMMER, G. W. A., GRIFFIN, N. Electronic Equipment Reliability. *Pitman*, 1960, 274 pp.

Basic Electronics. *Technical Press*, 1959, issued in 6 parts.

621.385.1 Valves

MIDDLETON, H. A. Master Receiving-Picture Tube Substitution Guide Book. *Rider*, 1959, 343 pp.

621.385.38 Thyratrons

SWENNE, C. M. Thyratrons. *Philips' Technical Library*, 1960, 73 pp.

621.389 Electronic Apparatus

VAN DER PLOEG, P. Industrial Electronics Apparatus. *Philips' Technical Library*, 1960, 97 pp.

621.392.1 Networks

THIRUP, G. Studies on Networks with Periodically Variable Elements. *Jul. Gjellerups Boghandel, Copenhagen*, 1959, 107 pp.

621.396 Radiocommunication

SCHLITT, H. Systemtheorie fur regellose Vorgange. *Springer-Verlag*, 1960, Berlin, 344 pp. (in German).

621.396.61 Transmitters

GRAY, Laurence, GRAHAM, Richard. Radio Transmitters. *McGraw Hill*, 1961, 462 pp.

621.396.62.004.2 Radio Servicing

BAILEY, A. R., BELL, E. C. Radio Servicing. Vol. 5, Specimen Answers to C. & G. Questions 1955/59, (Intermediate). *Norman Price*, 1960, 80 pp.

BAILEY, A. R., BELL, E. C. Radio Servicing. Vol. 6, Specimen Answers to C. & G. Questions 1955/59, (Final). *Norman Price*, 1960, 84 pp.

621.396.631 Calling Apparatus (radio)

A.R.R.L. The Mobile Manual for Radio Amateurs. *A.R.R.L.*, 2nd edn, 1960, 282 pp.

621.396.69 Printed Circuits

HORSEY, Eleanor F. (editor). Proceedings of the Symposium on Microminiaturization of Electronic Assemblies. *Hayden Book Co., Inc.*, 1960, 278 pp.

transistor work. Typical of design formulae is that for input resistance versus load resistance,

$$R_{\mathrm{IN}} = \frac{h_i - \Delta^h R_L}{1 + h_o R_L} \qquad \ldots\ldots(23)$$

A formula of this kind can be generalized as:

$$y = \frac{p + qx}{r + sx} \qquad \ldots\ldots(24)$$

When x is very small, $y = p/r$, and when very large, $y = q/s$. Plotted on log-log paper, the curve will have a step-like appearance (Fig. 3). The asymptote which can be used for the sloping portion of the curve is that which would occur with a zero coefficient for x in either numerator or denominator. If $q = 0$, then $y = p/(r + sx)$ and when x becomes very large, $y = p/sx$. This represents the inclined asymptote, and since logarithmic scales are used, the curve is that of log $y = (-1) \log x + \log p/s$, i.e. a straight line of slope -1. The inclined asymptote intersects each horizontal asymptote at a point which satisfies both equations, e.g. $y = p/r$ and $y = p/sx$, or $x = r/s$. The overall result for the general case is shown in Fig. 3.

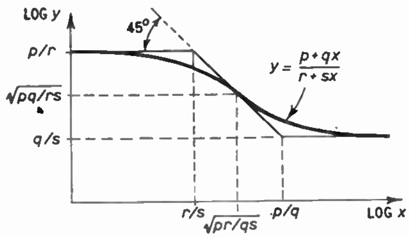


Fig. 3.

The ratios which appear in Fig. 3 resemble the deviations of Table 1, so that the points can very easily be relabelled with conventional parameters, as shown in Fig. 4.

5.2. *Transistor Parameters*

Figure 4 is typical of a common emitter circuit. Common base and common collector versions retain the same symbols, but the curve steps upward instead of downwards indicating, for example, that $z_{iB} > h_{iB}$.

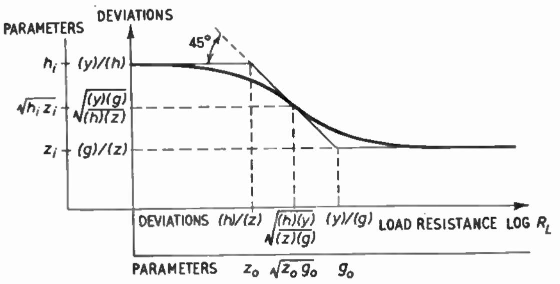
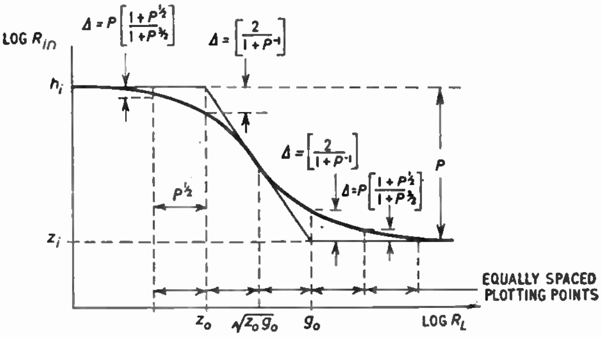


Fig. 4.

5.3. *Error in Asymptotes*

The numerical ratio of h_i to z_i (see Fig. 5) is referred to as P, the "projection ratio", and the error between curve and asymptote at the points z_o and g_o, is

$$\Delta = \frac{2}{1 - P^{-1}} \qquad \ldots\ldots(25$$



For nth plotting point from centre $\sqrt{z_o g_o}$, the error is

$$\varDelta = \left.\frac{1 + P^{\frac{n-1}{2}}}{1 + P^{-\left(\frac{n+1}{2}\right)}}\right]_{n \leqslant 1}$$

$$\text{or } \varDelta = P\left[\frac{1 + P^{\frac{n-1}{2}}}{1 + P^{\frac{n+1}{2}}}\right]_{n \geqslant 1}$$

Fig. 5. Error in asymptotic curves as a function of projection, *P*.

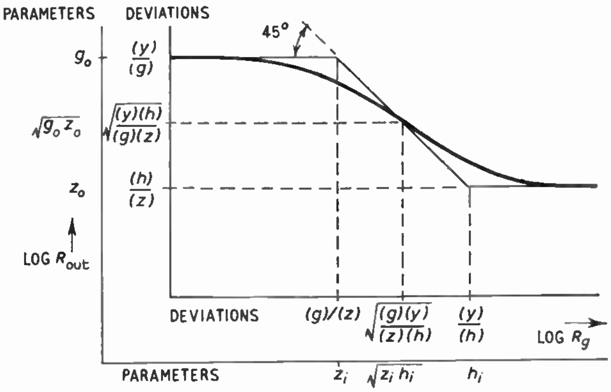


Fig. 6. Output impedance vs generator impedance (common emitter configuration).

To preserve the symmetry of the system, the error Δ is written as a ratio which is always greater than 1, and from the central cross-over point, several equally spaced plotting points are marked off. The error ratio for each plotting point then applies once on the left and once on the right of the centre of the system. Representative curves which are given in Figs. 5 to 9 are based on a detailed derivation.[7]

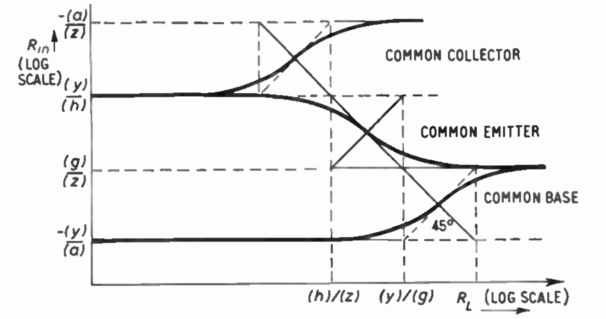


Fig. 7. Three-circuit configuration of input vs load impedance. All terms are common emitter deviations, with (*a*) assumed to be the predominant term in (*f*). A similar arrangement holds with output vs generator impedance.

5.4. *Active Two-port*

It can readily be shown that if a composite structure of R_{IN} vs R_L (or R_o vs R_g) is made for the three circuit configurations, CB, CE and CC, the central crossover points will fall in a straight line. This suggests the use of an approximate method for constructing design curves, as shown in Fig. 7, using only the common emitter *h*-parameters or the set of six common emitter deviations.[11]

5.5. *Gain Curves*

A typical formula for gain in a transistor is

$$A_i = i_o/i_i = h_f/(1+h_o R_L) \qquad \ldots\ldots(26)$$

This follows the pattern of the general case, eqn. (24), with $q=0$, and the asymptotes can be drawn, as in Fig. 8. The error ratio at the intersection of the asymptotes is 2.

To find the power gain, the current and voltage curves (which are on logarithmic scales) are added, and the result is as shown in Fig. 9. Note that the curve never attains the horizontal represented by the intersections of z_o and g_o with the inclined asymptotes. The maximum height of the curve is found by equating the first derivative to zero. Again, error ratios can be computed for equally spaced plotting points on either side of the centre, and the formulae are given in Fig. 9.

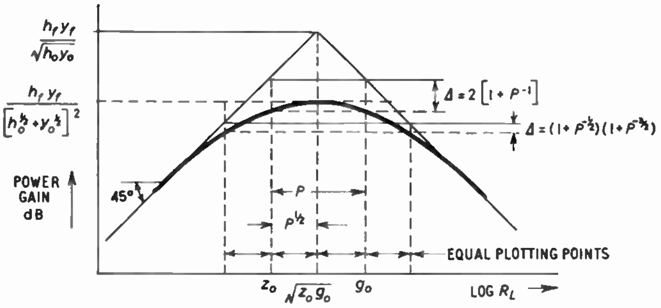


Fig. 9. Asymptotic plot of power gain vs load resistance, showing error at various plotting points and maximum power gain.

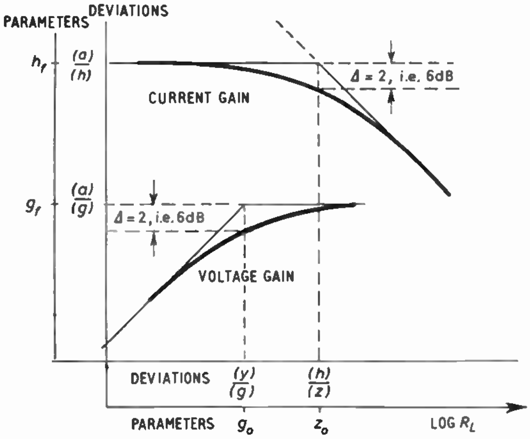


Fig. 8. Current and voltage gain vs load resistance.

6. Compound Networks

An essential in any discipline concerning network theory is the provision for coupling networks together. The combination may be simple or compound, and the problem is to express the deviations of the combined network in terms of the deviations of the component networks.

6.1. *Simple Combination Networks*

The addition of a one-port to a two-port is a straightforward manipulation, based on the rules given in Section 3. To find the deviation (*g*), for example, of the combined network, it is necessary only to write it as (v_i, i_o) and break up the term i_o into its component parts. Thus, from Fig. 10,

$$(g) = (v_i, i_o) = \left(v_i, i'_o + \frac{1}{z} \cdot v_o\right)$$

$$= (v'_i, i'_o) + \left(v'_i, \frac{1}{z} \cdot v'_o\right)$$

$$= (g') + \frac{1}{z}(y')$$

The effect of placing the impedance in different positions in the circuit is given in Table 3. *Z* may be resistive, reactive or complex, and the formulae are simplified somewhat if *Z* is described as a ratio, *m*/*n*.

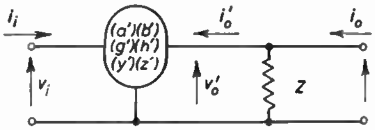
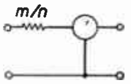
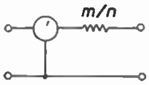
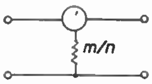
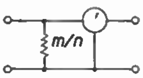
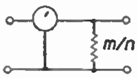
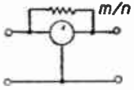


Fig. 10.

Table 3

Deviations of Simple Combinations of Networks

Deviations for composite network						
	In terms of deviations of component network and impedance m/n					
(a)	$n(a')$	$n(a')$	$n(a')+m(z')$	$m(a')$	$m(a')$	$m(a')+n(y')$
(b)	$n(b')$	$n(b')$	$n(b')+m(z')$	$m(b')$	$m(b')$	$m(b')+n(y')$
(g)	$n(g')+m(z')$	$n(g')$	$n(g')+m(z')$	$m(g')$	$m(g')+n(y')$	$m(g')+n(y')$
(h)	$n(h')$	$n(h')+m(z')$	$n(h')+m(z')$	$m(h')+n(y')$	$m(h')$	$m(h')+n(y')$
(y)	$n(y')+m(h')$	$n(y')+m(g')$	$n(y')-m(f')$	$m(y')$	$m(y')$	$m(y')$
(z)	$n(z')$	$n(z')$	$n(z')$	$m(z')+n(g')$	$m(z')+n(h')$	$m(z')-n(f')$

6.2. *Compound Two-Ports*

A more difficult situation—and in many respects a more remarkable one—is encountered when the deviations of a compound two-port are expressed in terms of the deviations of two component two-ports.

In any compound connection of two-ports, there are always two variables which can be expressed in terms of either network. These two variables are referred to as v_x and i_x, and, in conjunction with two desired variables, they form a set of four variables to which the uniqueness condition (Rules 3.2.1) applies.

An example will make the procedure clear. Figure 11 shows a compound network which consists of two cascaded two-ports, characterized by single and double primes, with unprimed symbols referring to the compound network. To evaluate the deviation (g) of the compound network, we first make a general statement of the uniqueness condition by expanding eqn. (12).

$$(p, q)(r, s)-(s, q)(r, p)+(s, p)(r, q) = 0 ......(27)$$

Requiring $(g) = (v_i, i_o)$, we apply eqn. (27) to the four variables v_i, i_o, v_x and i_x, giving

$$(v_i, i_o)(v_x, i_x)-(i_x, i_o)(v_x, v_i)+(i_x, v_i)(v_x, i_o) = 0 \quad ...(28)$$

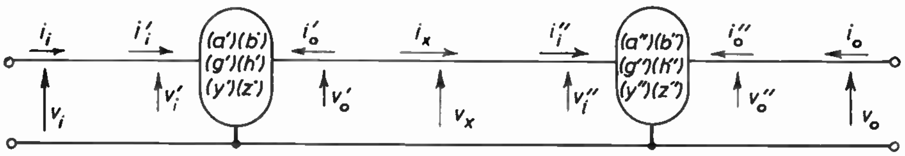
and by expressing v_x and i_x in terms of one or other of the component networks, this becomes

$$(g).[-(g')]-(z'')[-(y')]+(g')(g'') = 0 \quad ...(29)$$

It can be seen from eqn. (29) that the requirement has been met; the only terms present are the (known) deviations of the two component networks and the required deviation (g) of the compound network.

