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THE STATUS OF THE ENGINEER

CURRENT problems in Great Britain centre largely on the engineering industry. In comparison with the economic issues involved, argument on nomenclature may seem unimportant; it is nevertheless opportune once more to express regret that the term *engineer* should haphazardly be given to anyone engaged in the engineering industry, irrespective of whether he be a highly qualified head of a development department, or an unskilled or semi-skilled machine operator.

The problem of securing wider and better understanding of the status of the engineer is not a new one. In our own Institution the the problem was ventilated by Sir Louis Sterling in his Presidential Address just 15 years ago.* After reviewing the need, in wartime, for the registration of engineers, Sir Louis Sterling examined the means by which the status of the engineer might be more clearly understood. He emphasized that many felt that proper appreciation of the title *engineer* might ". . . come about in time through the general practice of those engaging engineers. That is probable, but it implies the need to educate the public and employers . . ."

A report given to the Parliamentary and Scientific Committee† stressed the professional status enjoyed by the engineer in the U.S.S.R. In many other countries, notably Canada, there is also legislation for the registration of engineers, but it seems most unlikely that such legislation will be promoted in Great Britain in the near future. Indeed, it is argued that a law for the registration of engineers might not be desirable; to be effective it would certainly need a very large measure of agreement between all the professional bodies whose members should be covered by such legislation.

Such agreement may not be easily obtainable. The exigencies of war demanded some form of registration, but, as Sir Louis Sterling pointed out, ". . . many alterations and amendments (were) made in the compilation of the wartime Central Register because, although excellent in principle, the Register did not at first take into consideration the specialist aspects of each and every profession—especially in the engineering industries. Consequently there was almost a fatal tendency to generalize *instead of specialize.*"

Whilst the medical profession, for example, is able to maintain its distinctive status and yet afford its members varying degrees of specialization, the younger profession of engineers has not yet been wholly able to resolve its differences of opinion on the need for specialization. Thus, although it has been argued that the granting of Charters to engineering bodies will hasten the process of securing recognition of the professional engineer, there is very little cohesive action in that group towards securing wider public recognition of the status of the engineer.

It cannot be denied, however, that there is major agreement about the basic training of an engineer. This is reflected in the subjects common to the examinations of our own and other Institutions. Possibly the main reason for lack of greater co-operation is inability to agree on the extent of specialization—one may put it as finely as that of being unable to agree that the physician, surgeon or gynaecologist, is entitled to come under the general description of being a member of the medical profession!

* *J.Brit.I.R.E.*, 3, pp. 33-36, September 1942.

† December 1956. The Parliamentary and Scientific Committee, 31 Palace Street, Westminster, London, S.W.1.

"ELECTRONICS IN AUTOMATION"

First List of Papers Selected for Presentation at the 1957 Convention

Synopses of papers in the other four sessions and the remaining papers to be included in Sessions 1 and 5 will be published in the April and May issues of the Journal

Session 1—OFFICE MACHINERY AND INFORMATION PROCESSING

(Thursday, June 27th—9.15 a.m. to 12.30 p.m.)

The Development of a Business Computer System

by A. St. Johnston, B.Sc., and S. L. H. Clarke, B.A. (Elliott Bros. (London) Ltd.)

The shortcomings of the scientific type of digital computer when applied to data processing jobs in general, and the business accounting type in particular, has led to the evolution of an electronic digital system more in tune with the requirements. The basic logical arrangement of the Elliott 405 system is outlined, followed by a more detailed description of the units. Particular reference is made to a new type of store using 35-mm film coated with magnetic oxide. Two specific applications of the use of the system are quoted as examples.

Ferroresonant Circuits for Digital Computers

by D. A. Bell, Ph.D., M.A., and C. B. Newport, B.Sc. (Electrical Engineering Department, University of Birmingham)

The characteristics of the ferroresonant circuit are analysed and conditions for bistability obtained. Problems of high frequency operation are considered and the composition and configuration of suitable ferrites discussed. A carrier frequency of 1 Mc/s seems to be the present limit. Practical circuits are presented for a shift register, 3-stage binary counter and 2-input logical adder.

Automatic Reading of Typed or Printed Characters

by C. E. G. Bailey, M.A. (Solartron Electronic Group Ltd.)

In order to enable the high input capacity of computers to be economically used in typical business applications, the automatic reading of conventionally printed or typed documents is required. The Solartron ERA (Electronic Reading Automaton) is designed to this end.

Typed matter typically suffers from mutilation, blurring, and displacement. To achieve reliable recognition, the input signal, produced by a flying-spot scanner, is first "cleaned up" and then passed on to a temporary store as a series of "black" or "white" units. An extensive process of "AND" and "OR" logic transforms these to a line-per-character output, from which binary or other transformations may be made. To overcome displacement, a "pre-scan" determines the position of the character and applied shifts to the reading scan.

The pilot model of ERA reads numerical data at 120 characters/sec. The production model is planned to read 200-300 characters/sec, including some of the alphabet. The input can be continuous roll, microfilm, or separate documents; the output can be punched cards, punched or magnetic tape, sorting, or direct to a computer.

Integration of Computers with Factory Processes

by A. H. Cooper (E.M.I. Electronics Ltd.)

Manufacturing processes in which each unit of the product may need individual treatment can be integrated with, and controlled by, computers. At each decision point in the flow of the product through the factory, the computer takes account of the overall production plan, the conditions in the neighbourhood of the decision point, and the past history of the product-unit involved.

This application of computers involves more transfers of information than in a mathematical or business computer, but the speed at which the information must flow is much slower. Hence, the preferred basis and form of the computer are unusual.

The paper describes typical computers for this class of application and the manner in which they fit into the requirements.

Session 5—SIMULATORS

(Saturday, June 29th—9.15 a.m. to 12.30 p.m.)

The Use of Radar Simulators in the Royal Navy

by P. Tenger, B.Sc. (Admiralty Signal and Radar Establishment)

This paper will describe an air defence tactical trainer developed for the Royal Navy and designed for tactical problems involving several ships and aircraft disposed over an area of some 500 miles radius. The equipment is based on an idealized narrow beam radar. Individual radar pictures are generated for each of the ships to enable them to "see" all targets within range. Complete realism is not aimed at since the cost and complexity of the equipment involved would be excessive, but features adding to the realism are included. These special circuits and the basic circuits for the generation of the radar and tactical pictures will be described in the paper.

An Analogue Computer for Fourier Transforms

by Professor D. G. Tucker, D.Sc. (Electrical Engineering Department, University of Birmingham)

The paper gives the basic principles of a computer for obtaining a graphical display of the Fourier Transform of a function which can be represented by a finite number of ordinates over a finite range of the input variable. A delay line with linear phase-shift/frequency response is used to give the exponential term in the transform, and a time-variation of carrier frequency in sympathy with the time-base of the display unit gives the required variation of the index of this term.

The Application of Analogue Computer Techniques to the Design of Aero Engine Control Systems

by A. D. Jeffery and J. M. Dempsey (Elliott Bros. (London) Ltd.) and H. Saville, B.Sc.(Eng.) and K. Gill, B.Sc. (Armstrong-Siddeley Ltd.)

The problem of aero engine control systems is shown to have reached the degree of complexity at the present time that makes it necessary to embark on analogue methods of simulation in order to predict the behaviour of the system. The parameters to be controlled are discussed and the way in which interaction may affect the individual loops is explained together with a review of the requirements of a control system from the pilot's handling, and the performance view points. As illustrations, examples of control system functional components are used to illustrate the methods of adaptation of a practical hydraulic system to analogue methods. The use of the computer results is discussed in formulating an optimum system design.

In order to show how systems can be simulated the concept of transfer functions will be outlined. This method of establishing relations between points in the computer will be described, and mention made of special units used in simulation including a time delay unit of particular importance in engine control problems. Other aspects of control system development are discussed and the place of analogue computer technique in the overall picture and its increasing importance in this field is shown.

The Determination of System Transfer Functions from Normal Operating Data

by J. G. Henderson, B.Sc., and C. J. Pengilley, B.E. (Electrical Engineering Department, (University of Birmingham)

It is often not possible to remove a plant or system from service in order to measure its transfer function by the usual process of applying sinusoidal or impulse test signals. Use can, however, be made of statistical data of input and output when the system is operating in normal service, since in a linear system, undisturbed by noise, the cross-correlogram between the input and output signals is given by the convolution of the system weighting function and the auto-correlogram of the input.

The experimental determination of the system transfer function or weighting function involves finding the function which, when convoluted with the auto-correlogram of input, will most closely fit the cross-correlogram between the input and output signals. The paper discusses various methods of doing this, with particular emphasis on analogue methods.

Registration forms have now been sent to all members. Those intending to attend the Convention are urged to return their forms as soon as possible.

MEMBERS OF THE PAPERS COMMITTEE

Born in 1917, **Ieuan Maddock** was educated at Gowerton County School and at the University of Wales, graduating in 1938 with 1st class honours in Physics. He was awarded a Research Studentship but his work in gas discharge phenomena was interrupted by the war.

Between 1940-1949 Mr. Maddock was concerned with electronic methods of measurement in armament research with the Ministry of Supply. In 1949 he was promoted to Principal Scientific Officer in charge of the instrumentation section of the Atomic Weapons Research Group; in 1952 he was appointed Assistant Director at the first atomic weapons trial at Montebello, where he was in charge of radio control, telemetry and high speed measurements, as well as the immediate pre-firing phase. He was appointed an O.B.E. in the 1953 New Year Honours List.

Since 1953 he has been Superintendent of Electronic Research at the Atomic Weapons Research Establishment (now part of U.K.A.E.A.) and last year was granted the rank of Deputy Chief Scientist in A.W.R.E. During the 1956 nuclear weapon trials he was responsible for all scientific aspects.

Mr. Maddock was elected an Associate Member in 1941 and was transferred to full Member in 1956. He has served on the Programme and Papers Committee since 1955.

George Percy Thwaites was born in London in 1915 and received his general education at Askes' Hatcham School. In 1934 he commenced his technical training with Standard Telephones and Cables Limited, North Woolwich, taking a sandwich course at Woolwich Polytechnic and gaining a National Diploma in Electrical engineering and the Woolwich Engineering Diploma in 1936. A year later he received his B.Sc.(Eng.) degree.

After completing his studentship training Mr. Thwaites went to the valve factory of Standard Telephones and Cables at Footscray where he was employed successively in production engineering, quality control and, from 1939-1947 as engineer of manufacture. He was then appointed a technical superintendent in charge of a number of departments which included inspection and production engineering. Following a company reorganization in 1949 he received his present post of head of the advanced plant development department where he is particularly concerned with special quality valves for Services use.

Elected an Associate Member in 1944, Mr. Thwaites was transferred to full Member in 1955. He has served on the Papers Committee since 1951 and has presented papers on valve subjects before Institution Sections.

Arthur George Wray, born in Retford, Nottinghamshire in 1924, was educated at Culford School, Bury St. Edmunds. He read natural sciences at Cambridge where he was resident in Emmanuel College, receiving his B.A. degree in 1944.

He then joined Marconi Instruments Limited where, in the Design and Development Section, he was responsible for the design of many communication instruments, particularly for f.m. equipment. In 1952 Mr. Wray was made editor of the Company's house journal and he has recently been appointed leader of the Advanced Development Group.

Mr. Wray was elected an Associate Member of the Institution in 1952 and joined the Papers Committee in 1953. He has read papers on electronic instrument design before Institution Sections and is the author of a paper on f.m. measuring equipment published in the *Journal* in July 1953.

BP/36

PRINCIPLES OF THE LIGHT-AMPLIFIER AND ALLIED DEVICES*

by

T. B. Tomlinson, B.Sc., Ph.D., (Associate Member)†

Read before the Institution in London on December 12th, 1956.

In the Chair: L. H. Bedford, C.B.E., M.A., B.Sc., Past President.

SUMMARY

The paper first considers the necessary component parts of a light amplifying system and briefly points out the advantages of using solid state devices. A system is outlined which consists of a photoconducting layer in series with an electroluminescent phosphor layer. The properties of the constituent parts are discussed in detail and the behaviour of the series combination is examined. Various practical constructions are described and attention drawn to the difficulties of manufacturing a picture reproducing device of large area.