Table 4

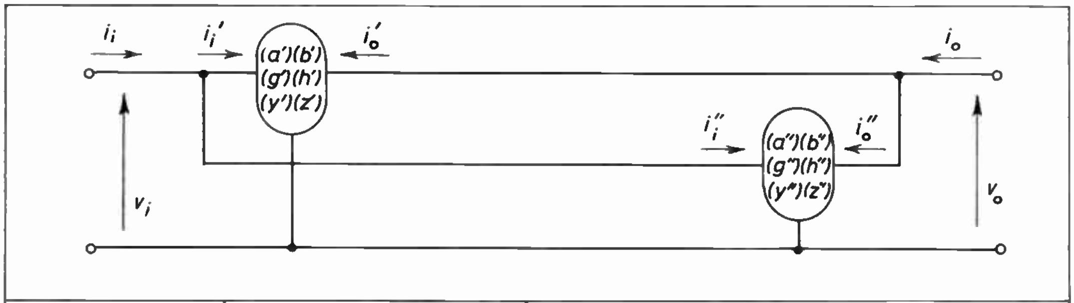
Deviations of Cascaded Two-Ports



Deviation to be evaluated	Selection of variables for substitution in eqn. (27)				Deviations of compound network in terms of deviations of individual networks
	p	q	r	s	
(a)	v_o	i_o	v_x	i_x	$(a) = (g'')(h'')-(y'')(z'')$
(b)	v_i	i_i	v_x	i_x	$(b) = (g')(h')-(y')(z')$
(g)	v_i	i_o	v_x	i_x	$(g) = (g')(g'')+(y')(z'')$
(h)	i_i	v_o	v_x	i_x	$(h) = (h')(h'')+(y'')(z')$
(y)	v_i	v_o	v_x	i_x	$(y) = (g')(y'')+(y')(h'')$
(z)	i_i	i_o	v_x	i_x	$(z) = (z')(g'')+(h')(z'')$
Starting condition: $(a') = (b'')$					

Table 5

Deviations of Parallel Two-Ports



Deviation to be evaluated	Description	Deviations of compound network in terms of deviations of individual networks
(a)	(v_o, i_o)	$(a) = [(a')+(a'')](y')$
(b)	(i_i, v_i)	$(b) = [(b')+(b'')](y')$
(g)	(v_i, i_o)	$(g) = [(g')+(g'')](y')$
(h)	(i_i, v_o)	$(h) = [(h')+(h'')](y')$
(y)	(v_i, v_o)	$(y) = (y')^2$ or $(y'y'')$
(z)	(i_i, i_o)	$(z) = [(z')+(z'')](y')$
Starting condition: $(y') = (y'')$		

The uniqueness condition can be imposed on any consistent set of four variables, and a different set can be chosen for the evaluation of each deviation of the compound network, as shown in Table 4.

A similar method may be employed for other combinations of two-ports, though a simple substitution is all that is necessary for the parallel case, Table 5. In each case, and in the case of series-to-parallel and parallel-to-series (which have been omitted here and are dealt with later), there is a "starting condition" which is implied by the fact that the deviations of the common variables (v_x, i_x) in Fig. 11, must be the same whether expressed in terms of the first or second network. Thus, in Fig. 11, the starting condition is $(a') = (b'')$, since

$$(v_x, i_x) = (v_o', -i_o') = -(a')$$

$$= (v_i', i_i'') = -(b'')$$

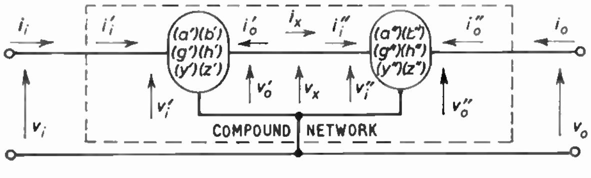


Fig. 11.

6.3. *Starting Condition*

The starting condition for the four cases is as follows:

Cascaded: $(a') = (b'')$
Parallel: $(y') = (y'')$
Series to parallel: $(h')-(a') = (h'')-(b'')$
Parallel to series: $(g')-(a') = (g'')-(b'')$

It is apparent that in each case, a simple combination of deviations is involved. If, for example, the deviations of the component networks in a cascaded two-port are known, but the starting condition has not been fulfilled, we can exercise our freedom to use an arbitrary multiplier on each set so that

$$C(a') = D(b'') \qquad \ldots\ldots(30)$$

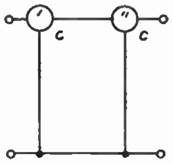
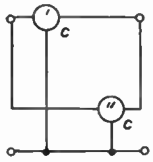
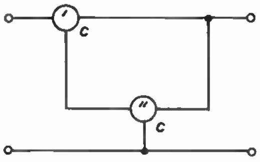
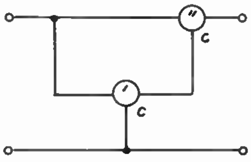
The formula for (z) in Table 4 then becomes

$$(z) = CD[(z')(g'')+(h')(z'')] \qquad \ldots\ldots(31)$$

If we write all the formulae so that the deviations of component networks occur only in pairs, one from each network, then each member of the resultant set of deviations has been multiplied by a constant, CD, as in eqn. (31). Such a multiplication does not affect the deviations (Rule 3.2.5) and accordingly the starting condition can be dispensed with provided that each formula is expressed in terms of pairs of deviations.

Table 6

Short Table of Deviations of Compound Two-Ports

	Cascaded	Parallel	Series-to-Parallel	Parallel-to-Series
				
	′ = first network	″ = second network	c = common terminal	
a	$a'a''$	$a'y''+y'a''$	$(a'-h')(a''-h'')-h'h''-z'y''$	$(a'-g')(a''-g'')-g'g''-y'z''$
b	$b'b''$	$b'y''+y'b''$	$(b'-h')(b''-h'')-h'h''-z'y''$	$(b'-g')(b''-g'')-g'g''-y'z''$
g	$g'g''+y'z''$	$g'y''+y'g''$	$(a'-h')(a''-h'')+(b'-h')(b''-h'')-$ $-f'f''-y'z''-h'h''-z'y''$	$-g'g''-y'z''$
h	$h'h''+z'y''$	$h'y''+y'h''$	$-h'h''-z'y''$	$(a'-g')(a''-g'')+(b'-g')(b''-g'')-$ $-g'g''-y'z''-f'f''-z'y''$
y	$g'y''+y'h''$	$y'y''$	$f'y''-y'h''$	$y'f''-g'y''$
z	$z'g''+h'z''$	$z'y''+y'z''$	$z'f''-h'z''$	$f'z''-z'g''$
	Note: $f = a+b-g-h$ All terms are deviations—brackets omitted			

Due to practical difficulties in the manipulations, it was found best to deal with the series-to-parallel and parallel-to-series cases in a sequence of steps, rotate-cascade-rotate, in order to obtain results with the deviations expressed in the desired form. The entire analysis of compound two-ports, with "built-in" starting condition, is presented in Table 6. It may be noted in passing that $(a)+(b)-(g)-(h)$ is a frequently recurring quantity in the various tables. It is useful to give it the symbol (f) and evaluate it along with the other deviations.

7. Miscellaneous Network Properties

7.1. *Active and Passive Networks*

For a passive network $(a) = (b)$. For an active network $(a) \neq (b)$. Thus all the work described in this paper related to active networks, with passive networks as a special case in which $(a) = (b)$.

7.2. *Symmetrical Networks*

In a symmetrical network (which can always be passive), $(g) = (h)$ and $(a) = (b)$.

7.3. *Iterative Matched Networks*

If the input and output impedances are identical, then $(g) = (h)$. The network may be active or passive.

7.4. *Interchange of Ports*

If a two-port is turned around so that the input port becomes the output and vice-versa, the deviations follow by inspection of eqn. (11) and obey this rule: Interchange of input and output ports corresponds to an interchange of (a) with (b) and (g) with (h), while (y), (z) and (f) remain unchanged.

Table 7

Stability Criteria for Two-Port Network

Port Considered positive value = stable negative value = unstable		Condition of Remaining Port
Input	Output	
$(g)(z)$	$(h)(z)$	OC
$(h)(y)$	$(g)(y)$	SC

7.5. *Stability Criteria*

Inspection of the values of a set of deviations shows whether any short- or open-circuit instability exists. The open-circuit input resistance of a frequency-independent network, for example is $(g)/(z)$. For this resistance to be negative, (g) and (z) must have opposite signs, yielding a stability criterion $(g)(z) < 0$. Similar reasoning for other conditions yields Table 7.

From this table, it is clear that a system is potentially unstable if the deviations (g),(h),(y) and (z) do not all have the same sign. This is another way of saying the h- and z-determinants must be positive, a well-known stability criterion.

A more general theorem has been stated[3] as follows: If a closed-loop system is opened at any point, its h-matrix may be written

$$\begin{bmatrix} v_1 \\ i_2 \end{bmatrix} = \begin{bmatrix} h_{11} & h_{12} \\ h_{21} & h_{22} \end{bmatrix} \begin{bmatrix} i_1 \\ v_2 \end{bmatrix} \qquad \text{......(32)}$$

and for oscillation on closing the loop, $v_1 = v_2$ and $i_1 = -i_2$, whence

$$h_{21} - h_{12} + \Delta^h = -1 \qquad \text{......(33)}$$

This is a general stability criterion, of which the Nyquist criterion is a special case where $i_1 = i_2 = 0$. Translated into terms of deviations, eqn. (33) becomes

$$(a)+(b) = (g)+(h) \qquad \text{......(34)}$$

or simply

$$(f) = 0 \qquad \text{......(35)}$$

For the frequency-independent case, connecting the output to the input terminal (or "closing the loop") will cause oscillation if

$$(g)+(h) > (a)+(b) \qquad \text{......(36)}$$

A more convenient way of writing eqn. (36) is that (f) must have the opposite sign from (g), (h), (y) and (z) for the system to be stable when the loop is closed. Reference to Table 2 will show that this corresponds to closing one of the ports of a rotated network.

If frequency-dependent networks are used, the deviations may be complex quantities. In this case, the system will oscillate at complex frequencies which are roots of

$$(g)+(h) = (a)+(b) \qquad \text{......(37)}$$

with positive real parts.

7.6. *Deviation Zero*

Certain network properties are implied when any of the deviations becomes zero. If $(a) = (v_o, i_o) = 0$, then one of three conditions must be present :

(1) $f(v_o, i_o) = 0$, that is to say, v_o and i_o are functionally dependent only upon each other.

(2) v_o is a constant

(3) i_o is a constant

That is to say, the output port may be open or short-circuited, or uninfluenced by conditions at the input.

7.7. *Verification*

The uniqueness condition eqn. (12) is a useful check since it applies to any consistent set of deviations. It is comforting to be able to verify one's calculations by ensuring that $(a)(b) - (g)(h) + (y)(z)$ is equal to zero.

8. Frequency-Selective Networks

All the methods described in this paper apply with simple or complex variables. This leads to some interesting methods of dealing with frequency-responsive curves.

8.1. *Frequency Response*

Consider, for example, the two-port network shown in Fig. 12. The z-parameters are easily written, from inspection, giving column (1) of Table 8. Referring to Table 1 with the deviation (z) set equal to 1 (Rule 3.2.5), we can write a set of deviations as in column (2) of Table 8. Any required transfer function can be written as a ratio of two deviations, e.g.

$$\left.\frac{\partial v_i}{\partial i_i}\right|_{i_o} = \frac{(v_i, i_o)}{(i_i, i_o)} = \frac{(g)}{(z)} = R + 1/sC$$

and its frequency-response is immediately evident. Frequency-dependent load, generator or intermediate impedances can be taken into account with the aid of Table 3.

In order to use only positive powers of s, it is convenient to multiply the set of deviations by sC, and to write τ for the time constant, RC, giving results as shown in column (3), Table 8.

Table 9

Calculation of Deviations of Fig. 13 Network

	First network deviations	First network rotated "*CE* in terms of *CB*"	Second network deviations	First and second networks cascaded
(a)	$s\tau$	$s\tau - 1 - s\tau = -1$	1	$s\tau$
(b)	$s\tau$	$s\tau - 1 - s\tau = -1$	1	$s\tau$
(g)	$1 + s\tau$	$-1 - s\tau$	$1 + s\tau$	$1 + 3s\tau + s^2\tau^2$
(h)	$s\tau$	$s\tau + s\tau - 1 - s\tau - s\tau = -1$	1	$2s\tau$
(y)	τ/C	$-\tau/C$	τ/C	$(2\tau/C) + (s\tau^2/C)$
(z)	$s\tau/R$	$-s\tau/R$	$s\tau/R$	$(s\tau/R) + (2s^2\tau^2/R)$

Table 8

Calculation of Deviations

(1) z-parameters	(2) Deviations with $(z) = 1$	(3) Deviations with $(z) = sC$ and $\tau = RC$
$z_i = (1/sC) + R$	$(a) = R$	$(a) = s\tau$
$z_r = R$	$(b) = R$	$(b) = s\tau$
$z_f = R$	$(g) = (1/sC) + R$	$(g) = 1 + s\tau$
	$(h) = R$	$(h) = s\tau$
$z_o = R$	$(y) = R/sC$	$(y) = \tau/C$
$\Delta^z = R/sC$	$(z) = 1$	$(z) = s\tau/R$

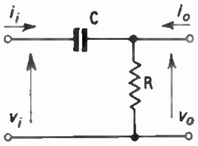


Fig. 12.

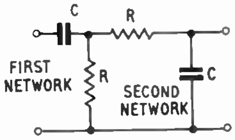


Fig. 13.

To illustrate a typical procedure, the steps are shown in Table 9 for the computation of the deviations of a compound network consisting of two Fig. 12 networks, rotated and cascaded, as in Fig. 13. Table 2, though designed for transistor rotations, is equally applicable to networks in general; the first network of Fig. 13 corresponds to *CB*, and the second to *CE*.

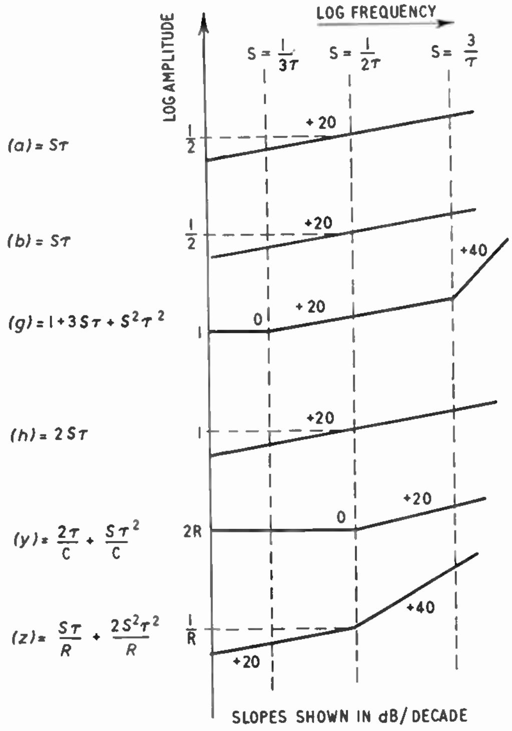


Fig. 14. Frequency response of Fig. 13 deviations.