Optical feedback depends on the spectral emission distribution of the electroluminescent layer and the spectral response of the photoconductor. An amplifier with optical feedback such that the loop gain exceeds unity can be triggered by a light pulse and then remains in the light emitting condition until the supply voltage is reduced. Such an electro-optical binary switch is described using ZnS and CdS powder layers or single crystals. Complex combinations of such units could be employed in shift registers, scaling circuits, etc., and novel circuits are suggested which employ light as the means of triggering and coupling. These devices would be of miniature size and need few, if any, associated conventional components. Limitations of switching time and possible future developments are also discussed.

1. Introduction

A light-amplifier system, taking its input power from an electrical supply must have two basic properties. These are:

(*a*) Provision of the output light either as uniform illumination or as a picture; in the latter case the device is also called an "image intensifier" and may be thought of as a multiplicity of units which amplify each picture element separately.

(*b*) Sensitivity to incident radiation so as to control the amount of electrical energy converted to light radiation.

It is possible to make a light-amplifier in which these two properties are combined in one material. Thus, thin films (about 10 microns) of ZnS:Mn have been synthesized[1] by evaporating Zn dust with a small percentage $ZnCl_2$ and $MnCl_2$ on to a hot glass substrate at 500–600° C in an atmosphere of H_2S. Such films luminesce when exposed to ultra-violet radiation or X-rays, and the application of a high field across the film results in an enhancement of the luminescence ("photo-electroluminescence"). Intensification of the image by a factor 11 times (photon gain) has been achieved for low intensity u.v. excitation. The emitted light is yellow-orange; the amplifier characteristic is fairly linear, but the response time is slow, varying from 0·01 sec at high light levels to several seconds at low levels. The most probable application of such a device would be in X-ray fluoroscopy; image intensification permits a valuable reduction in dosage, with consequent increased safety to the patient. A difficulty here is the low absorption of the X-rays in such a thin film and this largely counteracts the advantage gained from the field intensification. Another possible application is in cathode-ray tubes where the equivalent of post deflection acceleration can be achieved by making the phosphor screen in the form of

* Manuscript received 30th January 1957. (Paper No. 388.)

† Research Laboratories of The General Electric Company Limited, Wembley, England.

U.D.C. No. 621.383.4:535.37:621.318.57.

such a thin film. At present, the intensification of cathodo-luminescent images is much smaller than with u.v. excitation. Nevertheless, there is an advantage of a sharper spot and improved contrast since the normal microcrystalline screen diffuses the light over quite a large area.

An image intensifier in which the two basic properties are provided by separate materials can be made from a combination of a cathodo-luminescent screen and a photo-emissive surface. This system requires an evacuated envelope and is therefore bulky; it also needs high voltage supplies. In order to reproduce a picture, the photo-electrons must be kept to their appropriate paths and this requires a focusing system, usually provided by an enclosing magnetic focusing coil.

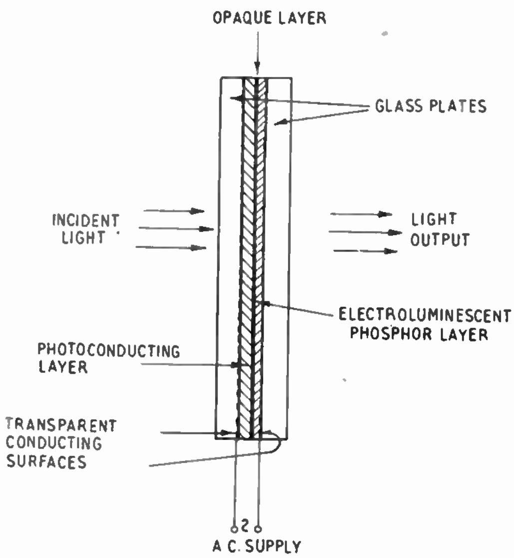

Fig. 1. Schematic diagram of a panel amplifier.

A simpler construction is achieved in the "panel amplifier" which is made up of a photoconducting layer superimposed on an electroluminescent layer, sandwiched between electrodes to which an alternating voltage is applied. (Fig. 1). This panel amplifier is the main topic of this paper.

2. The Panel Amplifier

The electroluminescent layer converts a.c. energy directly into light energy; the photo-conducting layer is electrically in series so that the voltage appearing across the electroluminescent layer, and therefore the light output, is determined by the incident light on the photoconductor. A low intensity image incident on the photoconducting layer may lead to the reproduction of the image, at an increased brightness level, by the electro-luminescent panel. In the dark regions of the incident image the dark current of the photoconductor is so low that only a small voltage appears across the phosphor layer in that region and there is negligible local light output. In regions of higher light intensity the increased photocurrent leads to a larger proportion of the applied voltage appearing across the phosphor layer with consequent increase of output light intensity. To determine the overall behaviour of the combination it is necessary to examine in detail the properties of the two parts.

2.1. *The Electroluminescent Layer*

The first interest in electroluminescence was as an alternative to fluorescent and filament lamps for general lighting purposes. In spite of considerable research effort, the light output is disappointingly low. There are several applications for which electroluminescent panels are particularly suitable; for instance as a panchromatic safelight of less than 0·1 ft-lamberts working from 240 V at less than 1 W, and 50 times more efficient than the conventional type. Another application is as "EXIT" and other signs in cinemas, etc. There was considerable stimulus to the study of electroluminescence because of possible television applications and the popular press has several times referred to an imminent television "picture on the wall." Thoughts turned to a colour system; green and blue phosphors are readily available, but a good red has not yet been produced so far as the author is aware.

An electroluminescent panel (Fig. 2) is manufactured by depositing a layer of the phosphor powder in a dielectric, which also serves as a binder, on to a glass base having a transparent conducting surface. Such a surface may be formed by blowing a jet of dry air, which has been bubbled through stannic tetrachloride, on to the glass surface which is heated to a temperature near the softening point. Water vapour in the atmosphere is essential to the action. The tin oxide layer, so formed, has a light absorption of some 15 per cent. and a surface resistance of about 500 ohms per square. The dielectric should have high permittivity, high dielectric strength and sound binding qualities. A layer formed by spraying a mixture of phosphor powder in an alkyd resin has been found satisfactory; more recent methods use a plastic such as styrene, or a vitreous dielectric. The second electrode is conveniently made from sprayed silver paint which combines good conducting and light reflecting qualities. A reflecting film of titanium dioxide has also been advocated.

The most common phosphor is ZnS activated by copper; Cl and Al have been used as co-activators and the addition of a trace of Pb is said to be beneficial. Copper activation can give blue and green electroluminescence. In ZnS:Cu,Cl phosphors the relative emission of the blue and green bands depends in a complex manner on the chlorine concentration (Fig. 3) and also on the temperature.[3] Addition of Mn yields an orange electroluminescence by an energy transfer process. A white panel is made from a suitable admixture of green, blue and orange phosphors.

The mechanism of electroluminescence[4] in phosphor powders is most probably due to the excitation of activator centres by collisions, the electrons being accelerated across a thin barrier layer on the surface of the crystals.

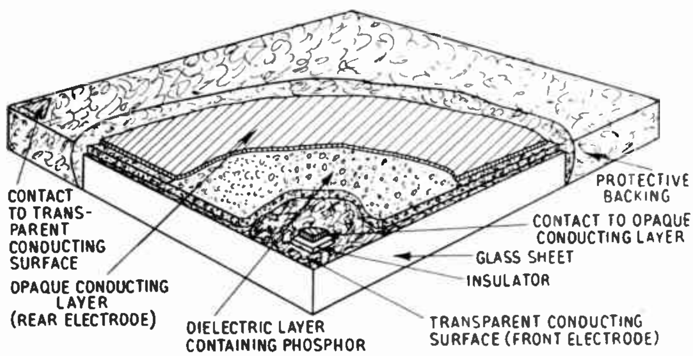

Fig. 2. Construction of an electroluminescent panel.

This layer might be caused by "excess" Cu on these surfaces.[5] Certainly, light can be stimulated at low voltages where the field strength across whole crystals is far too low to cause direct field excitation, which would require a field strength $\sim 10^7$ V/cm. Other investigators[5] have evaporated Cu on to a surface of ordinary fluorescent crystals and have claimed to observe electroluminescence. In phosphor powders the light is stimulated at small specks on the surface and not generally in the body of the crystallites (Fig. 4). We have found with single crystals[6] that the light pattern

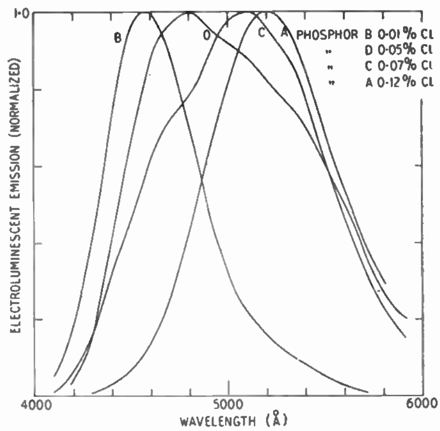

Fig. 3. Dependence of the spectral emission distribution of a ZnS:Cu,Cl electroluminescent phosphor on the Cl concentration: the same concentration of added Cu (0·3%) was used in the preparation of all four samples.

is one of streaks and spots (Fig. 5) which coincide in direction with dividing planes between regions of different degree of stacking disorder. We interpret these as directions of easy access for Cu diffusion.

The brightness B of a panel increases rapidly with increasing voltage V (Fig. 6), and this is the reason why a high dielectric strength medium is necessary in order to obtain useful brightness at high field strengths. The relationship is generally agreed to be

$$B = \omega A \exp\left(\frac{-k}{\sqrt{V}}\right)$$

where A and k are constants, nearly independent of ω if ω is not too high. In the working range of a normal electroluminescent phosphor a very

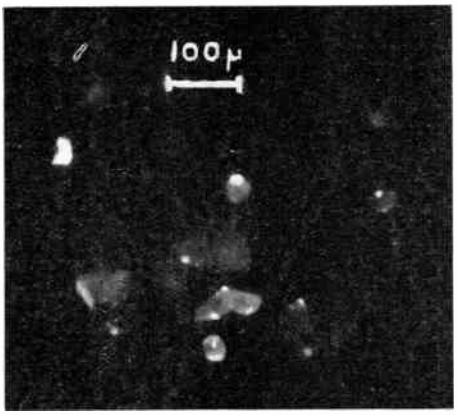

Fig. 4. Microphotograph of single particles of an electroluminescent phosphor during excitation by a 10 kc/s electric field. The stimulation of light at small specks on the surface is clearly visible.

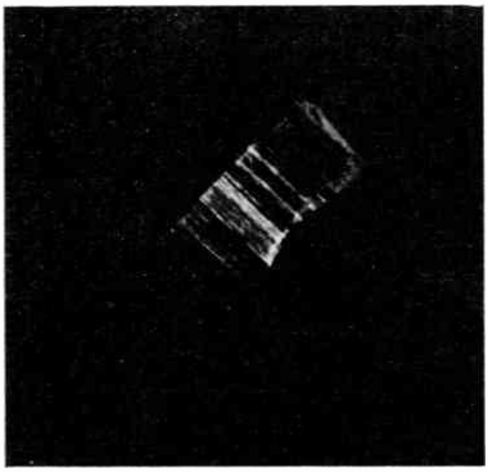

Fig. 5. The light pattern produced in a copper activated single crystal of ZnS when stimulated by a 10 kc/s electric field.

approximate formula is $B \propto V^n$ where n lies between 3 and 4. As the first formula indicates, the output luminance increases with frequency (Fig. 7). There is one complex light pulse for every half cycle of the supply voltage, and increase of frequency leads simply to more pulses per second without substantially reducing the pulse height. A saturation effect gradually sets in and this is at a much lower frequency for green phosphors than for blue phosphors. Consequently a panel made from a ZnS:Cu phosphor which has both green and blue centres, tends to become more blue with increase of frequency.

Electrically the panel behaves as an almost perfect capacitor of the order of 200 pF/cm^2 for a layer 0·001 in. thick. There are negligible losses at audio-frequencies and the capacitance is virtually independent of the light output.