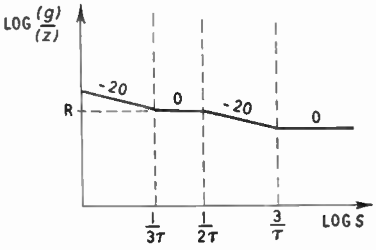
The last column of Table 9 gives the required deviations. If their frequency-response is plotted on log-log paper, subtraction of one curve from another is all that is required to give the frequency response of any required transfer function. It is useful, therefore, to plot the deviations asymptotically (Fig. 14).

A function such as $(g) = 1+3s\tau+s^2\tau^2$ has a value of $(g) = 1$ for low values of s; when the second term predominates, it acquires a slope of +20 dB/decade, since $(g) = 3s\tau$, and the intersection of these asymptotes occurs at $s = 1/3\tau$. Similar reasoning yields asymptotes, intersection points and slopes as indicated in Fig. 14.

To find, for example, the input impedance with open-circuited output, we can write this as

$$\left.\frac{\partial v_i}{\partial i_i}\right|_{i_o} = \frac{(v_i, i_o)}{(i_i, i_o)} = \frac{(g)}{(z)}$$

Subtracting the curve (z) from the curve (g) gives an asymptotic structure, as shown in Fig. 15, of the frequency-response of the input impedance. A simple operation such as this is all that is required for depicting the frequency response of any required transfer function.



Asymptotic plot of $\frac{(g)}{(z)} = \frac{1 + 3s\tau + s^2\tau^2}{\frac{s\tau}{R} + \frac{2s^2\tau^2}{R}}$ where $s = j\omega$

Fig. 15.

8.2. *Phase Response*

What has been done for frequency response can similarly be done for phase angle.[16] The "break points" in the frequency-response asymptotes correspond to those in the phase-response asymptotes. A slope of 20 dB/decade corresponds to a phase-shift

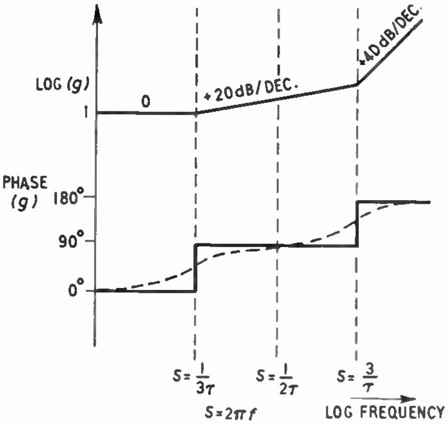


Fig. 16.

of 90 deg, 40 dB/decade to 180 deg, and so on. In Fig. 16, one of the deviations from Fig. 14 has been taken and the phase response indicated by the step-like "asymptotes", or break-points through which the actual curve will pass. The phase-responses are subtracted to give the phase-response of a ratio of two deviations, after the manner illustrated for the amplitude case in Fig. 15. The general subject of the correlation between phase and amplitude plots is dealt with very thoroughly by Bode.[17]

8.3. *Synthesis*

In synthesizing a network, it is helpful to be able to break up the frequency response into a numerator and denominator, backtracking through the procedure of Section 8.1. However, only the simplest examples have so far been tried out by the writers.

9. Distributed Parameter Networks

Recently, the development of micro-system solid-state elements has stirred activity in "distributed parameter networks"; a basic form of such networks consists of a resistive film capacitively coupled to an electrode as indicated by the circuit in the last column of Table 11. Analysis of such networks using matrix methods has been performed.[12, 13] Our purpose here is to show that the technique of using deviations applies *a forteriori* to the design problems associated with distributed parameter networks.

Table 10

Derivation of Distributed Parameter Deviations

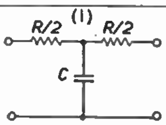
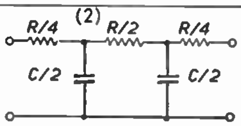
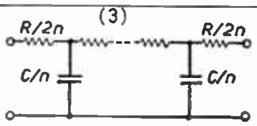
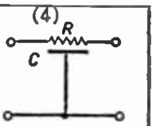
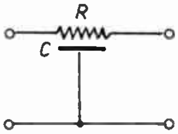
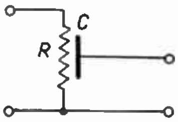
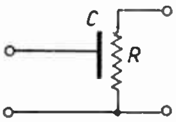
Deviations	(1) 	(2) 	(3) 	(4) 
(*a*) = (*b*) =	1	1	1	1
(*g*) = (*h*) =	$1+\theta^2/2$	$1+\theta^2/2+\theta^4/32$	$1+\theta^2/2+\theta^4/24+\ldots$	$\cosh\theta$
(*y*)/*R* =	$1+\theta^2/4$	$1+3\theta^2/16+\theta^4/128$	$1+\theta^2/6+\theta^4/120+\ldots$	$\theta^{-1}\sinh\theta$
(*z*)*R* =	θ^2	$\theta^2+\theta^4/8$	$\theta^2+\theta^4/6+\theta^6/120+\ldots$	$\theta\sinh\theta$
Note: $\theta = (j\omega RC)^{1/2}$				

Table 11

Deviations of R-C Distributed Parameter Two-Ports

			
(*a*) = (*b*) =	$\theta/\sinh\theta$	$\theta\tanh(\theta/2)$	$\theta\tanh(\theta/2)$
(*g*) =	$\theta\coth\theta$	$2\theta\tanh(\theta/2)$	$\theta\coth\theta$
(*h*) =	$\theta\coth\theta$	$\theta\coth\theta$	$2\theta\tanh(\theta/2)$
(*y*) =	*R*	*R*	*R*
(*z*) =	θ^2/R	θ^2/R	θ^2/R

9.1. *Basic Derivations*

Using the methods developed in previous sections, the deviations of a network such as that shown in column (1) of Table 10 are found. (Notice that for a passive, symmetrical network, $(a) = (b)$ and $(g) = (h)$). Two sections can be cascaded, as described in Section 6, and the result when the *total* resistance and capacitance are the same as in column (1) are given in column (2). When the operation is repeated for n sections, with n a large number, the deviations are as shown in column 3. For $n = \infty$, the well known hyperbolic functions result, as shown in column (4). The deviations of the basic distributed-parameter network follow identical rules to those for lumped-parameter circuits.

9.2. *Frequency Response*

In determining the frequency response, however, additional use is made of Rule 3.2.4. When the distributed-parameter network is manipulated, the resulting set of deviations may be multiplied by any convenient quantity, and in the selection of this quantity, the normalization imposed on the deviations is governed by the following rules, in order of importance:

(1) Avoid poles at finite frequencies.
(2) Avoid exponential poles.
(3) Minimize the number of exponential zeros, if possible.

In Table 11, a multiplier of $\theta/\sinh\theta$ has been used, and the distributed-parameter deviations are listed for various network configurations. The deviations are expressed in terms of $\theta = (j\omega RC)$ i.e. a normalized function of frequency. In establishing frequency-response curves, the various hyperbolic functions are plotted asymptotically on logarithmic scales as shown in Fig. 17. This table gives all the information necessary for computing the frequency response of any desired transfer function.

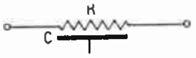


Fig. 18.
Distributed parameter one-port.

The procedures described in other sections of this paper enable any desired combination of networks to be evaluated. For example, the impedance of a one-port such as shown in Fig. 18 can be computed as the open-circuit input impedance of the circuit in the centre column of Table 11, or

$$\left.\frac{\partial v_i}{\partial i_i}\right|_{i_o} = \frac{(v_i, i_o)}{(i_i, i_o)} = \frac{(g)}{(z)} = \frac{2\theta \tanh(\theta/2)}{\theta^2/R}$$

$$= \frac{R}{(\theta/2)\coth(\theta/2)}$$

where $\theta = (j\omega RC)^{\frac{1}{2}}$

The frequency response of this function can be found by a simple manipulation of the curve for $\theta \coth \theta$ in Fig. 17.

The subject of distributed-parameter network analysis, using these methods, has been dealt with more fully elsewhere,[15] where a method of assessing

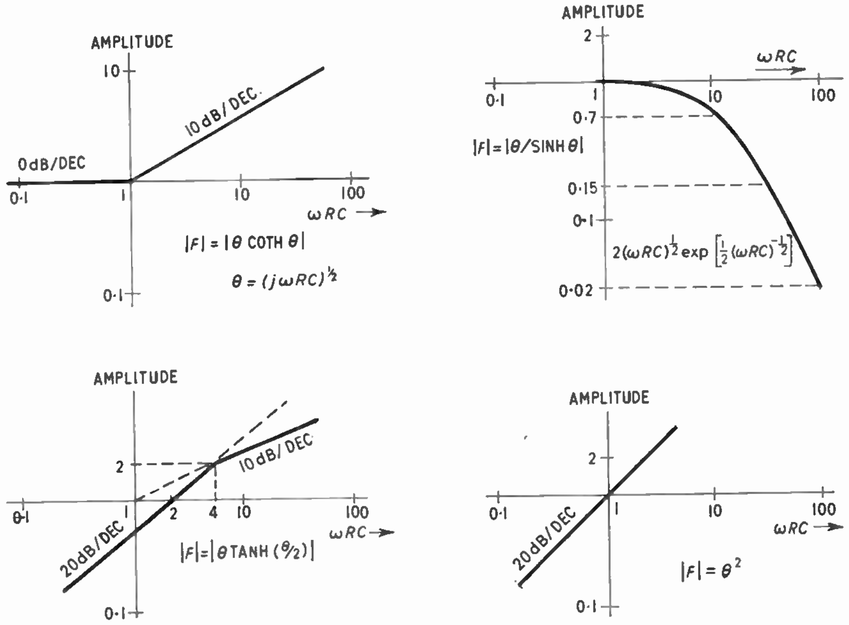


Fig. 17. Asymptotic frequency response curves for distributed parameter networks.

the error between asymptotic and true curves is also described.

9.3. *Phase Response*

What has been done for frequency response in the previous section can similarly be done for phase response.

10. The Transistor as a Distributed Parameter Network

The high-frequency response of transistors using a conventional equivalent circuit can be represented quite conveniently by the use of deviations.[9] However, a more interesting result is just around the corner. In many respects, the diffusion of carriers in a semiconductor corresponds to a distributed rather than a lumped parameter network, and some of the established methods of describing transistor frequency-response are based on lumped equivalent circuits. The benefits from using distributed equivalent circuits are within reach using the techniques discussed here.

11. Conclusions and Acknowledgments

At various times during the past few years,[4, 5, 6, 14,] the usefulness of applying methods paralleling those described here has been shown. Yet there has been little evidence of any general acceptance of them by the profession. In this paper, we have attempted to gather together some of the highlights of the subject, in the hope of stimulating the process of acceptance. It seems clear that in the calculus of deviations we have an extraordinarily powerful mathematical tool, worthy of much more attention than has hitherto been given it.

It must be acknowledged that we have merely picked up the mathematical methods of other branches of science and adapted them to our own purposes. Others have done the same—Matthaei, for example.[5] We have used our own terminology, for no better reasons than that we have grown accustomed to it, and that it has stood up to the test of four or five years of use in our own immediate circles.

Acknowledgment is made to Lockheed scientists, Dr. P. S. Castro, K. R. Morsette and F. A. Lindholm, who have worked on the problems of distributed-parameter networks, and to the Lockheed Missiles and Space Division for the support of this programme.

12. References

1. E. A. Guillemin, "Communication Networks", (John Wiley New York, 1935).
2. R. F. Shea, "Principles of Transistor Circuits", (John Wiley, New York and Chapman & Hall, London 1953).
3. N. A. Shaw, "Derivation of thermodynamical relations for a simple system", *Trans. Roy. Soc. (London)*, **A234**, 1935.
4. W. W. Happ, "Dynamic characteristics of four-terminal networks", *IRE Convention Record*, 1954.
5. G. L. Matthaei, "Some simplifications for analysis of linear circuits", *Trans. Inst. Radio Engrs. (Circuit Techniques)*, CT-4, September, 1957.
6. T. R. Nisbet and W. W. Happ, "Jacobians—a new computational tool for converting transistor parameters", *Electronic Industries*, November, 1958.
7. T. R. Nisbet and W. W. Happ, "For transistors, asymptotes solve design problems", *Electronic Industries*, August, 1959.
8. T. R. Nisbet and W. W. Happ, "Compound two-ports systematized", A.I.E.E. Pacific General Conference, 1960. (CP 60-1127).
9. T. R. Nisbet and W. W. Happ, "Jacobians of frequency-selective networks," A.I.E.E. Pacific General Conference, 1960. (CP 60-1128.)
10. P. S. Castro and W. W. Happ, "Distributed parameter circuits and microsystem electronics", *Proc. Natl. Electronics Conf.* 1960.
11. T. R. Nisbet, "Transistor input impedance", *Trans. Inst. Radio Engrs. (Circuit Techniques)*.
12. C. Holm, "Distributed-parameter networks for micro-miniaturization", *Electrical Manufacturing*, April, 1960.
13. P. S. Castro and W. D. Fuller, "Distributed-parameter networks for microminiaturization", A.I.E.E. Pacific General Conference, 1960. (CP 60-1097.)
14. L. Weinberg, "Network analysis," *Electrical Manufacturing*, January, 1960.
15. W. W. Happ and K. R. Morsette, "Analysis of the Frequency Response of R-C Distributed Parameter Networks by Means of the Calculus of Deviations", Tech. Memo, July 1960, Solid State Electronics Dept., Lockheed, Palo Alto, Calif.
16. K. R. Morsette and W. W. Happ, "Nyquist and Bode Plots of R-C-NR Components", Tech. Memo, Nov. 1960, Solid State Electronics Dept., Lockheed, Palo Alto, California.
17. H. W. Bode, "Network Analysis and Feedback Amplifier Design", §13 and 14. (Van Nostrand, New York, 1945.)

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A High-speed Graph Plotter

By

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(*Associate Member*)†

Presented at the South Western Section's Convention on "Aviation Electronics and its Industrial Applications", held in Bristol on 7th–8th October, 1960.

Summary: The paper describes a high-speed digital graph plotter capable of plotting up to 10 points per second on foolscap size Teledeltos paper with an accuracy of 0·2%. The plotter simultaneously inserts the graph ordinates at preselected intervals. It employs transistors throughout.