2.2. *The Photoconductive Layer*

This layer is also made by the deposition of a finely divided powder—this time a photosensitive powder, usually CdS with suitable activator. For mechanical stability the powder is put down in a plastic binder, which must be transparent and insulating, such as ethyl cellulose or polystyrene. The binder must simply bond together the particles and hold them in electrical contact. If it is too concentrated so that the interstices between particles are completely filled, then the layer becomes completely insulating. Successful layers have been made by settling, spraying, and dusting techniques. Such methods require powders of small, or at least uniform size particles but the CdS material, as prepared, may contain particles

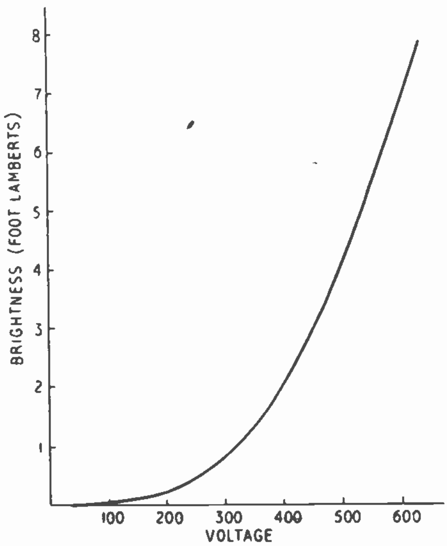

Fig. 6. Relation between the brightness of a typical green electroluminescent panel and the applied voltage (50 c/s).

from microns to "boulders" of millimetre dimensions. It is not possible to grind or mill to smaller size since this greatly decreases the photoconducting properties. Screening is used to remove the larger particles, and similar treatment of the initial material, prior to furnacing, gives a higher yield of small particles in the finished product. Another difficulty in preparing the photoconducting layer is that of critical thickness. If, as in the light-amplifier, the conductivity is to be effective through the layer (i.e. normal to its surface), then the incident light must penetrate to the bottom-most particles otherwise these are non-conducting. The particles are largely opaque to visible light, and this means that the layer must be very thin indeed and therefore the possibility of electrical breakdown is greatly increased. Furthermore, in the light-amplifier, this thin layer will have a high capacitance and this provides a parallel path for the a.c. supply to the electroluminescent layer thereby "masking" the photoconductive current, especially at low light intensities. The effect is to impair the contrast at low light levels, and the reproduction of a dark area may be impossible.

For many applications it is possible to make use of conductivity in the surface of the layer.

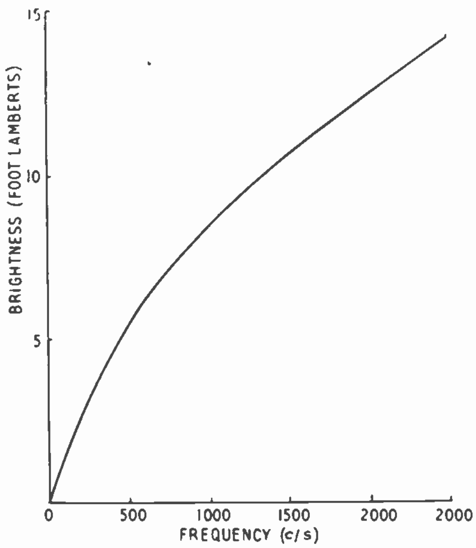

Fig. 7. Relation between brightness and frequency for a typical green electroluminescent panel.

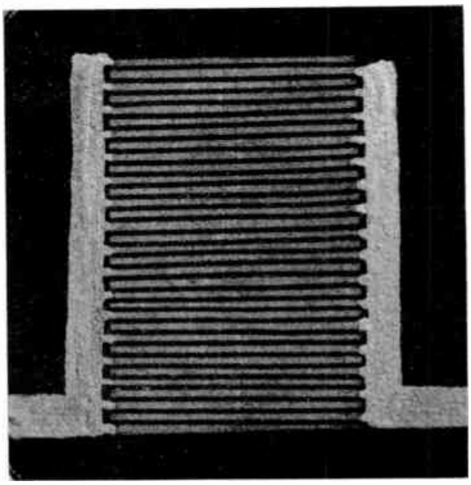

Fig. 8. Photograph of an experimental interdigital electrode structure on a "surface type" photoconducting cell.

Thus an interdigital arrangement (Fig. 8) of electrodes sprayed or evaporated via a mask on to the surface gives an efficient use of the available surface area. The sensitivity of this type of photoconducting cell is independent of the thickness of the layer. Such a cell has many applications, e.g. street light control, car headlight dipping control, industrial safety controls, etc. It has the further advantage that the substrate need not be transparent thus avoiding the use of glass. The sensitivities of some laboratory experimental types are given below—recent development has been rapid and the figures quoted may be out of date before publication but some typical specifications are as follows:

Cell 1 (high voltage, e.g. for street light control):

Area	7·5 cm^2
Voltage	250 V
I_{dark}	0·05 μA.
Photocurrent, 1 ft-candle	7·5 mA.

Cell 2 (low voltage):

Area	2·5 cm^2
Voltage	6 V
I_{dark}	0·003 μA
Photocurrent, 1 ft-candle	320 μA.
10 ft-candle	5·3 mA.

For a given intensity of illumination, the variation of photocurrent with applied voltage follows a power law

$$I = kf(B)V^n + I_{\text{dark}}$$

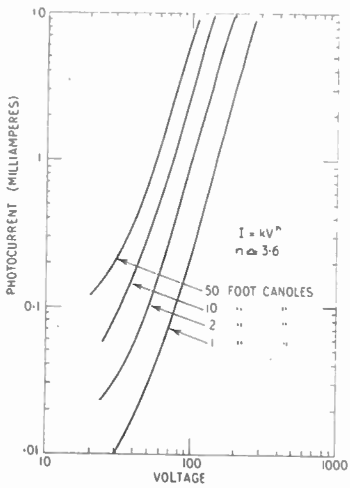

Fig. 9. Relation between photocurrent and applied voltage for a typical activated CdS powder layer.

For less sensitive powders and for voltages not too high, n is approximately 4 (Fig. 9); its value is less at high voltages. For sintered powder layers n tends to lower values and is unity for single crystals. The variation of dark current with voltage generally follows a similar power law but the dark current can be made so small as to have negligible effect on the operation of light-amplifiers. The photoconductivity decreases very slightly with the frequency of the a.c. electrical supply in the range d.c. to 20 kc/s; the capacitance of the layer is nearly independent of incident light intensity.

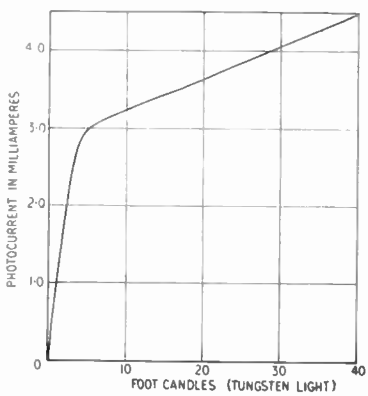

Fig. 10. Relation between the photocurrent and the intensity of illumination from a tungsten lamp.

The photocurrent is not a linear function of the incident light intensity (Fig. 10). At very low intensities the current increases super-linearly, then varies more or less linearly in an intermediate range, and tends to saturate at higher intensities.

3. The Amplifier Characteristic

When the two layers are superimposed, i.e. connected electrically in series, an approximate equivalent circuit is as shown (Fig. 11). The photoconductor resistance R varies with incident light intensity but the capacitance C of the electroluminescent cell is constant.

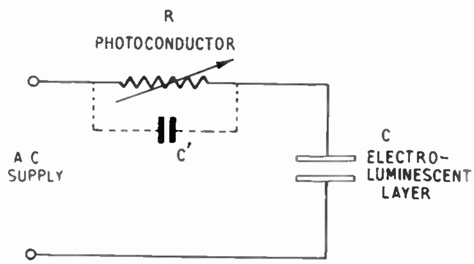

Fig. 11. The equivalent electrical circuit of a panel amplifier.

The overall output brightness (B_2) versus input brightness (B_1) characteristic has been calculated by several authors[7,8] with results in fair agreement with experiment. The calculation is of limited use because the appropriate formulae for the behaviour of the two layers vary considerably depending on the method of preparation. They also vary over the working brightness levels. Some typical measured B_2/B_1 characteristics for several values of supply frequency are shown in Fig. 12. The output brightness B_2 increases slowly at first as a result of the exponential type characteristic of the electroluminescent layer (not apparent in the diagram). At the higher input brightness levels the output brightness increases slowly due to the saturation effect of the photoconduction current. At intermediate light levels the slope of the $\log B_2/\log B_1$ curve is approximately linear and of slope γ; γ is the well known parameter used in photography etc., and is sometimes called the contrast amplification. A typical value for γ is between 4 and 6. For a given B_1 i.e. for a given value of R, the ratio B_2/B_1 tends to decrease with increase of frequency simply because of the lower reactance of the capacitance C. On the other hand the maximum output brightness B_2 is greater because the output of the electroluminescent cell increases with frequency. If the dark current were high, B_2 would be higher at low B_1 values; this effect would be greater at low frequencies where the reactance of C is large—too high a dark current would make it impossible to extinguish the cell. If the photoconductive layer is made thin in order to assist light transmission through it, a second capacitance C' appears across R—this also makes it impossible to "extinguish" the electroluminescent cell at low values of B_1. This is indeed the over-riding difficulty in the manufacture of this sandwich-type light amplifier for picture reproduction. It would appear that no panel has yet been produced in this country suitable for display in public. No great effort has been put into the development of such a panel because of the uncertainty of its eventual suitability unless the maximum brightness level can be increased. R.C.A. has produced, by a complicated process, a 12-in.

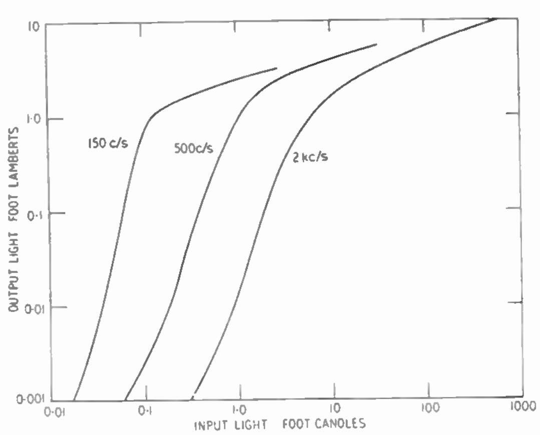

Fig. 12. The light amplifier characteristic at various frequencies of the supply voltage.

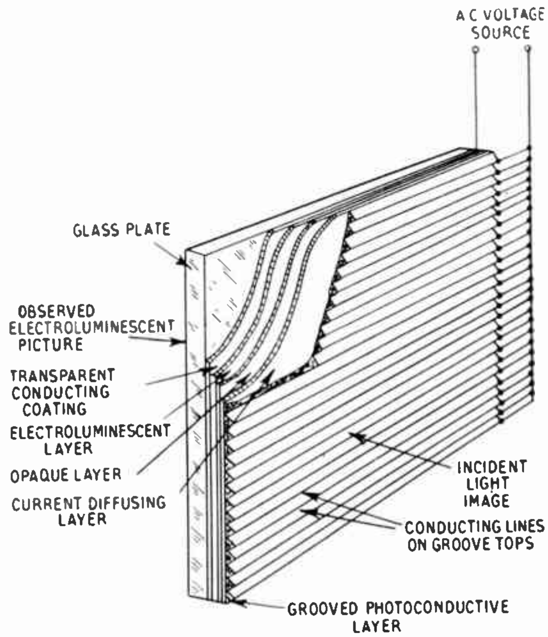

Fig. 13. Grooved-photoconductor type light amplifier.

panel[9] which is claimed to have uniform resolution comparable in quality with commercial television pictures (Fig. 13).