1. Introduction

The equipment to be described was developed by the author's company in conjunction with the Rocket Propulsion Establishment and forms part of an advanced recording and analysis system.[1] This system, designed by the Establishment, is concerned with data processing of test results associated with rocket motor development.[2] The complete system will be described briefly in order that some of the points demanded in the specification for the plotter may be better appreciated. Experience gained in constructing the prototype instrument has naturally indicated that several detailed modifications can either increase the reliability, or the operating convenience, or reduce the costs of production and a few such cases are mentioned later.

2. The R.P.E. Recording and Analysis System

The instrumentation for the static firing of solid propellent rocket motors at present used at the Rocket Propulsion Establishment is based mainly on the use of strain-gauge transducers, with 2 kc/s carrier amplifiers feeding 130 c/s reflecting galvanometers writing on 6-in. photographic paper.

In general it appears that accuracies in the order of 2% may be readily obtained with a comparatively small amount of effort in extracting the results from the photographic traces. Accuracies in the order of 0·5% are attainable with great care and an impracticable amount of effort devoted to computation of results. Semi-automatic trace readers are available which reduce this burden to some extent, but as far as is known no successful completely automatic trace-reader is available.

For accuracies better than 0·5% associated with the minimum of data-reduction effort, a completely new approach is necessary and this prompted the Rocket Propulsion Establishment to devise the digital recording and analysis system of which the plotter forms a part.

The system has been developed primarily for the instrumentation of the static firing of solid rocket motors but it is capable of being adapted to many other functions. Eight channels of analogue input of 20 mV amplitude are each digitized 100 times per second with a resolution of 1 part in 2000, and recorded on magnetic tape. Calibration and error-correcting information is also recorded, together with coding which enables a computer to deal with each item in the appropriate manner.

The special-purpose digital computer is programmed with a patch-board which has a capacity of 128 orders. This is more than sufficient to enable the computer to derive a linear term and a first order non-linear term from the calibrations, to apply zero drift and gain-drift correction, to compute the physical value of each point and cause it to be plotted on an automatic graph plotter, to compute time integrals of the recorded data, peak pressures and the times at which they occur, and ignition and burning times and to type these at the end of the analysis. The computer analyses one channel at a time, at a rate of 10 points per second.

The recording system comprises the transducers, multiplexer (i.e. commutating switch), and digitizer feeding into a high-speed magnetic tape recorder. The reel of recorded data is then passed to the analysis system for playback at reduced speed into the special purpose computer feeding the high-speed graph plotter.

3. Principle of Operation

The recording medium in the form of foolscap size Teledeltos recording paper is wrapped around and secured to a recording drum the length of which is equal to the height of the paper, i.e. 13 in. The diameter of the drum is sufficient to avoid overlap and is about 3 in. The recording paper has lines printed in one direction only, these being parallel to the longest dimension of the paper. Each tenth line is printed slightly thicker in order to facilitate subsequent reading and scaling of the curved curve, but this is not essential to the method of plotting.

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† Newman Industries Ltd., Yate, Bristol.

A recording head is mounted close to the drum which rotates during recording, the head being arranged to traverse the length of the drum mechanically. The carriage supporting the recording head is caused to engage with the running leadscrew when an electro-magnet is energized. This is associated with the start of plotting and causes the carriage to drop thereby engaging a half-nut on the carriage with the rotating leadscrew. The recording head is composed of two portions both in transverse alignment—the writing stylus and the line reading mechanism.

The former consists of a thin tungsten wire mounted on a cantilevered phosphor bronze strip. The method of recording on Teledeltos recording paper is as follows. The paper consists of a thin layer of kaolin deposited on a black conducting background paper and presents a grey/white surface. If a high voltage is applied between a stylus in light mechanical contact with this surface and the backing material, the resulting arc chips off a small patch of kaolin in the region of the stylus thereby revealing the black background in contrast. It will therefore be evident that a suitable pulse applied at the stylus whose duration is short relative to the transit time of the paper causes a point to be made, whilst an applied voltage of longer duration results in a line. The relative contrast of points and lines for a given writing speed is decided by the order of voltage, the higher the applied voltage the more dense the mark.

The line-reading mechanism consists of a miniature photocell and lamp focused in the plane of the paper via a cylindrical lens. The photocell detects the variation in reflected light occasioned by the passage of the printed lines on the recording paper. If 100 lines are printed on the recording paper, it follows that 100 pulses will be detected by the reading mechanism during each revolution of the recording drum. The reading portion of the recording head is mounted on the right-hand side of the writing stylus so that, during plotting, the head traverses from left to right and thus the line reading is not confused by the plotted curve. A magnetic transducer is also arranged to provide a pulse for each revolution of the drum and this is angularly positioned so that it occurs shortly before the first of the 100 lines passes the recording head, i.e. the abscissa line.

If it is required to plot a point, the magnitude of which is contained as a number in the computer store, the signal from the reading mechanism is arranged to subtract from the stored number. At the instant this number is counted down to zero the stylus is touching that portion of the recording paper appropriate to the magnitude of the original stored number. A pulse is then transmitted to the stylus and causes the point to be plotted. This procedure occurs on each revolution of the recording drum, the magnetic transducer being arranged to reset the computer store and permit reloading with the new number which is subsequently plotted. An important feature of this system is that close mechanical alignment and accurate printing of the paper is unnecessary. It will be appreciated that the speed of the recording drum must therefore be synchronized to the computer system and if 10 points per second are being presented a synchronous drum speed of 600 rev/min is necessary. In order to insert ordinate rulings at, say, each 50 plotted points, it is necessary to count 50 revolutions of the drum and then use this to gate a signal to the stylus for the duration of one revolution which commences with the reset from the magnetic transducer. The signal from the magnetic transducer is fed into counting circuits to count the revolutions of the drum. An alternative form of presentation is also provided whereby the ordinate ruling can be discontinued beyond the plotted point. This feature can be used to accentuate the form of the curve. The ordinate spacing is controlled by pre-selecting the counters and is conveniently arranged on a rotary switch.

The scaling of the curve to make best use of the paper is arranged in the ordinate sense by suitable programming of the computer, i.e. altering the magnitude of the stored count, and the abscissae scaling is dictated by the relative gearing between the drum and the recording head leadscrew.

The fact that the ordinate lines are portions of a helix is not normally noticeable except on the coarsest of traverse speeds when the ordinates subtend a small angle with the printed graticule. This does not introduce any errors into the interpretation of the graph since the points are observed relative to these ordinates. Again the low pitch of the helix results in the plotted points being close enough together to form an interpolated curve.

Briefly therefore from the operator's viewpoint the steps in plotting a graph are as follows: The operator having loaded the paper on to the drum and returned the recording head to the start position, switches on the motor which rotates the drum and leadscrew. The start signal from the computer lowers the recording head into position close to the paper and engages the leadscrew whereupon the plotter is synchronized to the system and the curve is plotted from the incoming information. The ordinates are inserted as preselected by the operator. When the recording head has fully traversed the paper and has come to rest at the right-hand side, the plotter is stopped by the operator, the paper removed, and the plotted curve is available for analysis without further processing.

4. Mechanical Features

The main structure is designed to afford sufficient rigidity for the drum and leadscrew driving gear whilst

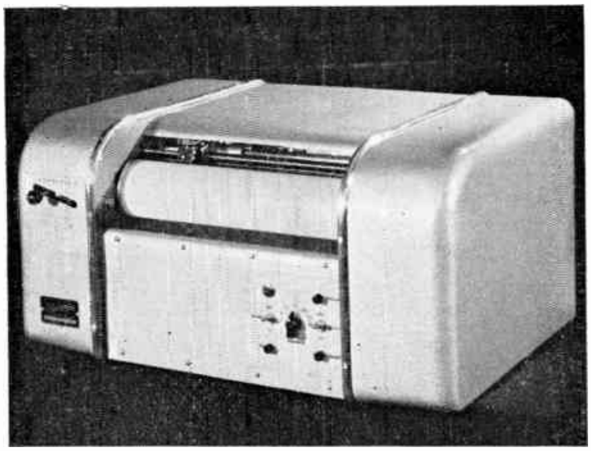

Fig. 1. General view of plotter.

ensuring a satisfactory environment for the plug-in transistor units (Figs. 1 and 2). The frame consists of two plates arranged in a vertical plane and maintained parallel to each other by means of suitably positioned bolsters. The recording drum is arranged at the front of the instrument to facilitate paper loading and unloading. A motor of the induction start, synchronous run, type provides power and the primary gear drive reduces the 3000 rev/min motor speed to 600 rev/min at the recording drum. The leadscrew drive is taken via a countershaft ten-ratio gear cluster, the idler gear being connected to the gear change lever which is accessible from the front of the instrument. The leadscrew has a blank portion without a thread at the right hand end to limit the travel of the recording head. The gear ratios are arranged in a geometric progression so that a scaling factor can be chosen which is sufficient to ensure a full-scale abscissae value of not less than two-thirds of the full-scale. The prototype instrument has a maximum leadscrew speed equal to the recording drum speed of 600 rev/min and the minimum speed is $\frac{1}{25}$th of this. In conjunction with a leadscrew pitch of 0·1 in. the full-scale range of plotting times ranges from 5 minutes with 3000 points to 12 seconds with 120 points.

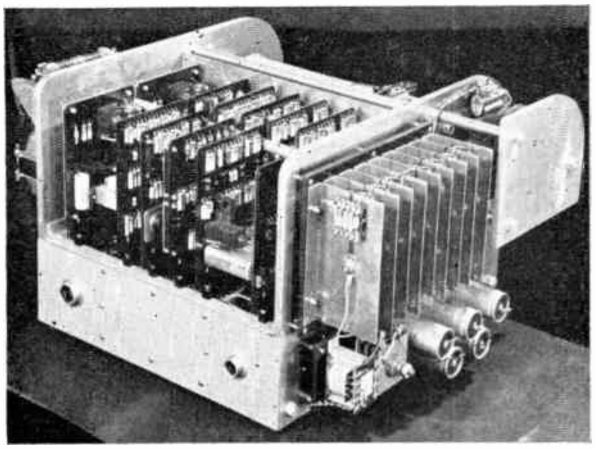

Fig. 2. View of the internal construction.

In view of the inherent speed of the system it was considered essential that time should not be wasted in the loading and unloading of the recording paper on to the drum and a simple but effective means was developed. The leading edge of the paper is inserted into a slot which prevents the stylus tearing the paper during rotation. As the paper is then wrapped around the drum, angular location is provided by spikes engaging the edge of the paper. The trailing edge is allowed to "flap" but a fixed bar mounted parallel to the drum face and in close proximity maintains the paper in contact with the drum in the essential region of the recording head so that no plotting errors are introduced. The recording head can be hinged up to allow ready replacement of the stylus.

The electronic parts of the system are positioned so that the effects of heat produced by the motor, transformers, power transistor cooling fins, etc., are kept separate from each other and from the plug-in units containing the transistorized logical parts of the instrument.

5. Electronic Features

Reference to the block diagram (Fig. 3) shows that the circuitry is divided into two distinct groups. Firstly there is the synchronous motor drive circuit whereby the motor speed is locked to a 100 c/s reference. In the Rocket Propulsion Establishment application this reference signal is contained on a synchronizing track on the magnetic tape. The 100 c/s signal after passing through a divide-by-two circuit is fed to the motor power amplifier. The motor is interlocked so that on switching on it starts via an auto transformer from the 50 c/s a.c. mains. This avoids rating the transistor amplifier for the motor starting current. On receiving the synchronizing signal the motor resynchronizes so that the plotter follows variations in the speed of the magnetic tape. A signal sensing and delay circuit detects when the synchronizing signal is present and the delay prevents the plotter attempting to follow the frequency run-up when the tape recorder starts.

As previously mentioned, the recording head consists of the writing stylus and the line reader. The stylus can be energized from either the line-writing amplifier which is basically a 2-kc/s oscillator and step-up transformer or the point-writing amplifier which is a mono-stable circuit providing a pulse width of approximately 200 μs at a peak voltage of the order of 1000 V. The line-writing amplifier is gated by the line control unit from the resetting pulse at

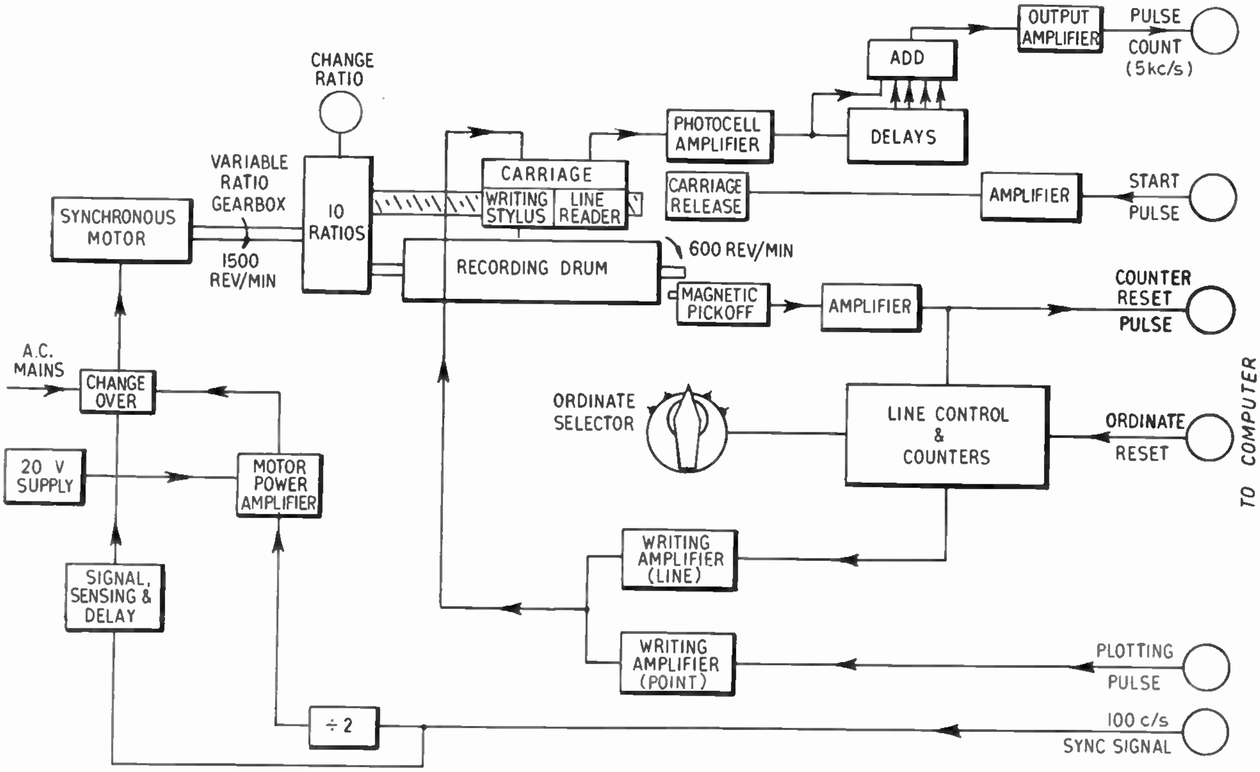

Fig. 3. Block diagram of the plotter.

each revolution of the drum. Two decade counters are associated with the ordinate selector and since each decade consists of a divide-by-five and a divide-by-two circuit the selector can be preset to plot an ordinate for each 1, 2, 5, 10, 20, 50 or 100 plotted points, i.e. revolutions.