A phosphor layer, 0·001 in. thick, is sprayed on to a conducting glass plate in the normal manner. Next comes a thin ($<$0·001 in.) opaque layer of lamp black in Araldite, the purpose of which is explained later. Above this is spread a thick current-diffusing layer of semiconducting CdS. This layer is then machined to a thickness of 0·01 in., covered with a thick layer of photoconducting CdS powder in Araldite and the surface again machined flat to leave the last layer approximately 0·014 in. thick. The photoconducting surface is sprayed with silver paint, and then "V-shaped" grooves 0·015 in. deep are cut into it at 0·025 in. intervals. Thus the bottom of the "V" grooves just cuts into the current diffusing layer and the tops of the grooves are left as narrow lines of conducting silver which are connected in parallel as a common electrode. Light incident on the photoconductor causes a "surface type" photoconduction current to flow down the sides of the grooves. This would cause the electroluminescent layer to be activated only in narrow strips but the current diffusing layer serves to prevent this. The scheme successfully overcomes the difficulty of photoconduction through a thick layer (of small capacitance) but gives one the impression of being difficult to manufacture on a large scale.

Unless the previously mentioned opaque conducting layer is interposed between the electroluminescent and photoconducting layers the light output of the former is incident on the latter thereby providing optical feedback. This feedback might be used to increase the "gamma" of the overall system but in the reproduction of a picture it is undesirable since the half tones are lost. The feedback "ratio" will depend on the relative spectral response curves of the two parts. The sensitivity of the usual CdS:Cu,Cl photoconducting material varies with wavelength as shown in Fig. 14; the peak at ~7000 angstroms is conveniently placed for operation with tungsten light. The overlap of this curve with the green electroluminescent spectral emission distribution curve is such that the feedback ratio is of the order of 5 per cent. or less. Consequently an amplifying system with a gain of 20 times or more will become unstable; a sufficiently high incident light intensity will then cause the output light intensity to build up into a saturated condition. This condition will be self-maintained after the incident light has been removed and can only be terminated by sufficient reduction of the a.c. supply voltage. This "thyratron-like" switching action promises to be of considerable use in certain novel switching arrangements to be described later.

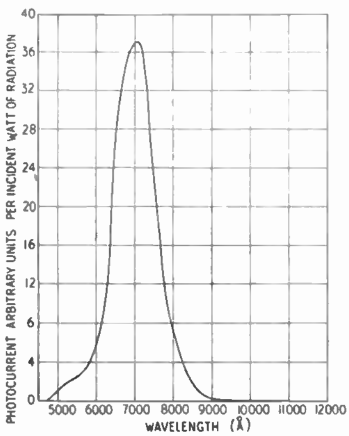

Fig. 14. The spectral variation of the sensitivity of a CdS:Cu,Cl photoconducting powder.

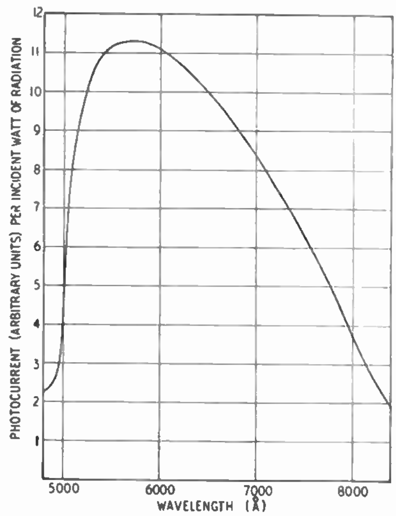

Fig. 15. The spectral variation of the sensitivity of a CdS:Ag, Br photoconducting powder.

For switching applications the maximum feedback is desirable; one method of improving the overlap of the response curves is to increase the wavelength of emission of the phosphor. This again calls for a good red electroluminescent material. The present (ZnS:Mn(Cu)) orange phosphors have a lower brightness level than the green phosphors so that little advantage is gained by their use. A photoconducting CdS:Ag, Br powder has been developed with a broad response curve having a maximum near 5500 angstroms (Fig. 15). However, this powder has not yet been made with a photosensitivity as high as that of the one previously described so that, again, little advantage has been gained. It is hoped that future work will lead to suitably matched materials which will enable these electro-optical binary switching units to be made to work from reasonably low voltage supplies, in particular from the a.c. mains directly.

Cadmium sulphide single crystals have been activated to have a photo-sensitivity with a peak in the region of the absorption edge

(5100 angstroms). Such crystals have been mounted on small strips of green electroluminescent panel, thereby forming units of very small dimensions. The first experimental types require a working voltage of about 300 V. Future developments must include the preparation of thin crystal slices; these might well be combined with thin slices of electroluminescent ZnS single crystals to form very compact, robust switching units.

4. Electro-optical Switching "Circuitry"

An electro-optical pair will be said to be "on" when in the self-maintained light emitting condition and "off" when the light output is negligible, the unit not having been "triggered" by a light "pulse." The electroluminescent brightness attains almost its final value after a few cycles of the supply voltage at audio frequencies; the decay time is often about 1 msec. For fast switching, then, a high frequency source would be required but, as explained previously, this leads to reduction in the B_2/B_1 ratio and regeneration is less; larger trigger light "pulses" and a higher value supply voltage would necessary for satisfactory action.

At the present stage of development the most significant limitation to switching time is the response time of the CdS layer. This depends on the illumination, being much faster at higher intensities. It also varies between photoconducting powders and as a general rule it is slower for more sensitive powders. Some binders also lead to a slower response; the reason for this is obscure. The photoconductive "gain" depends on the lifetime of an electron excited into the conduction band so that one would expect the more sensitive material to have slower response. However it is thought that such lifetimes are considerably shorter than the rise and decay times actually measured and that the present effects are mainly due to electron traps. It is therefore possible that faster response time (order of milliseconds) may yet be obtained with adequate sensitivity.

Although there is this difficulty of response time (the best switching time reported is 2msec using a CdS crystal) there are numerous applications where it is not a restriction and where the electro-optical switch has several advantages, such as:

(*a*) very small physical size (e.g. 1/16"–¼" square)

(*b*) minimum of subsidiary components

(*c*) low cost raw materials

(*d*) low power consumption, e.g. 1 μA r.m.s.

(*e*) the state of the circuit could be read off optically

(*f*) in slow speed devices, operation directly from a.c. mains.

Suggestions are now given for arrangements of electro-optical pairs which might be further developed to replace existing equipments with saving of space and cost.

4.1. *Reversible Binary Switching Unit (for use as scale of 2 counter, gating circuit, etc.)*

This device consists of two electro-optical pairs with electrical cross-connection as shown in Fig. 16 which also depicts the practical arrangement. In the "rest" condition one of the electroluminescent cells, say EL1, is "on," the other "off." A pulse of light on D2 causes its impedance to be reduced and this has two effects:—

(*a*) the voltage across EL1 is reduced causing its light to decrease,

(*b*) the voltage across EL2 is increased, leading to increase of light output.

As a result of (*a*) the impedance of D1 is increased and as a result of (*b*) the impedance of D2 is further decreased. The effect of the pulse is therefore to switch off EL1 and switch on EL2. Because of its symmetrical arrangement, the device is ready to be switched back into its initial condition by a pulse of light incident on D1. The state of the circuit can be read off by direct observation of EL1 and EL2.

The circuit would act as a scale-of-2 counter if it were arranged for the input pulses to be incident on D1 and D2 alternately. Fortunately, the photo-conductor which was previously non-illuminated is the more sensitive of the two because of the higher voltage across it and because the photoconducting material is more sensitive at lower illumination. This means that the input pulses can be applied to both D1 and D2 simultaneously and the circuit will be reversible, there being one output signal for every two input light pulses. For this purpose it would be advantageous to include Z; the supply voltage and the magnitude of Z are so chosen that the voltage drop across Z is too large for both EL1 and EL2 to be on at the same time but one cell can be steadily maintained in the "on" condition. This additional component would make for a more reliable switching action and switching by square light pulses would be possible. The light output of EL1 (or EL2) could then be used to trigger a following scale-of-two.

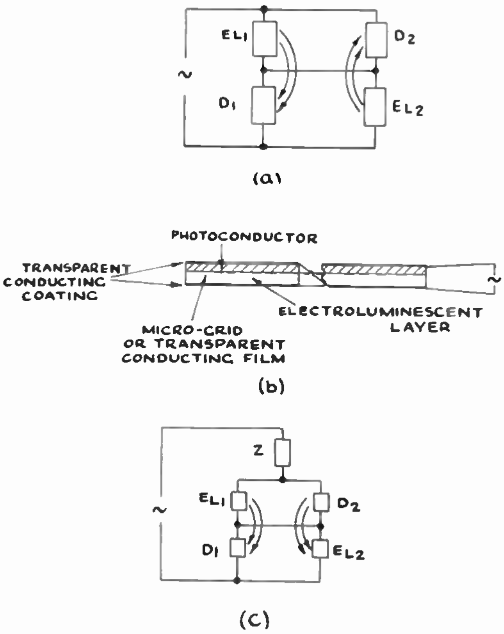

Fig. 16. Reversible binary electro-optical switching unit.

If necessary, electrical input pulses could be used by applying them to subsidiary electroluminescent cells whose output is incident on D1, D2. Electrical output pulses could be obtained by use of subsidiary photoconductors exposed to EL1, EL2, or by using the a.c. voltage appearing across EL1, EL2.

This application requires a photoconductor of high sensitivity since it is used "in shunt" across an electroluminescent cell: its dark current can be quite high, however, since it is in parallel with the capacitance C of the electroluminescent cell.

4.2. *Shift Register*

This circuit is based on a logical design involving two chains of parallel connected electro-optical pairs, there being as many pairs (n) as there are required units in the register (Figure 17). Construction of each pair is simplified since no connection is made to the junction of the photoconducting part and the electroluminescent part. The shift (clock) pulses must fulfil two functions; they must reduce the voltage supply to chain A so that all cells are extinguished, at the same time they must increase the voltage supply to chain B so that the appropriate cells may come on. Because the light output of E_{Am} in chain A is incident on D_{Bm} of chain B, the content of E_{Am} ("on" or "off") is transferred to E_{Bm} of chain B. After a suitable pulse width to allow the two chains to settle down in the new condition, the supply voltages revert to their original values, so that all cells in B are extinguished and cells in A can come on again. Because light from cell E_{Bm} in chain B was incident on D_{Am+1} in chain A, the contents of E_{Bm} (previously E_{Am}) are transferred to E_{Am+1}. Thus, all digits in the system are caused to move one place along chain A. The last digit is used to operate a further device or is lost. Incoming digits take the form of a square light pulse (to register 1) or absence of pulse (to register 0) incident on D_{A1} in chain A. These pulses are synchronized with the shifting pulses. Outputs may be taken, in parallel, from all of $E_{A1} - E_{An}$ in chain A, either optically, or electrically using the a.c. voltage across the electroluminescent cells.

4.2.1. Shift pulses

These could be supplied electrically by applying the a.c. supply voltage to the two chains via gating stages. The same input pulse would increase the supply voltage to chain B via an "open" gate and decrease the supply voltage to A via a "shut" gate, during the pulse. It might be preferable to employ optically triggered gates and for this the arrangement of Fig. 17 (*b*) could be used. D_A and D_B are photoconductors capable of passing sufficient current to maintain all n electroluminescent cells in the "on" condition. Shift pulses are in the form of positive light pulses to D_B and "negative" light pulses to D_A, since in the rest condition D_A is continuously illuminated, and

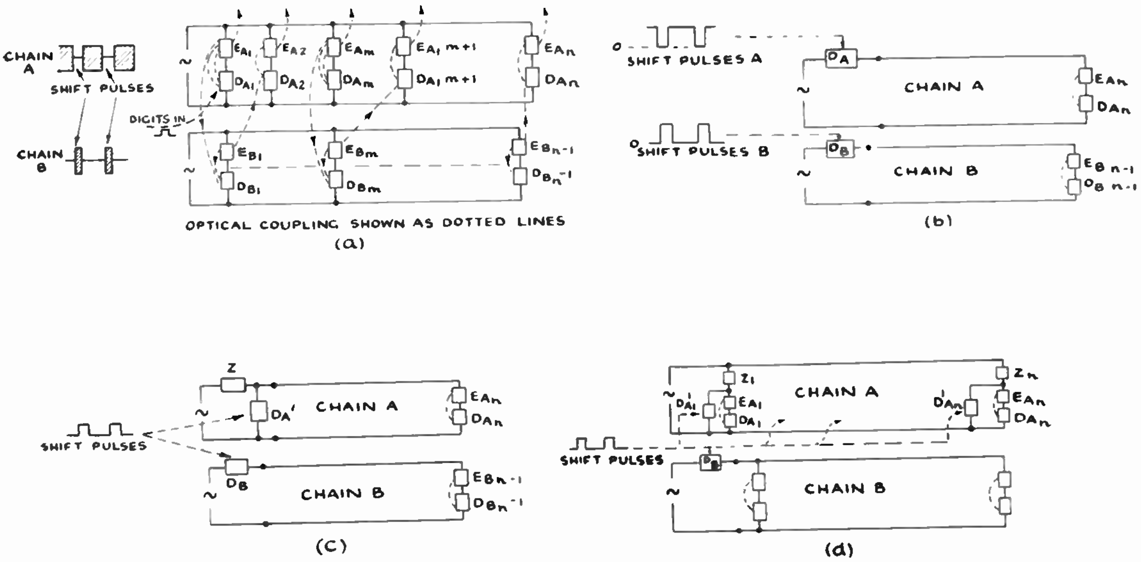

Fig. 17. Shift register using electro-optical pairs.