The line-reading mechanism has its lamp supply arranged so that it is only illuminated when the drum is running and ready to receive information. This is purely in the interest of lamp life since it is necessarily very small and difficult to ventilate adequately. The pulses detected by the photocell with the passage of the printed lines on the recording paper are amplified and fed into a set of four delay units. The outputs from these delays are then summed with the original signal from the photocell amplifier. The delay times are set at approximately 200 μs so that with 100 lines being read each 0·1 second the nominal 1 kc/s signal is divided into fifths and summed to produce a nominal 5 kc/s signal. This 5-kc/s pulse count signal is directed into the computer store or counter thereby providing an interpolation to one-fifth of a line spacing or 0·2%.

The carriage release mechanism is actuated by an electromagnet energized by the start pulse through an amplifier.

The logic plug-in units together with the 8 V supplies are of the standard type, each board being approximately 4½ in. × 6½ in. with a 15-way gold-plated connector on the lower edge. These plug-in units are mounted in plate-rack style in the central section behind the recording drum and are accessible with the top cover removed.

6. Comparison with other Plotters

The advantages and disadvantages of this type of instrument are compared in Table 1 with those of the two most common alternatives.

The first is a parallel head digital plotter of the type utilizing a "comb" of styli spanning the paper, the position of the mark being controlled by exciting the appropriate stylus. If separate amplifiers are associated with each stylus, many points can be simultaneously plotted. However a more economic method is to use a binary decoder or relay "tree" which simplifies the system but reduces the speed of plotting and restricts the plotter to one curve at a time.

The two co-ordinate analogue plotter is rather better known, one form of which utilizes a gantry motored to the position as dictated by one parameter in the form of a d.c. voltage. The gantry carries a stylus holder driven along the gantry and at right angles to its direction of motion in accordance with the magnitude of the other parameter. The paper is normally laid flat in the horizontal plane. A slide wire or lineal potentiometer associated with each axis provides position feedback to the driving servo system.

Table 1

Features	High-speed Graph Plotter	Parallel-head Digital Plotter	Two Co-ordinate Analogue Plotter
Ordinate resolution.	0·2%	Dependent on styli spacing.	Dependent on resolution of position feedback transducer—0·5%
Speed of plotter.	10 points/second.	Can be faster.	Limited by mechanical inertia.
Simultaneous plotting of more than one curve.	Not possible.	Possible if separate input circuits per stylus; not if stylus switching "tree" used.	Not possible.
Form of recording medium.	Sheet.	Sheet or roll.	Sheet.
Paper alignment.	Not critical.	Critical.	Critical.
Accuracy in plotting step function.	100% following of input data.	100% following of input data.	Dependent on moment of inertia of head and positioning servo stiffness.
Abscissae resolution.	Dependent on pitch of helix.	Dependent on speed of recording medium.	Dependent on resolution of position feedback transducer—0·5%
Form of input.	Digital synchronous.	Digital non-synchronous.	Analogue non-synchronous.

7. Conclusions

In conclusion some suggested fields of application can be mentioned. Firstly, applications exist which are fundamentally similar to that inspiring the design, many of these being in the aircraft and allied industries for the rapid assessment of test results. Much of the value of, say, a test-bed run of an engine may well be lost due to the time taken to collate fully and analyse the multitude of readings before the next run. The costs of this assessment can also be reduced since a larger number of readings can be recorded per test run and their full value reaped.

Whereas the instrument was originally designed for use with a special-purpose computer this can equally well be a suitably programmed general-purpose computer. In this context it is probably desirable to provide an auxiliary counter-store with the plotter to avoid utilizing too much of the computer storage facilities. Fundamentally this is a binary counter which can be supplied with synchronized data either in serial or parallel form. The value of the count is reduced to zero by means of the graph plotter line counting mechanism whereupon a signal is produced to energize the point writing circuit. The counter is then reset to the next value to be plotted.

Another suggested application for the plotter in the commercial field when used in conjunction with a general purpose computer is as an alternative output to the typewriter, high-speed printer, etc. Some commercial information is best displayed in graphical form, an example being certain types of statistical information.

As already stated, development experience has shown details where improvements are to be made. One example of this is the synchronous drive system which despite its apparent simplicity for the application has introduced other difficulties. In order to reduce the wattage dissipation and size of the transistor drive amplifier, this must be used as an on-off device thereby passing square waves into the motor via its matching transformer. The motor and the transformer reflect high inverse voltages into these transistors which then require series protection, resulting in a significant lowering of efficiency. The photocell amplifier and associated circuitry had also to be very carefully screened from this radiated square wave. A digital servo drive is to replace this system. Another refinement which is to be fitted to production models is a stroboscope working off the point-writing amplifier so that the curve can be observed during the plotting process. Other small refinements are also to be incorporated which it is hoped will make the high-speed graph plotter a useful contribution to general data processing equipment.

8. Acknowledgments

The author wishes to record his thanks to the Directors of Newman Industries Limited for permission to present this paper. The author would also thank Dr. W. R. Beakley of the Rocket Propulsion Establishment, Westcott, who was responsible for the overall instrumentation project, for his valued assistance during the course of the project.

9. References

1. British Patent Application No. 2231/60.
2. W. R. Beakley and D. S. Dean, "A digital recording and analysis system for static testing of rockets". Internal memorandum, Ministry of Aviation, Rocket Propulsion Establishment.

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APPLICANTS FOR ELECTION AND TRANSFER

As a result of its meeting on 27th April the Membership Committee recommended to the Council the following elections and transfers.

In accordance with a resolution of Council, and in the absence of any objections, the election and transfer of the candidates to the class indicated will be confirmed fourteen days after the date of circulation of this list. Any objections or communications concerning these elections should be addressed to the General Secretary for submission to the Council.

Direct Election to Member

BRIGHTMORE, Air-Commodore Anthony George, R.A.F. *Weybridge, Surrey.*

Transfer from Associate Member to Member

SINGH, Kirpal. *Kuala Lumpur, Malaya.*

Direct Election to Associate Member

HOWSON, David Philip, B.Sc., M.Sc. *Solihull, Warwicks.*
JAMES, Lt. Allan, M.A.(Cantab.). *Garston, Herts.*
JOHNSTON, James Stewart, B.Sc. *New Malden, Surrey.*
LAI, David Soson. *London, N.3.*
LISTER, Lt.-Col. Anthony Wynter, R.A. *Salisbury, Wilts.*
LOWE, Ronald, B.Sc. *Reading, Berks.*
McKEE, Kenneth MacIntosh, B.Sc. *Enfield, Middlesex.*
MEACOCK, Eric Anthony, B.Sc. *Norwich.*
PARSONS, David Ernest, B.Sc.(Eng.). *Egham, Surrey.*
RICKARDS, Arthur John. *Greenford, Middlesex.*

Transfer from Associate to Associate Member

PERKS, Flt.-Lt. Roger Ian, R.A.F. *Watford, Herts.*
YOUNG, Alexander Macnab. *Newcastle-upon-Tyne.*

Transfer from Graduate to Associate Member

BRADFIELD, Derek William, B.Sc. *Ilford, Essex.*
FRAZER, Charles Henry. *Whiteley Bay, Northumberland.*
HALL, Major Wilfred Francis. *Crowthorne, Berks.*
McARTHUR, James. *Retford, Notts.*
MILES, Ronald Boyce. *Welwyn, Herts.*
NARENDRANATHAN, Ponnudurai, B.Sc. *Wellawatte, Ceylon.*
SAUNDERS, Patrick William. *Southend-on-Sea.*
WHEELER, Eric Gilbert. *Cheltenham, Glos.*

Transfer from Student to Associate Member

FERNANDES, Antonio Taurino, B.Sc. *Chesham, Bucks.*
MACNAMARA, Patrick, Colman. *Dublin.*
WINTERBOTTOM, Keith. *Wallasey, Cheshire.*

Direct Election to Associate

BONNER, Trevor Frank Kirkpatrick. *Bushey, Herts.*
CLEGG, Ronald Wright. *Barnard Castle, County Durham.*
O'BRIEN, Daniel. *Dublin.*
RUTTER, Jack Arthur William. *Sawston, Cambs.*

Direct Election to Graduate

ALLCOCK, Kenneth Malcolm Alan. *Barking, Essex.*
BEARD, Douglas Albert. *Bracknell, Berks.*
BLANCHARD Colin Reginald. *Chelmsford, Essex.*
BRITTAIN, Austin Furey, B.Sc.(Eng.). *Teddington, Middlesex.*
DOWNES, Flt.-Lt. John Philip, R.A.F. *Woodhall Spa, Lincolnshire.*
EN, James Lee Khiu, B.Sc.(Hons.). *Sarawak, Borneo.*
NEEDHAM, Laurie Faragher. *Towcester, Northamptonshire.*
SOUCH, Geoffrey. *Cheadle, Cheshire.*
*THOMAS, Edison Symonds, M.Sc., B.Sc. *Bangalore.*
WAN, Seng Kong, Dip.El. *Brighton, Sussex.*

Transfer from Student to Graduate

BUZZELLI, Giovanni Emilio. *Turin, Italy.*
CURLEY, Michael Joseph Colum. *Dublin, Eire.*
GOVINDAKRISHNAN, Tirunilayi Krishnan, B.Sc. *Madras.*
JAGJIT RAI. *Bombay, India.*
MacCALLUM, William Anthony. *Haddington, East Lothian.*
SENG KIM GOH. *London, N.W.6.*
UPENDRA, Dattatreya Bhardvaj. *Bombay.*
VENKATESWARAN, K. *Mallasamudram, India.*
WEST, Noland. *Poole, Dorset.*

STUDENTSHIP REGISTRATIONS

The following students were registered at the 28th March and 27th April meetings of the Committee.

LOWRY, John S. *Stevenage, Herts.*
LUKE, Ebenezer George. *London, N.19.*

MACLEAN, Alexander M. *Isle of Lewis.*
MARSDEN, Norman Wain. *Sheffield.*

NASMYTH, David, B.A. *Wells, Somerset.*
NOBLE, David. *Middlesbrough, Yorkshire.*
NWADUKWE, Dennis Enweluzo. *Lagos.*

PANG, Peter Guan. *Singapore.*
PASS, Brian James. *Fairview, Canada.*
PENROSE, Ronald J. *Camborne, Cornwall.*
PETERS, Martin Bryant. *Berkhamsted, Herts.*
PHILLIPS, Edmund. *Cambridge.*
PREWER, Brian E. *Wells, Somerset.*

RAO, Yellanki Syama Sundara, B.Sc. *Visakhupatnam, India.*
RICHARDS, Frederick J. *Oldham, Lancs.*

SARKAR, Dwijendra Nath. *Parganas, India.*
SCOTT, Henry George. *London, W.10.*
SHOTTER, Raymond J. R. *Paisley, Renfrewshire.*
SOWUNMI, Isola Babalola. *London, N.16.*
SRIDHAR RAM, N. S., B.Sc. *Bangalore.*
STEDMON, Roy Frank, B.Sc. *Harrow.*

TARR, Anthony F. *Glastonbury, Somerset.*
TOWNSEND, Leslie John. *East Croydon.*

UNNITHAN, K. Padmanabhan. *Trivandrum, South India.*

WARRINGTON, Richard Arthur L. *Newmarket.*
*WILLIAMS, Gordon Wilfred. *Port Talbot.*
WILLIAMS, Lascelles A. *London, N.20.*
WILLIMOTT, Ernest John Conway, B.Sc. *Glastonbury, Somerset.*
WOODS, John F. *Wembley, Middlesex.*
WROE, Frank. *Banbury, Oxfordshire.*

ABADALE, Kolapo Adegare. *Ibadan, Nigeria.*
ALUPAKUSADI-OBO, J. M. *Kampala, Uganda.*
ANAND, Satish Chander, M.Sc. *Jammu Tawi, Kashmir State.*

BATTEN, Derek John. *Abergavenny, Mon.*
BHATIA, Sunder Sagnomal, B.Sc. *Bombay.*
BRUMFITT, James R. *Edinburgh.*
BURDEN, Peter. *Liverpool.*

CARR, Philip, B.Sc. *Broxbourne, Herts.*
CHOO, Mun Ling. *Singapore.*
COHEN, Semah, B.Sc. *Haifa, Israel.*
COX. Edward. *Sydney, Australia.*
CROMARTY, David. *Bracknell, Berks.*
CROMWELL, James. *London, W.14.*
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Recent Developments in Coaxial Cables for Television Distribution

By

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Presented at the Symposium on New Components held in London on 26th–27th October 1960.

Summary: The paper outlines the requirements for coaxial cables for television distribution purposes in Bands I and III. The various factors affecting the performance of such cables are mentioned, and recent improvements in construction of the outer conductor and dielectric are discussed. A description is given of the test methods used for determining the uniformity of impedance of cable lengths and the relative screening efficiency of the various constructions. Details are given of the characteristics of the various newly developed cables and these are compared with the more standard braided constructions. The cables are divided into two classes: (*a*) those suitable for trunk routes, and (*b*) smaller cables for distribution purposes.

1. Introduction

For some time, fairly standard types of radio frequency cable have been employed in community antennae systems, such as exist in blocks of flats for distribution of television signals. In general, these single-braided types have been satisfactory, while as an alternative with improved screening, lead and aluminium sheathed versions have been used. However, recently more widespread systems are coming into being using television broadcasting Band I and Band III frequencies and in these cases cables are required which have simultaneously good screening and low attenuation.

Much thought has been put into this problem by cablemakers and this paper is concerned with various new cables which have been developed to meet this particular requirement which is of growing importance in the field of television distribution by wire.

2. Statement of the Problem

The main requirements for the cables are:

(*a*) low attenuation

(*b*) good screening

(*c*) good flexibility.

These requirements are very often conflicting and in addition must be obtained usually within a predetermined overall diameter. The latter consideration is necessary for aesthetic and economic as well as technical reasons.

To make a start on a problem of this nature involving a number of interdependent variables, it is necessary to fix certain of the parameters. Opinions of course vary on the optimum figures to be adopted, as system operators have to make a careful balance of costs between cable and terminal equipment. The attenuation and overall size of cable are usually the two parameters to be fixed as targets. Depending on the top operating frequency of the particular systems, this will determine the degree of difficulty in meeting the specified attenuation within a given overall diameter of cable.

Typical target figures for trunk cables set for attenuation and overall diameters are of the order of 1·0 to 1·5 dB/100 ft and 0·5 to 0·6 in. respectively. The top frequency at present under consideration is around 200 Mc/s and designs of cables with the above characteristics at this frequency have been considered. Smaller cables with attenuations in the range 2·5 to 3·0 dB/100 ft have been considered for distribution purposes.