digits can remain stored in A. During the shift pulse, D_B is a low resistance so that digits are transferred to B. An arrangement which makes possible the use of a single train of positive light pulses is shown in Fig. 17 (*c*). A shift pulse reduces the resistance of D_B as before, but D_A' becomes a low resistance across chain A, temporarily reducing the voltage supply to E_{A1}–E_{An} so that all are extinguished. At the end of the shift pulse D_B becomes high resistance, extinguishing $E_{B1} \cdots E_{Bn}$; D_A' is also high resistance and this enables $E_{A1} \cdots E_{An}$ to come on again. In these circuits it would be advisable to arrange for D_A', D_B to have faster response times than $D_{A1} \cdots D_{An}$, $D_{B1} \cdots D_{Bn}$. D_A' must be capable of carrying a current at least several times the current of the whole chain A. Because the supply voltage to chain A now becomes dependent on the number of "on" cells, a preferable arrangement is to split up D_A' and Z into individual parts $D_A{}'_1 \cdots D'_{An}$ and $Z_1 \cdots Z_n$ as in Fig. 17 (*d*). Similarly D_B could be split up into *n* individual parts. A second method of avoiding a double shift pulse is as follows. Using the circuit of Fig. 17 (*c*), positive shift pulses are applied to D_B only, bringing on appropriate cells in chain B. All B cells are caused to illuminate D_A' so that cells in chain A are automatically extinguished. In Fig. 17 (*d*) it would simply be necessary for $E_{B1} \cdots E_{Bn-1}$ to illuminate $D_A{}'_1 \cdots D_A{}'_n$ respectively, with similar results. This last arrangement gives a convenient practical layout.

4.2.2. Reset pulse

There are several possible methods, the basic requirement being the extinction of cells in A without transfer to B. This may be effected by reduction of supply voltage to A without corresponding increase of supply voltage to B.

4.2.3. Practical layout

One layout corresponding to Fig. 17 (*a*) is the simple "in line" arrangement of Fig. 18 (*a*). A second possible layout is the echelon form of Fig. 18 (*b*). This latter arrangement requires larger area but smaller depth and the outputs are more direct.

In order that the digit shall pass in the forward direction only, it is necessary for the light from any B cell, say E_{Bm}, which falls on a *preceding* photoconductor D_{Am} to be greatly attenuated compared with the light falling on the correct photoconductor D_{Am+1}. In the arrangements of Figs. 18 (*a*) and (*b*), this requires high absorption in the D_{Bm} and E_{Am} layers, i.e. in one complete cell thickness. In all preliminary experiments, this absorption has been found to be adequate; however, an alternative arrangement conveniently avoids such difficulties. Each electro-optical pair has its electroluminescent part divided into two

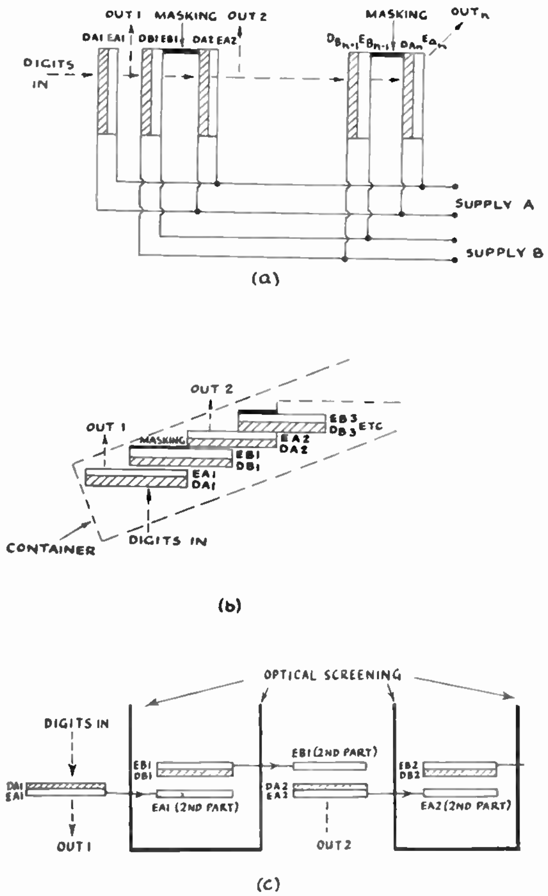

Fig. 18. Practical arrangements for an electro-optical type shift register.

sections, connected electrically in parallel. The photoconducting part, preferably a single crystal on account of its small size, is mounted directly in front of one of these electroluminescent areas. The second electroluminescent area is positioned directly opposite to the single crystal of the subsequent unit in the other chain (Fig. 18 (*c*)). With a minimum of optical screening, the passage of digits in the reverse direction is made impossible. These "double units" can be conveniently made on the same base simply by putting down the Ag backing layer as two separate areas with a narrow conducting link between them.

4.3. *Ring Counter*

It is clearly difficult to devise schemes involving time-constants and, therefore, it is suggested that a ring counter should be based on the logical design used for a shift register. A possible arrangement (Fig. 19 (*a*)) uses two chains A and B similar to the shift register but the last electro-optical pair in chain A now causes light from E_{An} to be incident on D_{Bn} of an extra pair in chain B. Also, light from E_{Bn} is incident on D_{A1} so forming a closed ring. In the rest condition only one of $E_{A1} \cdots E_{An}$, say E_{Am}, is in the "on" condition. An incoming pulse transfers this state to E_{Bm} and, on removal of the pulse, the "on" condition is transferred back to E_{Am+1}. Thus for every pulse in, the "on" condition moves along chain A by one position. After n input pulses, the "on" condition returns to E_{A1} and so on. If an output is taken from E_{An} (or any other E_A) there will be one output pulse for every n input pulses. The ring could either be triggered electrically or optically in the same manner as the shift register circuit.

An improvement in reliability results from including a suitable impedance Z in series with the supply voltage. Z should have a value such that the voltage drop across it is too great for two cells to be "on" together. The inclusion of Z lends itself readily to optical triggering. A photoconductor D_A across chain A, and a photoconductor D_B in series with chain B are all that are necessary (Fig. 19 (*b*)).

Other combinations of separate photoconductors are also possible. A particularly attractive circuit is that of Fig. 19 (*c*) in which cells of chain B are extinguished by the rise of illumination in any one of cells A. This should make for positive transfer B to A. D_B' might well be split up into individual parts $D_B'{}_1 \cdots D_B'{}_n$ located in front of $E_{A1} \cdots E_{An}$. The electro-optical ring counter has the advantage that the remaining count can be directly read off optically from $E_{A1} \cdots E_{An}$.

4.3.1. Practical layout

In the "in line" arrangement of Fig. 20 (*a*) the signal from E_{Bn} is brought back to E_{A1} by splitting the first electro-optical pair into two separate parts, D_{A1} being placed at the end of the line, opposite to E_{Bn}. An alternative (Fig. 20 (*b*)) is a closed ring which allows all electro-optical pairs to be similar. A layout using *n* separate photoconductors for D_B' is shown in Fig. 20 (*c*).

5. Future Outlook

5.1. *Electroluminescence*

The use of electroluminescence for general domestic lighting cannot be ruled out as impossible; the present rate of improvement is slow but definite. Recently there has been a demonstration of room lighting (50 ft-candles) by a complete ceiling and upper halves of walls made up of electroluminescent panels operated from a 3 kc/s, 350 V supply. The efficiency was quoted as 9 lumens/watt (cf. 16 lumens/watt for 100 W tungsten lamp). It may be that some form of high efficiency frequency convertor, supplying output power at a frequency of several kilocycles from the 50 c/s mains, will be an important factor in future development.

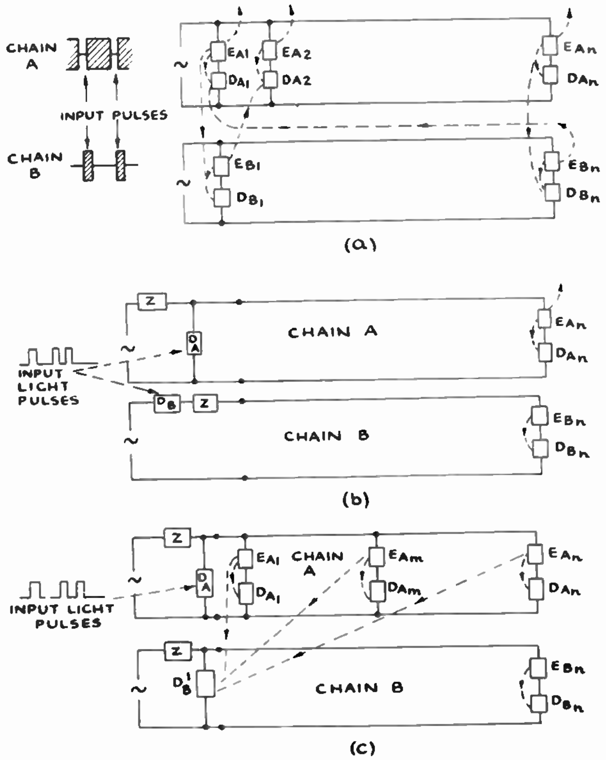

Fig. 19. A ring counter using electro-optical pairs.

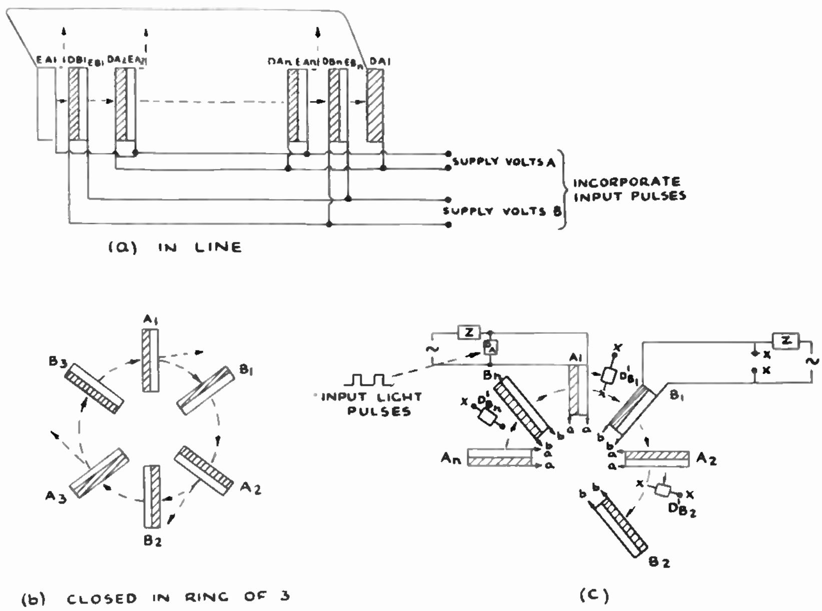

Fig. 20. Practical arrangements for an electro-optical type ring counter.

5.2. *Photoconducting Cells*

With the advent of increased automation in industry there will be many applications for these cells. Photoconduction will become an increasingly important part of electronics technique; lighting control is an obvious application, of course.

5.3. *Light Amplifiers*

The most promising use for this in the near future is for the intensification of X-ray images. Apart from possible dosage reduction it also means that the radiologist does not have to become dark adapted before using the equipment. The colour, "gamma," and response time limitations are not a disadvantage here.