3. Factors Affecting Cable Performance

The factors affecting the attenuation of a coaxial cable can be most readily appreciated by reference to the following relationship which is valid above frequencies of 1 Mc/s:[1]

$$\alpha = \tfrac{1}{2}R/Z_0 + \tfrac{1}{2}GZ_0 \qquad \ldots\ldots(1)$$

where α = attenuation

R = total a.c. resistance of inner and outer conductors

G = conductance of dielectric

$= \omega C \tan \delta \qquad \ldots\ldots(2)$

C = capacitance of cable

δ = dielectric loss of insulation

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† British Insulated Callender's Cables Ltd., Helsby, Cheshire.

$\omega = 2\pi \times$ frequency

Z_0 = characteristic impedance

$$= \sqrt{\frac{L}{C}} = \frac{138}{\sqrt{\varepsilon_r}} \log_{10} D/d \text{ (ohms)} \quad \ldots\ldots(3)$$

L = inductance

ε_r = relative permittivity

D = diameter over dielectric

d = diameter of inner conductor

The first term in eqn. (1) usually predominates, and thus it will be seen that the attenuation primarily depends on the resistance of the inner and outer conductors and the characteristic impedance. The latter is usually fixed at 75 ohms for reasons of standardization, although it is sufficiently near the optimum for minimum attenuation. Equation (3) shows that if the dielectric constant can be made as low as possible, this will result in a high characteristic impedance and thus a low attenuation. The impedance can however, be retained at 75 ohms by increasing the inner conductor diameter, and the reduction in attenuation is then achieved by the lower resistance of the inner conductor. Any further improvement in resistive losses can only be achieved by a change of construction. There is little that can be gained economically from a constructional change for the inner conductor, but attention must be concentrated on the form of the outer conductor, which also determines the screening performance of the cable. In this respect, it is well known that the conventional braids are, relatively speaking, electrically inefficient at radio frequencies, having a high resistance and poor screening performance for the amount of conducting material employed. Double braids, both with and without intersheath, have been tried and whilst these give some improvement in screening, there is no improvement in attenuation. The other factor mentioned earlier of flexibility is of course met satisfactorily with braided types and here lies the basis of the problem, i.e. to obtain a cable with a pre-determined overall diameter, having the lowest possible attenuation, highest screening performance and best possible flexibility.

4. Cable Constructions

A construction of outer conductor which undoubtedly has one of the best performances for attenuation and screening is the solid-drawn aluminium tube, but this is very stiff by comparison with a braided cable. The best alternative which has been developed to date is a longitudinally-applied copper tape having transverse corrugations in order to impart some degree of flexibility. Cables of this type have already been manufactured with the longitudinal tape having an overlap. There is, however, some field leakage at the overlap and from the point of view of screening a seam welded tube gives an improved performance. Again, however, the welding of the seam causes stiffening of the cable. The solution to these conflicting requirements depends largely on the input signal level, which is usually determined by an economic balance of cable and equipment costs. If the signal is relatively low, the more flexible type without seam welding is almost certainly satisfactory. If, however, on certain parts of trunk routes a higher voltage is required, then it may be necessary to have a better screened cable (either seam welded copper or drawn aluminium). For the smaller distribution cable, the accent is usually on flexibility and with lower signal levels the version with an overlapped seam should be adequate for this purpose.

Consideration must now be given to the second term of eqn. (1). It has already been shown that a dielectric with a low permittivity is desirable. However, for mechanical reasons there is a limit, and a cellular polythene dielectric having sufficient mechanical strength has a permittivity of the order of 1·45. A semi-airspaced construction has a somewhat lower permittivity of the order of 1·35 but most constructions of this type allow ingress of moisture from the ends unless special precautions are taken to seal them. The dielectric loss of the construction is also of importance, particularly at the top frequencies in Band III, since power losses in the dielectric are directly proportional to frequency (see equations (1) and (3)). At the present time, the dielectric loss angle of tough versions of cellular polythene is of the order of 0·0008. This compares with a figure of 0·0002 for solid polythene and is due to the polar nature of the blowing agent residue in the cellular version. Similar values are also being obtained at present with cellular polypropylene which is a possible alternative. The semi-airspaced construction is usually rather less than 0·0002. Thus, here again, present art offers two possible alternatives, namely, a cellular plastic with discrete bubbles which do not allow a direct moisture path through the cable, or semi-airspaced construction with lower dielectric loss but allowing moisture penetration. Opinions of most systems operators seem to favour the cellular version since special precautions at cable ends are not required.

5. Measurement of Cable Characteristics

Before giving details of the characteristics of the various constructions discussed above, it is of interest to give some details of the methods of measurement.

The mean impedance and attenuation are determined by conventional bridge and/or slotted line techniques[2] and will not be discussed further here.

5.1. *The Uniformity of Impedance*

It will be appreciated that there are variations in diameter and permittivity along the length of a cable and these cause local changes in characteristic impedance. These in turn cause reflections of the signal which are characterized by variations of the input impedance vs frequency performance of the cable. This may be measured by a sweep technique whereby the output from a sweep generator covering the required frequency band is fed to the cable via an attenuator. The signal variations at the input of the cable are detected, and amplified and then displayed on a cathode-ray oscilloscope (see Fig. 1). The change in signal level in decibels and the resulting change in impedance which this represents may then be calculated. A typical commercial instrument covers the frequency band 0·5 to 400 Mc/s in a number of steps and any "resonances" can clearly be seen. Resonances can occur in the impedance vs frequency curve due to periodic variations in the construction of the cable, which may only be very small but cause reflections which are in phase if the periodicity is a multiple of the wavelength. This summation of reflections can cause the input impedance of the cable to be several times the rated impedance at the "resonant frequency" and clearly can lead to undesirable effects in a system.

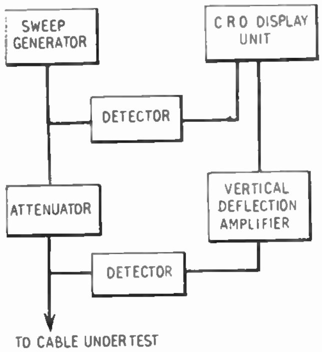

Fig. 1. Layout of impedance/frequency sweep test equipment.

Considerable experience has been obtained on braided constructions with 0·064 in. diam. conductor. Initially the impedance sweep test showed a sharp resonance at frequencies of the order of 140 Mc/s, the impedance at this frequency being of the order of 100 ohms and in some cases as high as 125 ohms. The general level at other frequencies on such cables was ± 10 ohms from the nominal of 75 ohms. After much experimental work had been carried out, this variation was reduced generally to ± 8 ohms over the whole frequency range of 40–220 Mc/s and with further precautions in manufacture a variation of ± 5 ohms can be obtained. This is as good as the length-to-length tolerance on impedance. With an aluminium sheath, which is drawn with great precision, the internal diameter of the sheath is very uniform and tends to annul any variation in diameter of insulated core. Hence, very consistent results can be obtained with this construction. Initial experience with cables having longitudinal copper tape outer conductors shows that the impedance uniformity approaches that of the aluminium sheathed cables, and a tolerance of ± 5 ohms over the whole frequency band can be maintained.

5.2. *Screening Efficiency*

The screening efficiency of a cable is usually expressed in terms of the surface transfer impedance[3] or coupling impedance. This is defined as the longitudinal voltage which appears at the outer surface of the screen when unit current flows down the inner surface and back along the inner conductor of a coaxial cable (Fig. 2).

The surface transfer impedance is usually measured by enclosing the cable coaxially in a metal tube thus forming the so-called "triple coaxial system". The transfer impedance of the cable screen is then measured by energizing the circuit formed by the outer screen and cable screen and measuring the voltage between the cable screen and the inner conductor. The apparatus in a comparatively simple form may be made to operate satisfactorily provided the cable sample under test is short compared with the wavelength. In order to avoid end effects this means an upper limit for the frequency of measurement of 30 Mc/s. Clearly, it is desirable to go higher in frequency for the present-day problems and apparatus has now been constructed for the higher frequencies where care must be taken to match all the coaxial circuits in order to avoid standing waves.

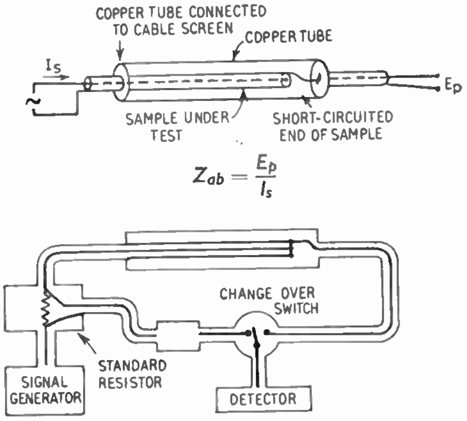

Fig. 2. Coupling impedance measurement.

The surface transfer impedance is equal to the d.c. resistance of the outer conductor at low frequencies

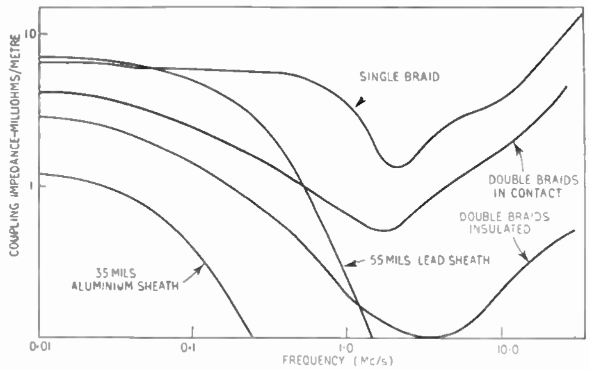

Fig. 3. Comparison of coupling impedance of braided and solid sheathed coaxial cables (0·285 in. over dielectric).

(i.e. usually less than 100 kc/s) and hence the larger the cable, the lower will be its surface transfer impedance and the better its screening efficiency. Also, the higher the conductivity of the screen, the better its screening performance. Thus, a homogeneous tube of a metal with high conductivity will give the best performance. As the frequency is increased, the surface transfer impedance of a homogeneous tube will decrease due to the familiar "skin effect" phenomenon. The rate of decrease is determined by the thickness of the screen in relation to the frequency. A braided or spiral lapped tape construction, on the other hand, has a turning point in its surface transfer impedance. This occurs where the inductance of the braid or spiral lapped tape begins to take effect. Thereafter, the transfer impedance increases linearly with frequency at least up to 1000 Mc/s (Fig. 3), the screen becoming less and less efficient. It is possible to design braids to have an optimum screening factor and this occurs where the inductance is kept to a minimum. Also double braids with an intersheath give a further improvement due to the reflection losses between the two screens.

6. Cable Dimensions and Characteristics

These can be conveniently divided into trunk and distribution cables. Trunk cables are larger, have lower losses and usually better screening performance.

6.1. *Trunk Cables*

Table 1 shows the results for trunk cables.

Firstly a word of explanation is required about the screening results. These are expressed in decibels relative to the single full cover braided cable. It should be realized that this is a parameter by which the *relative* screening efficiency of various constructions may be judged and does not give a ready answer to the amount of radiation to be expected from a particular cable in a particular location. With regard to the latter, the environment can affect the result considerably. On considering the results given in the Table, it will be seen that the attenuation figures at 100 and 200 Mc/s are comparable for the double-braided cable and the longitudinal copper-taped version. In this instance, the improved conductivity of the latter balances the improvement in dielectric properties of the fin and tube construction. Screening figures are in general, for this size of copper taped cable, comparable with those for the close-weave double-braided type.

Table 1

Characteristics of Trunk Cables

Inner Conductor diam. (inches)	Overall diam. (inches)	Type of dielectric	Type of screen	Maximum attenuation dB/100 ft		Minimum setting radius (inches)	Screening efficiency (dB)	
				100 Mc/s	200 Mc/s		30 Mc/s	100 Mc/s (estimated)
0·104	0·636	Cellular polythene fin and tube	Double braid	1·05	1·55	1½	20	30
0·104	0·590	Cellular polythene	Longitudinal corrugated copper tape	1·05	1·6	1½	30	30
0·104	0·600	Cellular polythene	Aluminium sheath	1·15	1·75	3	>140	>140
0·064	0·485	Cellular polythene	Double braid	1·75	2·6	½	10	10
0·064	0·384	Cellular polythene	Single braid	1·75	2·6	½	—	—

As predicted, the aluminium-sheathed version is far superior on screening performance but is at a slight disadvantage on attenuation. The minimum setting radius figures are in the expected order according to cable construction. The smaller double-braided cables with cellular polythene have also been included in this Table as showing attenuation figures at 100 Mc/s comparable with the 200 Mc/s figures for the other cables. These types may be used on narrow band systems having top frequencies of the order of 100 Mc/s.

6.2. *Distribution Cables*

Table 2 shows the characteristics for the smaller distribution cables with a top frequency attenuation in the range 2·5 to 3·4 dB/100 ft. Again, the lowest results in this size range are shown by the longitudinal copper tape version. The double-braided versions are the next lowest but it should be noted that the fin and tube type is smaller in overall diameter. Again a smaller size of double-braided cable is included for 100 Mc/s systems. The screening and setting radius figures show a similar pattern to the corresponding trunk types. Concerning the aluminium-sheathed types, it is of interest to note that the thinner wall version is still excellent on screening performance and, for a given overall diameter, allows a decrease in attenuation.

6.3. *Downlead Cables*

The system operators state that a well-screened downlead is likely to be important. This is required in order to prevent either direct radiation from the receiving set or pick-up in the set with the downlead acting as an aerial.

It is very necessary that this construction should be flexible and a suitable alternative to a double braid with intersheath would appear to be the best answer. Constructions are under consideration at the present, but no definite pattern is yet being offered.

6.4. *Radiation Measurements on Trunk and Distribution Cables*

The final criterion of the efficiency of a screen is obtained by measurement of the radiated field from the cable for a given input voltage. Comparative tests have recently been carried out on various cables under known conditions, and these have shown that, for input voltages up to 3 volts, the radiated field at distances of greater than six inches from the cable is less than 1 microvolt per metre for both the cables with double braided and longitudinal overlapped copper tape outer conductors. By comparison, the radiation at a distance of 12 inches from a single "close-weave" braided cable with the same input voltage is of the order of 100 microvolts per metre. Thus for input voltages up to 3 volts, which covers most practical cases, cables with double braids and longitudinal copper tape outer conductors give comparable screening performance and the general level of radiation from such cables also appears to be satisfactory. It is, therefore, unlikely that the more rigid seam welded or drawn tubes will be necessary for most cases.

Table 2

Characteristics of Distribution Cables

Inner Conductor diam. (inches)	Overall diam. (inches)	Type of dielectric	Type of screen	Maximum attenuation dB/100 ft		Minimum setting radius (inches)	Screening efficiency (dB)	
				100 Mc/s	200 Mc/s		30 Mc/s	100 Mc/s (estimated)
0·048	0·290	Cellular polythene	Single braid	2·3	3·4	½	—	—
0·048	0·369	Cellular polythene	Double braid	2·3	3·4	½	20	15
0·064	0·485	Cellular polythene	Double braid	1·75	2·6	½	20	15
0·056	0·394	Fin and tube	Double braid	1·85	2·7	1	20	15
0·056	0·374	Cellular polythene	Longitudinal corrugated copper tape	1·7	2·5	1	20	15
0·044	0·340	Cellular polythene	Aluminium sheath	2·3	3·4	1½	>140	>140

7. General Conclusions

The development of a longitudinal corrugated copper outer conductor in conjunction with a tough cellular polythene dielectric appears to offer the best compromise between attenuation, screening performance and flexibility. In addition, this type does not allow direct ingress of moisture and is the most economical construction for a given overall diameter. If, however, there are situations where exceptional screening is demanded, then the aluminium-sheathed version offers the best solution at present. Alternatively, if flexibility becomes of prime importance, then double-braided constructions are available with semi-airspaced dielectric to offset the increase in attenuation. In this case precautions are necessary to avoid moisture ingress.

These remarks apply particularly to cables for systems with a top frequency of the order of 200 Mc/s. For the narrow band systems with top frequencies of the order of 100 Mc/s smaller cables are possible and it is not necessary to have recourse to a semi-air-spaced dielectric construction.

8. Future Trends

Development has already proceeded fairly rapidly in this field and it is therefore difficult to predict the more long-term course of events. There is, however, a tendency to go to systems operating at higher frequencies as the available frequency spectrum is limited in Bands I and III. If this is the case, cables with lower losses will be required, which will inevitably lead to larger cables. The problem will then be mainly one of retaining or improving on the present mechanical properties, whilst maintaining the same degree of screening. The advent of new or improved materials having lower dielectric loss angles, in constructions not allowing direct ingress of moisture, could be of great significance in minimizing the increase in attenuation at the higher frequencies, and development work continues on this aspect.

9. Acknowledgments

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10. References

1. Willis Jackson, "High Frequency Transmission Lines". (Methuen Monographs, London, 1947.)
2. W. T. Blackband and D. R. Brown, "The two-point method of measuring characteristic impedance and attenuation of cables at 3000 Mc/s", *J. Instn Elect. Engrs*, **93**, Part IIIA, pp. 1383–6, 1946.
3. S. A. Schelkunoff, "The electromagnetic theory of coaxial transmission lines and cylindrical shields", *Bell Syst. Tech. J.*, **13**, pp. 532–79, October 1934.

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The Evaluation of Oxide-Cathode Quality by Shot-Noise Tests

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Summary: As is well known, the space-charge reduced shot noise of a triode can be used as a measure of the cathode emissivity. A simple test equipment is described and results of measurements are discussed. The noise tests seem to be more sensitive than other methods without having any detrimental influence on the valve under test. Application for life test is suggested.

1. Introduction

The routine emission test[10] which is carried out on mass-produced valves may have undesirable effects on the life of the valve under test. Because of the high value of emission current normally drawn from the cathode, the testing period must be reduced to a minimum which is sufficient to allow the stabilization of that current. Even then it may impair the valve and influence the life expectations.

The value of pulse tests or alternating-voltage tests which are used in lieu of the simple d.c.-saturation test is debatable. The pulse operation may decompose the barium layer normally deposited at the inner grid at the end of the activation and ageing process shifting the grid bias because of the change in contact potential towards negative values and rendering the valve characteristic unstable. Further, in practice, there is no exact and simple correlation between a.c. or pulse tests and d.c. tests.

The cathode emission is only slightly affected by the measuring procedure if the variation of mutual conductance with cathode temperature is accepted as a measure of the cathode emissivity. The transconductance is measured by any of the standard methods while the heater voltage is gradually reduced. For properly activated and aged cathodes, the transconductance and anode current decrease at a lower rate than for cathodes with poor emission.

From the variety of tests which employ underheating of the cathode for the purpose of evaluation of cathode emission, we found particularly useful the one in which a large sinusoidal signal is applied to the first grid of the triode-connected valve under test; whilst reducing cathode temperature, the amplified signal is observed in the anode circuit of the valve under test on the screen of an oscilloscope. At a certain heater voltage a characteristic distortion appears on the positive half-cycles, as a sign that the emission is becoming saturated.

All these tests are cumbersome and time-consuming for routine inspection, further, they have the disadvantage that it is rather difficult to arrange them as simple GO/NO-GO tests and much is left to the skill and conscientiousness of the operator.

During the development of low-noise long-life amplifier tubes, it has been observed that shot-noise measurement is a more sensitive test than the usual means for evaluation of cathode quality. Further, it is suitable for routine tests without having any detrimental effects on the valve under test. Activation and ageing processes can be quickly established and excellent correlation has been found between shot-noise tests and other cathode tests. In life tests the decrease of transconductance is usually the criterion of life, which can be replaced by a shot-noise deterioration figure and prediction of valve life can be made by measuring shot noise after a relatively short ageing period.

A simple test equipment, which utilizes well-known principles, will be described, together with the interpretation of the results, and comparison is made with other methods.

2. Theory

The mechanism of shot noise in the region of medium-frequencies where transit-time effects are negligible is fairly well understood,[1-5] and close agreement between theory and results obtained has been observed.[7, 8] An excellent and up-to-date treatment of the fundamentals and physical mechanism together with extensive bibliography has been given by Bell.[6] Application of noise measurements for cathode tests was the subject of a more recent communication.[9] To recall the fundamental equations, the mean-square shot noise current in a temperature limited diode or negative grid triode can be expressed as:

$$\overline{i_s^2} = 2eI_a\Delta f \qquad \ldots\ldots(1)$$

where e is the electron charge, I_a the d.c. anode current and Δf the frequency band in which noise is measured.

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† New Electronic Products Ltd., London, S.E.11.

The space charge reduces this to

$$\overline{i_s^2} = 2eI_a F^2 \Delta f \qquad \ldots\ldots(2)$$

where the space-charge reduction factor F^2 can be expressed as

$$F^2 = 1{\cdot}28\,\frac{kT_c}{e}\,\frac{g_m}{\sigma I_a} \qquad \ldots\ldots(3)$$

with

$$\sigma = \frac{1}{1 + \dfrac{1}{\mu}\left(1 + \dfrac{4}{3}\dfrac{d_{ga}}{d_{cg}}\right)} \qquad \ldots\ldots(4)$$

where μ is the static amplification factor and d_{ga} and d_{cg} the grid-plate and grid-cathode distances respectively, k the Boltzmann constant, T_c the cathode temperature and g_m the mutual conductance.

The anode shot-effect noise current can be represented as an equivalent noise voltage generator in series with the grid, or by an "equivalent grid resistance" in series with the grid of an ideal tube having no anode shot-effect noise. The value of $R_{\text{eq.}}$ can be calculated with sufficient accuracy for practical purposes from the following equations:[3]

for triodes

$$R_{\text{eq.}} = \frac{2{\cdot}5}{\sigma g_m} \qquad \ldots\ldots(5)$$

for pentodes

$$R_{\text{eq.}} = \frac{2{\cdot}5}{\sigma g_m}\,\frac{I_a}{I_a+I_{g2}} + 20\,\frac{I_a}{g_m^2}\,\frac{I_{g2}}{I_a+I_{g2}} \qquad \ldots\ldots(6)$$

For purposes of comparison of valves the concept of $R_{\text{eq.}}$ is particularly useful. On the other hand, especially for cathode test it is convenient to express the space-charge-limited shot-noise current in terms of temperature-saturated current generating equal noise.

3. Test Equipment

The noise-measuring equipment is basically a sensitive, high-gain low-noise amplifier, consisting of preamplifier, oscillator, mixer, demodulator and audio frequency amplifier stages. Essential parts of the input stage are shown in Fig. 1.

The valve under test is mounted on an interchangeable socket and fed from well-filtered d.c. supplies with appropriate voltages. All voltages and currents are continuously monitored. Three centre frequencies of 1·1 Mc/s, 2·1 Mc/s and 11 Mc/s can be selected by adjusting the anode load of the valve under test and the local-oscillator frequency. A saturated diode with pure tungsten filament can be switched in the position of the valve under test and by adjusting the heater voltage of the diode, the noise of the valve under test can be expressed in saturated diode current, or if the mutual conductance is known, in equivalent noise resistance. Facilities are provided to measure mutual conductance *in situ*. The pre-amplifier is of conventional construction for 1·2 and 2·1 Mc/s and a cascode-connected double triode is employed for 11 Mc/s. The amplified noise is fed into mixer and i.f. amplifier stages of 470 kc/s centre frequency and about 25 kc/s band-width and followed by demodulator and a.f. amplifier stages. A d.c. galvanometer is inserted into the circuit of the demodulator and used as an output meter. As this output meter is used as an indicator only, it is not essential to employ a thermocouple or other power measuring device. The a.f. amplifier has a flat response from 400 c/s to 20 kc/s, and a sharp cut-off at low audio frequencies. As it is desirable to keep the noise of the amplifier as low as possible, specially selected low-noise tubes are used in the input stage of the preamplifier.

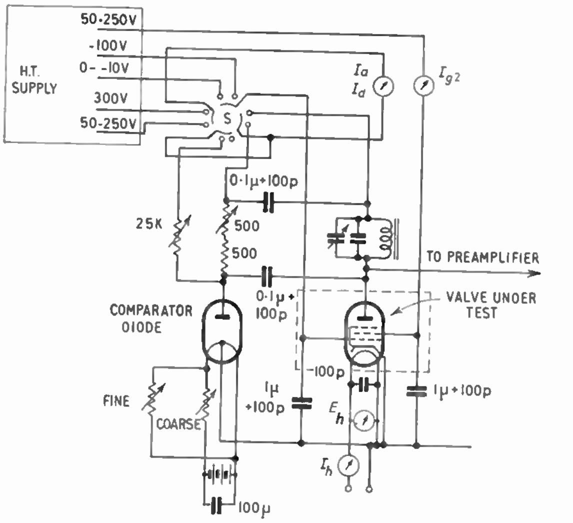

Fig. 1. Essential elements of the input circuit of the measuring amplifier.

The valve under test is placed into the interchangeable input socket and appropriate electrode voltages are adjusted and currents checked. The valve noise is read in arbitrary units on the output meter. Next, a high negative voltage is applied to the first grid of the valve under test, cutting off its anode current, and the heater voltage of the calibration diode is raised until the previous reading is obtained. The ratio of diode current to valve anode current gives the F^2 factor. The equivalent noise resistance is

$$R_{\text{eq.}} \simeq 20\,\frac{I_d}{g_m^2} \qquad \ldots\ldots(7)$$

where I_d is the diode current which generates the same noise as the valve under test. Care is taken to eliminate errors which arise from capacitance and resistance changes between operating and cut-off conditions of the valve under test.

Proper functioning of the equipment can easily be checked by placing saturated diodes in both 'Valve under Test' and 'Diode' positions and ensuring that equal currents produce equal noise output. The a.f. amplifier proved to be very useful for routine measurement as the operator was able to adjust the diode current to produce a noise equal to that of the valve under test purely by listening to it. Further, it made detectable any noise originating from external disturbances. Little difference has been found between values measured at 1·1, 2·1 or 11·1 Mc/s centre frequencies on well-activated tubes although it was assumed that flicker noise is not negligible even at 1 Mc/s.

4. Equivalent Noise Resistance of Receiver Valves

Early experiments were conducted with the aim of comparing the measured shot-noise figures for various types of valves with that of the calculated values. Some of these results are shown in Table 1. The measured values represent averages of 50 samples taken at random from various batches.

Generally speaking, there is a very good agreement between measured and calculated values but for high-current high transconductance triodes and pentodes, the measured $R_{eq.}$ values are 5 to 30% higher than those calculated. This is not the consequence of the lower-than-nominal transconductance, because if the results are expressed in equivalent saturated-diode current, the trend remains the same.

Table 1

Comparison between Calculated and Measured Values of $R_{eq.}$ for Various Types of Valve

Type	V_a (volts)	V_{g2} (volts)	I_a (mA)	I_{g2} (mA)	R_k (ohms) $-V_1$ (volts)	g_m (mA/V)	$R_{eq.}$ meas'd (ohms)	$R_{eq.}$ calcul'd (ohms)
1R5	90	67·5	1·6	3·2	0	0·3	210 000	240 000
1S5	67	67·5	1·6	0·4	0	0·62	20 000	21 000
1U5	67·5	67·5	1·6	0·4	0	0·062	20 000	21 000
6AC7	250	150	10	2·5	160Ω	9	760	760
6AK5	180	120	7·7	2·4	200Ω	5·1	2000	1900
6AT6	250	0	1	0	3	1·2	2400	2400
6AU6	250	150	10·6	4·3	68Ω	5·2	2400	2600
6BA6	250	100	11	4·2	68Ω	4·4	3600	3500
6BE6	250	100	2·9	6·8	0	0·475	180 000	180 000
6SJ7	250	100	3	0·8	3	1·65	6000	6100
ECH4	250	100	3	6·2	2	0·75	58 000	55 000
ECH21	250	100	3	6·2	2	0·75	58 000	55 000
ECH42	250	85	3	3	2	0·75	90 000	100 000
EF8	250	250	8	0·2	2·5	1·8	2800	2400
EF9	250	100	6	1·7	2·5	2·2	6800	6500
EF22	250	100	6	1·7	2·5	2·2	6600	6300
EF42	250	250	10	2·4	2	9	800	760
EF80	250	250	10	2·8	2	6·8	1200	1300

Having observed that the noise of some valves is considerably higher than the average, the defective samples were carefully tested again with the result that on about 75% of all defective valves, poor cathode emission was the cause. By further ageing at slightly elevated cathode temperature some noisy valves improved, others remained still noisier than the average.

It may be of interest to note that after 50 hours ageing of 100 samples of each type listed below, the equivalent noise resistance was found to be twice the average or more in the percentages shown:

6AK5	12%
EF22	4%
6BE6	2%
6BA6	2%
EF42	2%
6AC7	1%
EF80	1%
ECH21	1%

5. Cathode Tests

In order to allow a closer study of the correlation between valve noise and cathode emission, two different types of experimental triodes have been built with the following main characteristics:

	T1	T2
Heater voltage	6·3 V	6·3 V
Heater current	0·3 A	0·2 A
Anode current	8 mA	3 mA
Transconductance	6·8 mA/V	0·8 mA/V
Amplification factor	62	48
Control grid voltage	−1·2 V	−1 V

T1 utilized the elements of an ECC81 triode, but the cathode was made visible on the finished valve for the purpose of optical measurement of the cathode temperature. T2 was a low-transconductance triode of plane-parallel construction with d_{cg} = 170 μm and d_{gp} = 500 μm. The cathode assembly was constructed in a way to allow easy replacement and on some samples noise tests were carried out during the activation process whilst the valve was being pumped. These last experiments proved of little practical value owing to the difficulty experienced in applying optimum activation temperature and time data to valves in current production. The reason is fairly obvious: the complex process which takes place during the activation depends a great deal on the gas-liberation process from the other electrodes, the formation of barium-oxide layer on the inner grid, etc., and results obtained on one type can hardly be applied to another type with slightly differing characteristics.