There is a possibility of infra-red conversion if a suitable photoconductor is developed (CdSe, CdTe).

From the point of view of television, electroluminescent panels are not so efficient as cathode-ray tube screens and, more important, the maximum brightness is considerably less. However, the advantages of a flat screen and much less bulk will encourage further work and we may yet see successful results but there are many technical difficulties to be overcome.

5.4. *Electro-optical Switches*

At the present time the development of these switches is, in the author's opinion, a more promising field than the amplifier itself. The optical coupling and triggering enables complete electrical isolation of separate units. Combinations of series or parallel connected photoconductors and electroluminescent cells can be devised which perform similarly to many conventional networks with the advantages previously mentioned.

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A SIMPLE DIRECT READING THERMISTOR BRIDGE*

by

John Swift, B.Sc.†

SUMMARY

A simple balanced thermistor bridge is described which is direct reading and which can be used for accurate r.f. power measurements between about 10 microwatts and 1 milliwatt. The instrument requires no special components and incorporates a novel method of compensating for wide variations in ambient temperature. In addition, the bridge is suitable for field work since it may be operated from ordinary dry batteries.

1. Introduction

For the measurement of radio-frequency power the thermistor is placed in a matching device or mount designed so that ideally, all the incident r.f. power is absorbed by the thermistor bead. The thermistor mount is connected to an independent d.c. (or low-frequency) bridge circuit which enables the unknown r.f. power absorbed by the thermistor to be measured in terms of equivalent d.c. power.

Design information for various types of thermistor mount has been given elsewhere.[1,2,3] This paper is concerned only with the bridge circuit which is often more difficult to design than the mount especially when it is intended for low-level r.f. power measurements.

2. Balanced Bridge Circuits

Several distinct types of thermistor bridge circuit are in common use.[1] Of these, the balanced type has several important advantages arising from the fact that the thermistor is operated at a definite power level and therefore at a constant resistance. The balanced bridge is self-calibrating and the calibration is not dependent on the resistance–power characteristics of individual thermistors. In addition, the r.f. match to the thermistor mount is not changed when the power measurement is made.

In balanced bridges, the thermistor forms one arm of a conventional Wheatstone bridge. Initially, d.c. (or low-frequency) power is supplied to the thermistor to bring its resistance to a definite value and hence balance the bridge. When r.f. power is applied to the thermistor, balance is destroyed and the bridge is rebalanced by decreasing the d.c. power. In each balance condition the thermistor resistance and hence the total power supplied to it are the same. Thus the r.f. power is given exactly by the difference between the two d.c. power levels to the thermistor. This equivalence between r.f. power and a d.c. power difference is possible with thermistors because the heating effect and hence resistance change caused by a given power level is independent of frequency over a wide range.

In designing a balanced bridge suitable for low-power r.f. measurements, the main problem is to measure accurately the d.c. power difference which in some cases is as small as 10 microwatts. In addition, there are the problems (discussed later) of compensating the bridge for ambient temperature changes, of providing a direct reading scale and of obtaining high bridge stability.

A number of balanced thermistor bridges have been described previously[4,5,6,7] and reference to these articles will provide a good picture of the techniques and difficulties involved in measuring low-level r.f. power. All of these bridge circuits have, however, certain undesirable features such as complicated circuitry, special components, difficult operating procedure or the need for a high voltage regulated supply.

The simple thermistor bridge described below has none of these disadvantages yet the accuracy is as good as in previous instruments.

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U.D.C. No. 621.317.733.

3. A Simple Balanced Bridge

The basic circuit of the bridge is shown in Fig. 1. With no r.f. power supplied to the thermistor and with the switch S open (Fig. 1 (*a*)), R_A is adjusted until the thermistor resistor is R and the bridge is therefore balanced. When r.f. power is applied, S is closed and the bridge balance restored by adjusting R_B (Fig. 1 (*b*)). In the initial and final balance conditions the currents to the bridge are I_0 and I_1 respectively and that flowing in the shunt arm is I_B.

As the thermistor remains at a constant resistance R, the r.f. power P is given by the difference between the d.c. powers supplied to it.

Thus $$P = R(I_0^2 - I_1^2)/4 \qquad \text{.........(1)}$$

If $R_A = kR$ where k is a variable, I_0 is given by

$$I_0 = V/(R+R_A) = V/R(1+k) \qquad \text{......(2)}$$

At the final balance

$$V = (I_1 + I_B)R_A + I_1 R$$

Hence $$I_1 = (V - I_B kR)/R(1+k) \qquad \text{.........(3)}$$

Substituting (2) and (3) in equation (1)

$$P = \frac{I_B V k}{2(1+k)^2} - \frac{I_B^2 k^2 R}{4(1+k)^2} \qquad \text{.........(4)}$$

For r.f. powers below 1 milliwatt, I_B is small and the second term in equation (4) may be omitted since the resultant error (given approximately by 100 $I_B Rk/2V$) is less than one per cent. in practical bridge circuits.

Therefore $$P \approxeq \frac{I_B V k}{2(1+k)^2} \qquad \text{.........(5)}$$

Assume now that the control R_A is set to a value of R. Then $k=1$ and equation (5) becomes

$$P = I_B V/8 \qquad \text{.........(6)}$$

For a given bridge voltage V, the r.f. power P is directly proportional to I_B and the meter M in Fig. 1 may be calibrated to read r.f. power directly.

In practice however, the initial balance current I_0 depends on the ambient temperature at the time of the measurement and has a range of possible values. This arises because the d.c. power required to operate the thermistor at a definite resistance R is a function of temperature. Thus, for a given supply voltage V, R_A must be variable to provide sufficient range in I_0 and will not in general be equal to R as assumed in equation (6).

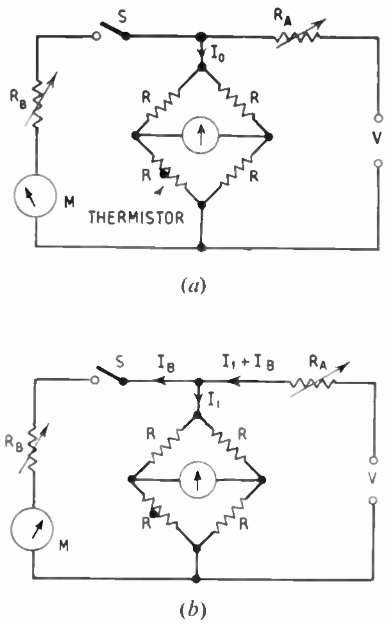

Fig. 1. Thermistor bridge; (*a*) initial balance condition; (*b*) final balance condition.

Referring to equation (5) it is seen that with $R_A = R$ (i.e. $k=1$), the function $k/(1+k)^2$ has a value of 0·250. This is the maximum value of the function but, as shown in the following Table, the maximum is very broad.

R_A	k	$k/(1+k)^2$
0·7 R	0·7	0·242
1·0 R	1·0	0·250
1·5 R	1·5	0·240

It follows therefore that R_A may be changed from 0·7 R to 1·5 R without equations (5) and (6) differing by more than 4 per cent. From equation (2) it is seen that this variation in R_A enables the initial balance current I_0 to be adjusted between the limits $V/1{\cdot}7\,R$ and $V/2{\cdot}5\,R$. In practice this current range is so large that the bridge may be balanced at widely different ambient temperatures.

The bridge voltage V is chosen so that with R_A at approximately its mean value (i.e. $R_A \cong R$), the value of I_0 obtained corresponds roughly to the ambient temperature conditions. The required bridge voltage therefore depends on the thermistor used and on R which is governed by the r.f. matching requirements of the thermistor mount. In practice V will seldom be larger than about 12 volts so that ordinary dry cells or storage batteries may be used to operate the bridge.

From equation (6) it is evident that the current I_B is directly proportional to the r.f. power P, only if V remains constant. In practice however, the voltage of the battery supply will change slowly with time. To avoid errors in the power reading, a low-resistance potentiometer is therefore placed across the battery to ensure that the correct value of V is always applied to the bridge. Occasional readjustment of this control will then compensate for slow changes in the battery voltage.

4. Sources of Error at Low R.F. Power Levels

It has been shown that the simple bridge of Fig. 1 is direct-reading and that the meter calibration is not appreciably upset by slow changes in ambient temperature or in bridge voltage. When however, the bridge is used to measure very low r.f. power levels, the short-term stability of the bridge becomes very important. If the bridge is balanced in the absence of r.f. power, the initial balance current I_0 should drift only slowly with time and be essentially constant over a time interval much longer than that required for the complete power measurement. The slow drift in I_0 is then easily corrected by occasionally readjusting the initial balance control R_A or the supply voltage potentiometer. If, however, I_0 drifts or fluctuates rapidly, the initial balance condition may be destroyed before the final one is made and a serious error can occur in the power measurement. The reduction of this effect is a major problem in all types of manually balanced bridges and is one reason for the development of the complex and expensive self-balancing bridge.[8]

Short term drift is usually caused by rapid changes in ambient temperature and in the bridge voltage V. In many cases, the fluctuations in ambient temperature are too small to be important. In field work, where the changes may be larger, error can be minimized by placing the thermistor mount in an insulated box providing a long thermal delay. Obtaining adequate stability of the bridge voltage V is a much more difficult problem since the short term stability needed at low r.f. power levels is extremely high. Storage batteries were found to have excellent voltage stability when used with this bridge. Even when the bridge was designed to give a full scale reading with a r.f. power level of only 20 microwatts the error due to drift in the initial balance condition was negligible. Similar results were obtained with miniature dry accumulators which are particularly useful if only short periods of bridge operation are required. In many cases, however, it is more convenient to use dry cells to operate the bridge. Ordinary No. 6 cells were found to have very good stability and errors due to drift were negligible for r.f. power measurements between about 1 milliwatt and 100 microwatts. At lower r.f. power levels, the drift in the initial balance is more noticeable but still not particularly serious.

One further important design factor is the correct choice of the balance indication meter. Obviously the calibration equation (6) is not valid if the two balance conditions cannot be made accurately. The bridge sensitivity, or the degree of unbalance produced by a given r.f. power level, should be high to ensure accurate balancing. In practice, however, the circuit parameters governing the bridge sensitivity are more or less fixed by other requirements of the bridge.[2] The usual procedure is to use, therefore, a balance meter which has a very low internal resistance and the highest possible current sensitivity. Reasonably robust, centre zero meters are readily available commercially, which will provide adequate sensitivity for this bridge.

5. Two Practical Bridge Designs

The above theory will now be used to design two practical thermistor bridges for the r.f. power ranges 1 milliwatt – 100 microwatts and 100 – 10 microwatts.* The thermistor mount used with these bridges consists of a short length of rigid 50 ohm coaxial line terminated

*Australian Patent Application No. 13048/55.

by two Stantel E 2361/20 thermistors in an untuned cavity.[1] The frequency range of the mount for the operating conditions given below is about 100–600 Mc/s. The bridge can easily be modified to suit other frequency ranges or to work with other available thermistor mounts. The steps in the design of the bridges are given below.

(*a*) The first point in the design is to decide upon the value of R, the thermistor operating resistance. This depends on the thermistor to be used, the r.f. mount and often on the frequency range to be covered. If the untuned cavity mount is used for the 100–600 Mc/s range, each E 2361/20 thermistor is operated at 100 ohms to give a total series resistance R of 200 ohms. From the characteristic curve of the E 2361/20, the average current required by the mount is approximately 12·5 mA. and therefore the average value of I_0 is 25·0 mA.

(*b*) Knowing R and I_0, the bridge voltage V may now be determined. With R_A at approximately its average value (i.e. $R_A = R$) and using equation (2)

$$V = I_0(R + R_A) = 2I_0R = 10{\cdot}0 \text{ volts.}$$

(*c*) The initial balance control R_A covers the working range of $0{\cdot}7\,R$ to $1{\cdot}5\,R$. As $R = 200$ ohms, the required range in R_A is 140 to 300 ohms. From equation (2) it can be seen that with $V = 10$ volts and R_A at its extreme limits, the maximum and minimum allowable values of I_0 are 29·4 mA and 20·0 mA respectively, corresponding to a very large variation in ambient temperature conditions. R_A is made up of fine and coarse potentiometer controls and a fixed current limiting resistor. It is most important that the potentiometers used here and elsewhere in the bridge should be of high quality.

(*d*) The design of the shunt arm across the bridge depends on the maximum r.f. power reading required. The r.f. power P is related to the current I_B in the shunt arm by

$$P = I_B V / 8$$

With $V = 10$ volts and P having a maximum value of 1 milliwatt, I_B is 800 microamperes. The meter M will then have a 0–800 microamperes movement and the scale may be calibrated to read r.f. power directly.

The final balance control R_B must have sufficient range to provide maximum and minimum currents of 800 microamperes and 80 microamperes in the shunt arm if the r.f. power range of the bridge is 1 milliwatt to 100 microwatts. When I_0 has its minimum possible value of 20·0 mA, the voltage across the shunt arm also has a minimum value of about 4·0 volts. The minimum value of R_B required is therefore given approximately by

$$R_B\,(\text{Min.}) \cong 4 \times 10^6/800 = 5{,}000 \text{ ohms.}$$

When I_0 has its maximum possible value of 29·4 mA, the voltage across the shunt arm has a maximum value of about 5·9 volts. Then

$$R_B\,(\text{Max.}) \cong 5{\cdot}9 \times 10^6/80 = 74{,}000 \text{ ohms.}$$

Thus the required range of R_B is roughly 5,000 to 74,000 ohms. Some overlap should be allowed, however, since the above calculation is only a first approximation and, in addition, the internal resistance of the meter M has been ignored. As was the case with R_A, R_B is composed of fine and coarse balance controls, together with a current limiting resistor.

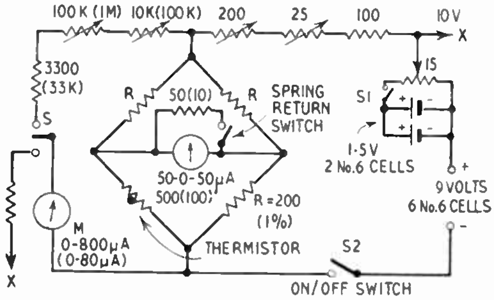

Fig. 2. Practical thermistor bridge circuit for the range 1,000 to 100 microwatts. For the range 100 to 10 microwatts, the values shown in brackets are used.

If the r.f. power range desired is 100–10 microwatts, a similar procedure is followed. In this case M is a 0–80 microampere meter and the control R_B has an approximate range of 50,000 to 740,000 ohms.

(*e*) The balance indication meter in both circuits can be a standard 50–0–50 microamperes unit with an internal resistance of

Fig. 3. Thermistor bridge.

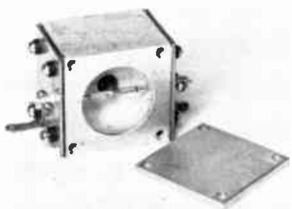

Fig. 4. Thermistor mount.

100 ohms. For the 1 milliwatt–100 microwatts range, a less sensitive meter with an internal resistance of between 500 and 2,000 ohms may be used if desired. The balance meter is protected by a shunt which is not switched out until approximate balance has been obtained.

(*f*) The power indication meter M may be used to monitor the voltage *V* applied to the bridge. This meter (in series with a suitable resistor) can be placed across the battery potentiometer when the switch S is in its initial position. A red mark placed in the centre of the scale then indicates the correct value of *V*.

The circuit for both bridges is given in Fig. 2 and photographs of a bridge and thermistor mount are shown in Figs. 3 and 4.

6. Acknowledgment

The author wishes to thank Standard Telephones and Cables Pty. Ltd. for permission to publish this paper and acknowledges the assistance given by Dr. R. Guertler of the N.S.W. University of Technology.

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# of current interest . . .

Premiums for Technical Writing on Radio

The Radio Industry Council has recently announced the award of a number of premiums of 25 guineas for articles published in the technical press during 1956. These include an article on "Particle Accelerators and Their Application," by D. R. Chick, M.Sc. (Member) (Research Laboratories, A.E.I.) and C. W. Miller, M.Sc. (Associate Member) (Research Department, Metropolitan-Vickers Electrical Co. Ltd.), published in *British Communications and Electronics*, October and November, 1956.

Mr. Chick has been a member of the Council and the Education and Examinations Committee and Mr. Miller, a past Chairman of the North Western Section, was author of two papers on the application of linear accelerators at the 1951 Convention.

American Manpower Requirements

It has recently been pointed out by the American periodical *Electronics* that during 1957 the output of the electronics industry in the United States would be at least 6½ per cent. greater than last year. Most of this increase would be accounted for by the growing demands for industrial and defence equipment; the domestic side of the industry, e.g. television receivers, etc., however, was expected to show only a very slight increase.

To help meet this expansion it is calculated that 10,000 graduates would be required during 1957; the Engineering Manpower Commission has however stated that only about 4,000 electronics engineers would graduate this year. Many universities report that all those students due to graduate this June have already received firm offers of employment.

Another report, by the National Science Foundation, estimates that developments in technology over the next five years will create a demand for 50,000 engineers each year, apart from the replacement requirements of 12,000-15,000 a year. In 1962 the total number of engineers in employment in the United States would therefore be 800,000 if all posts are filled compared with about 560,000 at present employed.

B.B.C. Permanent Television Station for the Isle of Man

The B.B.C. announces that, after consultation with the Broadcasting Committee for the Isle of Man, it intends to build the permanent Isle of Man television station at Carnane, near Douglas, on or close to the site of the temporary station which has been in operation there since December 1953.

It was planned originally to build the permanent station on the top of Snaefell Mountain from where the great geographical coverage could be achieved, but recent tests have shown that there would be reception difficulties in many parts of the Douglas area where viewers are at present getting good reception from the nearby temporary station. It is expected, however, that the permanent station now to be built at Carnane, together with the B.B.C.'s station at Divis, Northern Ireland, and the permanent transmitting station to be built at Sandale in Cumberland, will provide a strong television signal to well over 90 per cent. of the island's population.

The necessary approvals to build the station at Carnane are being sought and work on it will proceed immediately. It is hoped that it will be completed before the end of 1957.

Television in the U.S.S.R.

A recent report in the *Financial Times* stated that there are at present 13 television stations in Russia serving an audience of about 10M. By 1960, 75 stations should be working, covering an audience of around 25M—one-eighth of the entire population. This will mean laying 6,000 miles of land-lines between the major industrial towns where the new stations are to be built. At present only three stations—at Moscow, Leningrad and Kiev—are transmitting daily programmes.

In Russia a relatively large number of people share a television set and there are probably not more than 2M sets in the U.S.S.R. Their sets are generally smaller than in this country—models with 7-inch screens are still common, though production of a 25-inch model has just begun.

AN EXPERIMENTAL STUDY OF SOME FADING CHARACTERISTICS OF 10-CM WAVES IN THE SCATTER REGION*

by

D. G. Kiely, D.Sc.,† S. J. Robinson, B.A.‡ and F. C. Chesterman†

SUMMARY

This paper is primarily concerned with the short-term rapid fading of 10-cm waves in the scatter region. For a 100-mile path over the Bristol Channel measured results of the fading rate and amplitude, together with the correlation of the fading pattern of signals from the same source received by two spaced aerials are presented. The lack of correlation is illustrated by photographs of a particular type of c.r.t. presentation of the signals. Fading rates of the order of 1–5 c/s over periods of a few minutes were measured, with amplitudes in excess of 25 db.

Measured results of the amplitude of the slower mean-level fades which occur with periods of the order of an hour and greater are given from observations in the North Sea for the limited time of three weeks. In this time mean-level variations of 18 db were recorded during the winter of 1955.

1. Introduction

Since the early work by Pekeris,[1] Megaw,[2,3] and Booker and Gordon[4] on the propagation of radio waves well in the diffraction region by a mechanism which is believed to be associated with scattering from turbulence in the troposphere, many workers in the United States[5,6,7,8] have continued the study of the characteristics of these so-called scatter fields. Investigations have been concerned principally with the variations of the mean level of the field with season, geographical location, distance and general atmospheric conditions, as well as the theory of the propagation mechanism. Considerable experimental effort has been devoted to the confirmation of various aspects of the several theories[2,4,9,10] put forward to explain the mechanism of propagation which are still lacking in complete explanation of the observed phenomena.

This paper describes the results of observations of the short-term rapid fading characteristics of scatter fields at 10 cm wavelength. These were made on a 100 mile path totally over the sea in the Bristol Channel area between Milford Haven and Pendeen Lighthouse, near Penzance. The experiments were carried out in February, 1956, when weather conditions were such as to minimize super-refraction effects. If present, these would have partially or totally submerged the scatter signal by a stronger refraction signal of quite different characteristics. It is believed that the signal characteristics recorded are those of scatter signals as the prevailing weather conditions, cold air with a strong wind, were very unfavourable for other than normal atmospheric refraction. No direct proof of this was obtained but the signal correlation characteristics observed give strong confirmation that the received signal was a pure scatter signal, with very little or no refraction component.

There is little doubt that the fading characteristics are dependent on the atmospheric conditions at different seasons and at different localities. The present study does not include this variation and dependence; it is essentially a pilot experiment conducted over a few days in one locality at one season to indicate the order of magnitude of the fading amplitude and frequency and of the signal correlation with spaced receiving aerials.

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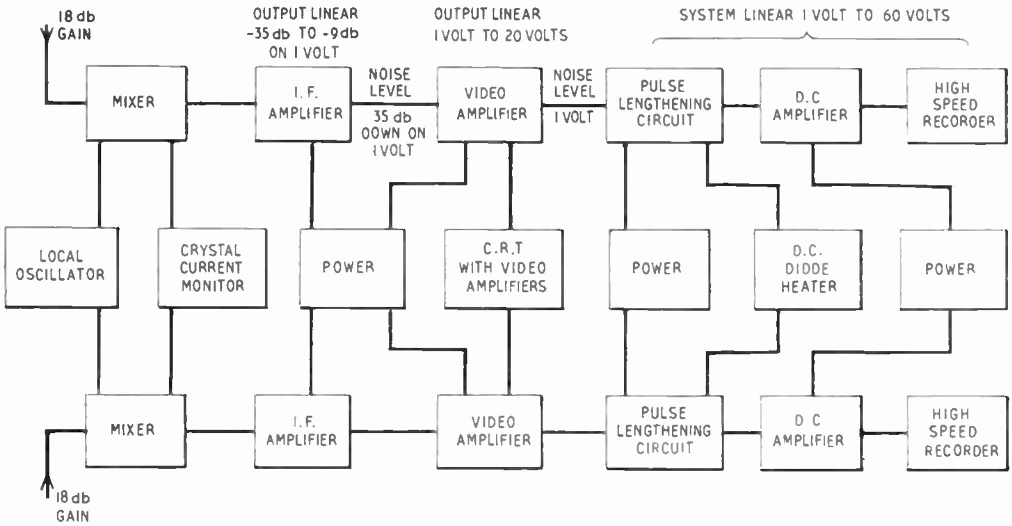


Fig. 1. Block diagram of the recording system.

A visual method of displaying the degree of signal correlation by means of a particular type of cathode-ray-tube display was employed. Simultaneous recordings of the signals from each aerial on a high-speed paper recorder enabled the correlation to be studied and the fading pattern to be observed and analysed.

A secondary object of the paper is to provide some measured results of long-term mean-level fades in British coastal waters. Information of this type has been obtained by Megaw[1] in the North Sea during the summer and by several workers for North American coastal areas. The present results relate to the North Sea area for a period of three weeks in the winter of 1955. Although this type of fading is also dependent on the atmospheric conditions at different seasons, and at different localities, the present results are included as representative of the locality of the experiment and of the season.

2. Rapid Fading Characteristics

2.1. *Object and Scope of the Measurements*

The objects of the investigation were to record the short term rapid fading characteristics of 10 cm waves from a pulse radar transmitter at H.M.S. *Harrier*, a shore Establishment near Milford Haven, by receiving the transmission with two spaced aerials associated with independent receivers. From these recordings the fading amplitude and rate were obtained, with some information on the fading rate spectrum.

The degree of correlation of signals received in the two aerials was observed for different aerial spacing distances from 2 ft to 21 ft. The aerials were spaced in directions transverse (horizontal and vertical) to the line joining transmitter and receiver, and also along this line. It was only possible to obtain signal records for two values of aerial spacing in the vertical direction due to the limitations of the site.