A typical result obtained on a sample of type T1, with a well-activated cathode is shown in Fig. 2,

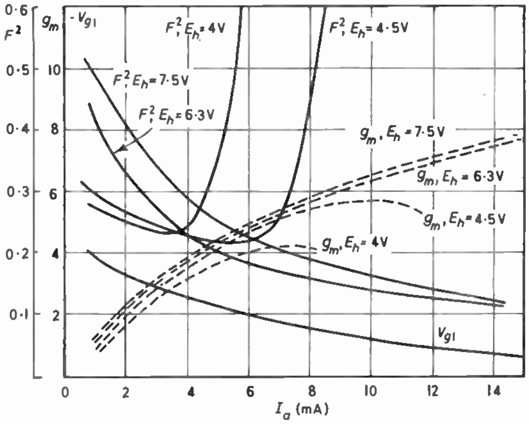

Fig. 2. Measured values of the space-charge reduction factor F^2 and mutual conductance g_m versus anode current I_a for various heater voltages on type T1 experimental triode. Averages of 5 samples.

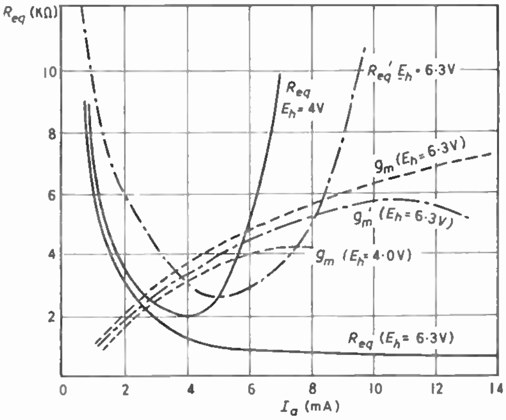

Fig. 3. Comparison between well-activated (g_m, $R_{eq.}$) and poorly activated (g_m', $R_{eq.}'$) type T1 experimental triodes. Equivalent noise resistance and mutual conductance are plotted against anode current for 4 and 6·3 V heater voltages.

where the space-charge reduction factor is plotted against anode current for various heater voltages.

Both the trend and value of the $F^2 = f(I_a)$ curves are in reasonably close agreement with the theoretical value. Although the nominal anode current is 8 mA, even with 14 mA there is no sign of saturation and the space-charge reduction factor is about 0·15, as expected. Decreasing the cathode temperature by reducing the heater voltage to 4·5 and 4 V respectively causes a sudden and marked increase in F^2 as some elements of the cathode are approaching saturation. For the purpose of comparison, the mutual conductance of the valve is measured at the same heater voltages and anode currents, exhibiting the well-known shapes. It should be noted that at $E_h = 4·5$ V and $I_a = 8$ mA, the drop in mutual conductance is hardly noticeable, whereas the increase in noise shows a sharp rise from 7 mA onwards. The same effect is apparent at 4 V heater supply. The noise test indicates the approach to saturation earlier than the transconductance test.

Comparative figures for well-activated and poorly activated cathodes are shown in Fig. 3. The optimum activation process for T1 type valves was found to be 2 minutes at 14 V, followed by 15 minutes at 13 V, without current drain. The sealed-off valve was then aged with the application of normal electrode voltages and currents for several hours, effecting a small improvement in noise. On some valves the activation was carried out at 9 V only for a period of 10 minutes.

In Fig. 3, the valve noise is presented in terms of equivalent noise resistance. For the well-activated valve with an applied heater voltage of 6·3 V, $R_{eq.}$ decreases gradually from 1000 Ω to 500 Ω and the mutual conductance increases from 4·8 to 7·5 mA/V with an increase in anode current from 6 to 14 mA. If the heater voltage is reduced to 4 V, the trend in $R_{eq.}$ is reversed and a sharp increase occurs beyond 4 mA current.

The $R_{eq.}$ of the poorly activated valve is about 3000 Ω at 6 mA, compared to the 1000 Ω of the well-activated ones, and increases rapidly with increasing current; at 10 mA the values are 700 Ω and 11 000 Ω respectively. It should be noted that at normal cathode temperature the mutual conductance of the poorly activated valve is hardly lower than that of the good one, even at 10 mA anode current. If the well-activated cathode is run at an applied heater voltage of 4 V, the shape of $R_{eq.}$ versus I_a curve reveals the saturation of the emission before that could be noticed on the relevant g_m curve. In Fig. 4, results of experiments carried out on the plane electrode triodes T2 are displayed. Anode current I_a, mutual conduc-

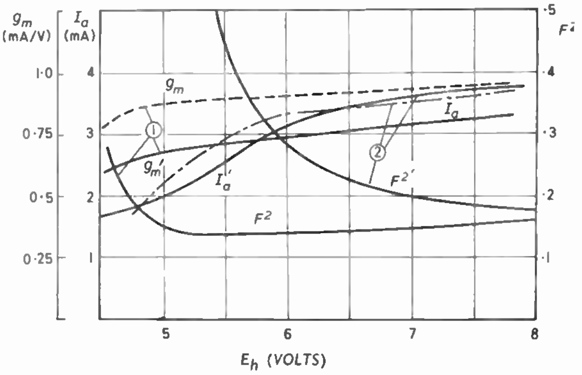

Fig. 4. Transconductance, anode current and space-charge reduction factor of well-activated (g_m, I_a, F^2) and poorly activated (g_m', I_a', F'^2) type T2 experimental triodes versus heater voltage. Average of 5 samples.

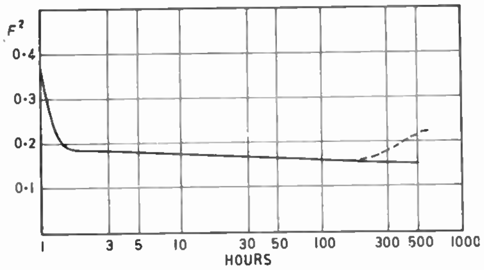

Fig. 5. Noise test on 6AK5 valves during life test. Average of 20 samples. Operating conditions: $V_a = 180$ V, $V_{g_2} = 120$ V, $I_a = 7{\cdot}6$ mA, $I_{g_2} = 2{\cdot}3$ mA, $R_k = 200$ ohm, $E_h = 6{\cdot}3 \pm 0{\cdot}05$ V.

tance g_m and space-charge reduction factor F^2 are plotted against cathode temperature for well-activated and poorly activated samples.

Figure 4 illustrates further the well-known fact that the dependence of anode current on cathode temperature is higher for the poorly activated cathodes than for well-activated ones, although as in the example, the current at nominal heater voltage may be higher for the poorly activated cathode. The correlation between g_m and F^2 is again obvious.

6. Noise and Life Tests

Some limited experiments have been carried out concerning the value of noise measurement in life tests. A quantity of twenty type 6AK5 valves was taken out of a large batch undergoing life test, and given a shot-noise test with measurements taken at intervals of 50, 100, 200, 300 and 500 hours of ageing. The results are plotted in Fig. 5.

Up to the first 300 hours nearly all samples show a slight improvement in emission, and thereafter 18 of the samples showed no further change. The remaining two samples deteriorated slightly. Subsequently the life of these samples proved to be about 30% less than average. Unfortunately, noise tests were discontinued after 500 hours.

The collected data are of course insufficient for any firm conclusion, but previous experience suggests that noise tests may be valuable in prediction of life expectancy and instead of the usual criterion of valve life which is a certain drop in mutual conductance, the shot-noise deterioration figure may be introduced. As this becomes apparent after a shorter ageing period than the indication given by drop in g_m, valuable time and equipment may be saved.

7. Conclusion

The shot-noise test for estimating the emissivity of oxide cathodes seems to be more sensitive than the other methods. Contrary to other tests it does not influence the cathode, being a particularly valuable feature if the emission of the same valve has to be measured repeatedly, as in life tests. It can be applied among other purposes for establishing optimum activation procedure. A simple test gear can be built, and as only direct voltages and currents have to be measured, its calibration is simple and permanent.

8. References

1. W. Schottky, "Spontaneous current fluctuations in various conductors", *Annalen der Physik*, **57**, pp. 541–67, 1918.
2. W. Schottky and E. Spenke, "Space charge reduction of shot noise", *Wissenschaft Veröff Siemens-Konzern*, **16**, pp. 1–41, 1937.
3. B. J. Thompson, D. O. North, and W. A. Harris, "Fluctuations in space-charge limited currents at moderate high frequencies", *R.C.A. Review*, **4**, pp. 269–85, 441–72, 1939; **5**, pp. 106–24, 244–60, 371–88, 1940–41; **6**, pp. 114–24, 505–24, 1940–41.
4. H. Rothe and W. Kleen, "Elektronenröhren als Anfangsstufenverstärker", pp. 291–9 (Akad. Verlag, Leipzig, 1953).
5. A. J. Rack, "Effect of space charge and transit time on the shot noise in diodes", *Bell Syst. Tech. J.*, **17**, pp. 592–616, 1938.
6. D. A. Bell, "Electrical Noise", pp. 88–184 (Van Nostrand, New York, 1960).
7. J. Dalbert, "Background noise of oxide cathodes", *Le Vide*, **9**, pp. 200–2, 1954.
8. L. G. Sebestyen, "On the shot noise of receiver and mixer valves", *Magyar Hiradàstechnika*, **7**, No. 1, pp. 8–20, 1956.
9. W. Dahlke and F. Dlouhy, "A cathode test utilizing noise measurements", *Proc. Inst. Radio Engrs*, **46**, pp. 1639–45, September 1958.
10. Standardized in JAN-1A Specification and described, for instance by J. F. Blackburn, "Components Handbook", M.I.T. Radiation Laboratory Series, Volume 17 (McGraw-Hill, New York, 1949).

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Radio Engineering Overseas . . .

The following abstracts are taken from Commonwealth, European and Asian journals received by the Institution's Library. Abstracts of papers published in American journals are not included because they are available in many other publications. Members who wish to consult any of the papers quoted should apply to the Librarian, giving full bibliographical details, i.e. title, author, journal and date, of the paper required. All papers are in the language of the country of origin of the journal unless otherwise stated. Translations cannot be supplied. Information on translating services will be found in the Institution publication "Library Services and Technical Information".

BRANCHED WAVEGUIDES

A recent German paper discusses the case of two parallel waveguides whose partition wall is discontinued from a certain point onward. If an H_{01}-mode enters one of the two waveguides, its power is scattered in the following way: one part is reflected into the primary waveguide, a second part is diffracted around the edge and returns along the adjacent waveguide, and the third part continues as propagating mode in the common waveguide of double width. This problem can be formulated mathematically by an infinite set of linear equations which expresses the mutual interaction of an infinite number of mode types. This set of equations is solved approximately with an electronic computer. The complementary problem, where a mode enters a wide waveguide and is split into two parallel waveguides of half the width, is discussed by the same method. The results of these two problems are used for calculating the coupling between two parallel rectangular waveguides which are coupled by a slot in the partition wall of an arbitrary length *l*. With its application in frequency separating and combining filters in mind the paper discusses the case in which the power divides evenly between the two outgoing waveguide sections.

"Electromagnetic waves in branched rectangular waveguides", H. Kaden. *Archiv der Elektrischen Ubertragung*, **15**, pp. 61–70, February 1961.

FERRITE CORE SWITCHING

Analysis of the behaviour of magnetic switching materials having rectangular hysteresis loop properties leads to expressions too complex to handle in practical circuit design. On the other hand, a rough approximation may render possible design procedures leading to quick results, but experimental corrections may be required at a later stage. In a Dutch paper a solution is presented in which the derived analytical expressions are in close agreement with the physical behaviour of the material and are at the same time fairly easy to apply to circuit design. Resistive and inductive loading of the cores is discussed, and also the arrangement of cores to drive a line of cores in a storage stack, current sources being used to drive the cores. The operation of voltage-driven cores is also described.

"Analysis of ferrite core switching for practical applications", P. A. Neeteson. *Electronic Applications, Eindhoven*, **20**, No. 4, pp. 133–52, 1959–1960. (In English.)

BACKWARD WAVE OSCILLATOR

A delay line with circular fingers has been developed at the Berlin Technical University for use in a backward wave oscillator operating in the frequency range 10 000–20 000 Mc/s. Its advantage is a stronger stray field with which the electron beam can interact. A particularly short and effective field absorber made from graphite is used as a reflectionless termination for this delay line. For the purpose of a more effective utilization of the stray field, an electron gun for producing bent ribbon beams is used. The effect of space charges, incident angle of the electrons and magnetic field strength on the cross-section of the beams is discussed. Operational measurements have revealed a good agreement between the calculated and measured frequency and transient current characteristics.

"A backward wave oscillator for the frequency range 10 000–20 000 Mc/s with a delay line consisting of circular fingers", W. Wendrich. *Nachrichtentechnische Zeitschrift*, **14**, pp. 203–11, April 1961.

SUPER-GAIN ANTENNA

The super-gain principle states that it is theoretically possible to obtain a gain of any magnitude with an antenna of any small dimensions provided no limits are given for the number of the individual radiators forming the antenna system. Furthermore no limiting conditions must be imposed for the amplitude of the current and the phase of the feed. However, calculations show that the feeding of the radiators becomes extremely difficult even for a reasonable degree of "over-designing". Some calculated values of small super-gain antenna have been given in the literature. For the purpose of illustrating the conditions for slightly larger antennas several values for "over-designing factors" have been used by a German engineer in calculations relating to antennas with length of 6λ. Three end-fire and side-fire arrangements have been studied. Radiator spacing was 0·43 λ, 0·3 λ and 0·2 λ and the number of radiators was 15, 21 and 31.

"Supergain antenna", Th. Heller. *Nachrichtentechnische Zeitschrift*, 14, pp. 113–8, March 1961.

MOBILE RADIO NETWORKS

In civil mobile radio systems, several receiving stations are usually installed at different sites to compensate for dead points of the service areas caused by the unbalance of transmitting powers of fixed and mobile stations. In such systems a receiving station with highest signal/noise ratio should be instantaneously selected at all times. Two new all-electronic methods are proposed by a Japanese engineer for such systems: a switching system and a combining system respectively. As the result of experiments it has been shown these systems operate satisfactorily, even in cases where signal/noise ratios at several receiving stations fluctuate at speeds of 100 c/s, without any transient effects.

"Automatic selection of fixed receiving stations in mobile radio systems", T. Morinaga. *The Journal of the Institute of Electrical Communication Engineers of Japan*, **44**, pp. 11–5, January 1961.