By arranging for the output of each receiver to provide a deflection voltage on two orthogonal plates of a cathode ray tube, the degree of correlation could be observed visually and studied in a qualitative manner. With complete correlation of the rapid fading patterns on each signal the cathode ray tube produced a steady single line trace; if the axial gain of each receiving channel was the same this line bisected the angle between the deflection directions of each plate. As the correlation deteriorated the display took the form of a sector whose width increased with aerial spacing, and within which the radial line varied in angle at the frequency of fading of some 1 to 5 c/s. A photograph of the cathode ray tube display with a time exposure of four or five seconds showed the maximum width of the sector, and gave a measure of the degree of correlation, zero correlation producing a 90 deg. sector.

2.2. *Description of Equipment*

The following is a summary of the relevant system parameters:—

Transmitter power output	500 kW peak
Transmitter pulse repetition frequency	500 p.p.s.
Transmitter pulse length	1·8 microseconds
Transmitter aerial gain	34 db (beamwidths approx. 8·5° × 1°)
Path length	100 nautical miles
Transmitter aerial height	250 ft.
Receiver aerial height	150 ft.
Receiver aerial gain	18 db
Polarization	Horizontal

Receiving Equipment:

Aerials.—A pair of identical cheese aerials were used. The vertical beamwidth was 4 deg. and the horizontal beamwidth was 90 deg. Each aerial had a gain of 18 db.

Receiver.—The twin-channel receiver, shown schematically in Fig. 1, was designed for maximum possible signal-to-noise ratio and was provided with pulse-lengthening circuits in the output stages to obtain sufficient signal power for the operation of a standard high-speed multi-channel recorder. It was necessary to reduce external interference to the absolute minimum to prevent the recording of spurious results.

The cathode-ray-tube display, with its associated video amplifiers, was incorporated in the receiver chain as shown in Fig. 1.

A balanced crystal mixer incorporating a coaxial hybrid was coupled to an i.f. amplifier. The overall receiver noise factor was 9 db.

The r.f. components and i.f. input circuit were carefully screened to prevent interference by means of induced surface currents. Due to the difficulty in adequately filtering the crystal monitor circuits these were switched out when not in use.

A narrow i.f. bandwidth discriminates against unwanted signals and contributes to good overall sensitivity. A bandwidth of 4 Mc/s was chosen to accommodate the pulse spectrum and allow for transmitter frequency instability.

The dynamic range of the recording system was determined by the i.f. amplifier. It will be shown that the pulse-lengthening circuits reduced the signal-to-noise ratio by increasing the apparent noise level. Thus, though the dynamic range of the amplifier was 26 db, the effective range of the complete recording system was 20 db. This was found to be insufficient to handle the fading amplitudes adequately, but did not materially degrade the results.

The amplifiers were provided with gain controls to allow balancing of the sensitivities of the two receiver channels.

The receiver characteristics at maximum gain may be summarized as follows:—

Signal-frequency range	2,500–4,000 Mc/s
Intermediate frequency	45 Mc/s
I.f. amplifier bandwidth	4 Mc/s
Noise factor	9 db
Sensitivity	95–100 dbm
Output noise level	35 db below 1 volt
Limit of linear output range	9 db below 1 volt

The receiver pulse-lengthening network is described and discussed in the Appendix.

2.3. *Experimental Procedure and Results*

The overall gains of the two receiver channels were balanced using an r.f. signal generator and each recorder trace was calibrated in decibels.

The radar transmitter at Milford Haven was beamed on to the receiving aerials at Pendeen. The aerial spacing was varied, in turn, along the three mutually perpendicular axes containing the direction of arrival of the signals. For each aerial spacing the high-speed paper recorder was operated for one minute, giving a record of the signals from both aerials. The same signals, displayed on the cathode-ray tube, were photographed with exposures of four to five seconds. Example of the signal recordings are shown in Fig. 2 for spacings of 2 ft to 18 ft in the (horizontal) transverse direction. It will be noted that the signal correlation for 2 ft spacing is very high, while for 3 ft spacing it is lower and for 18 ft spacing it is virtually zero. Precise evaluation of the correlation

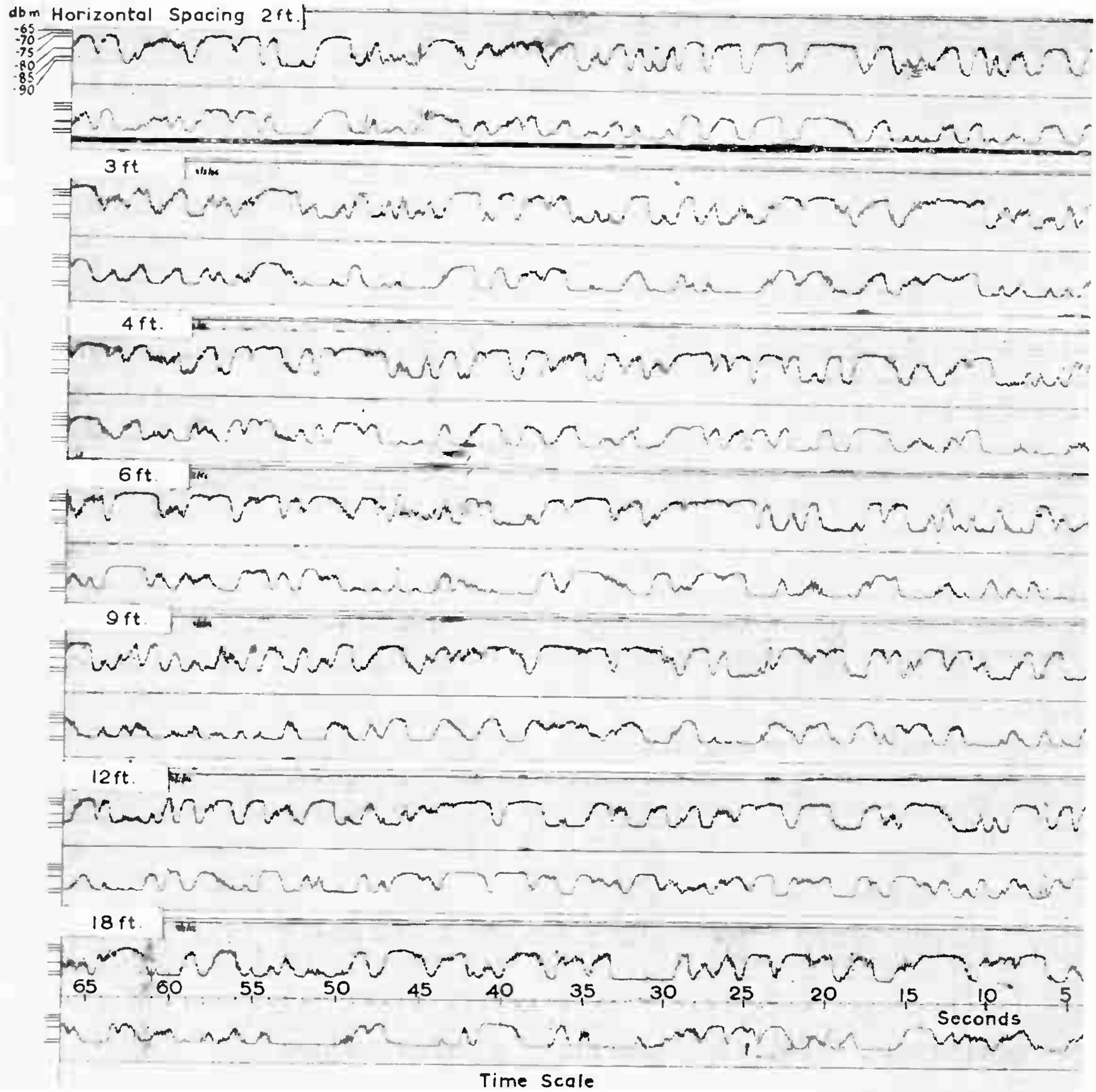
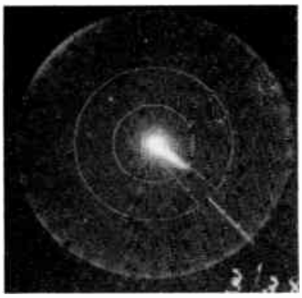


Fig. 2. Signal recordings for aerials spaced from 2 ft to 18 ft in the horizontal transverse direction.

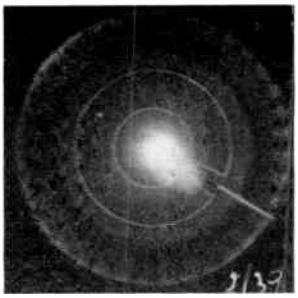
coefficients was not possible as the dynamic range of the fading was greater than that of the recorder, thus resulting in limiting at the top and bottom of its range. Fig. 3 shows photographs of the c.r.t. display corresponding to the records of Fig. 2. It is clear that for small values of aerial spacing the correlation is high, but falls off very rapidly as the spacing is increased. In this case, for a range of 100 miles and a wavelength of 10 cm, correlation is low beyond 3 ft spacing, i.e. approximately ten wavelengths.

It will also be noted that the fading rate varies from some 1 to 5 c/s and the amplitude of the fading exceeds 25 db.

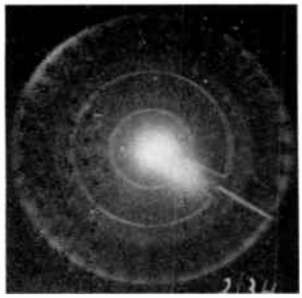
In Fig. 4 signal recordings for a range of spacings along the longitudinal (direction of arrival) axis are shown. The correlation along this axis is high and remains fairly constant for



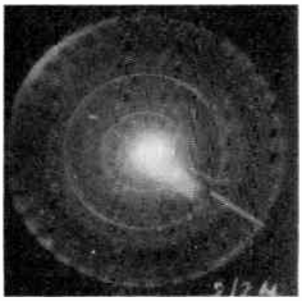
2 ft



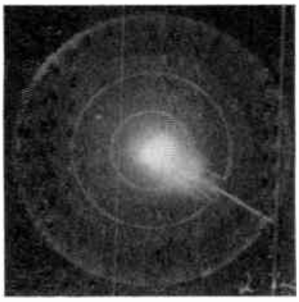
3 ft



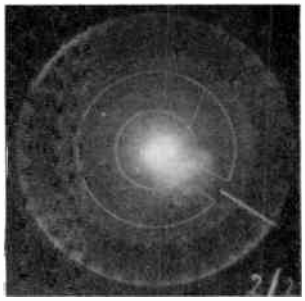
4 ft



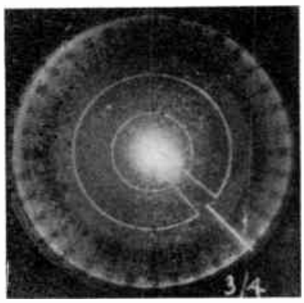
6 ft



9 ft



12 ft



18 ft

Fig. 3. Cathode-ray tube displays corresponding to Fig. 2.

the total range of aerial spacing investigated. The signal recording for longitudinal spacing of 6 ft shows a very regular and more rapid fading which was present for a few seconds. This is thought to be due to spurious reflections from an aircraft flying through the scattering centres contributing to this link. This type of fading occurred infrequently and, although it was not possible to verify the presence of aircraft in the correct vicinity, the fading is attributed to their presence from its resemblance to records taken in the U.S.A.,[12] which were associated with interference from aircraft. Fig. 5 shows c.r.t. display photographs for some of the recordings shown in Fig. 4. In Fig. 6 signal recordings for the only value of vertical spacing which could be achieved is shown. Correlation for this 12 ft spacing is very low.

The maximum components for the fading frequency are in the order of 1 cycle per second, and comparison with the white noise recordings in Fig. 7 shows that there is no evidence of a more rapid scintillation on the signal.

Some of the lack of correlation is due to a time displacement between the signal strength at the two aerials. This suggests a slow drift, possibly due to upper atmosphere winds, of the scattering centres. The time displacement involved, about 0·25 seconds at 10 feet spacing, is quite consistent with known wind speeds.

The rapid fading characteristics of 10 cm and 3 cm signals received over two paths of

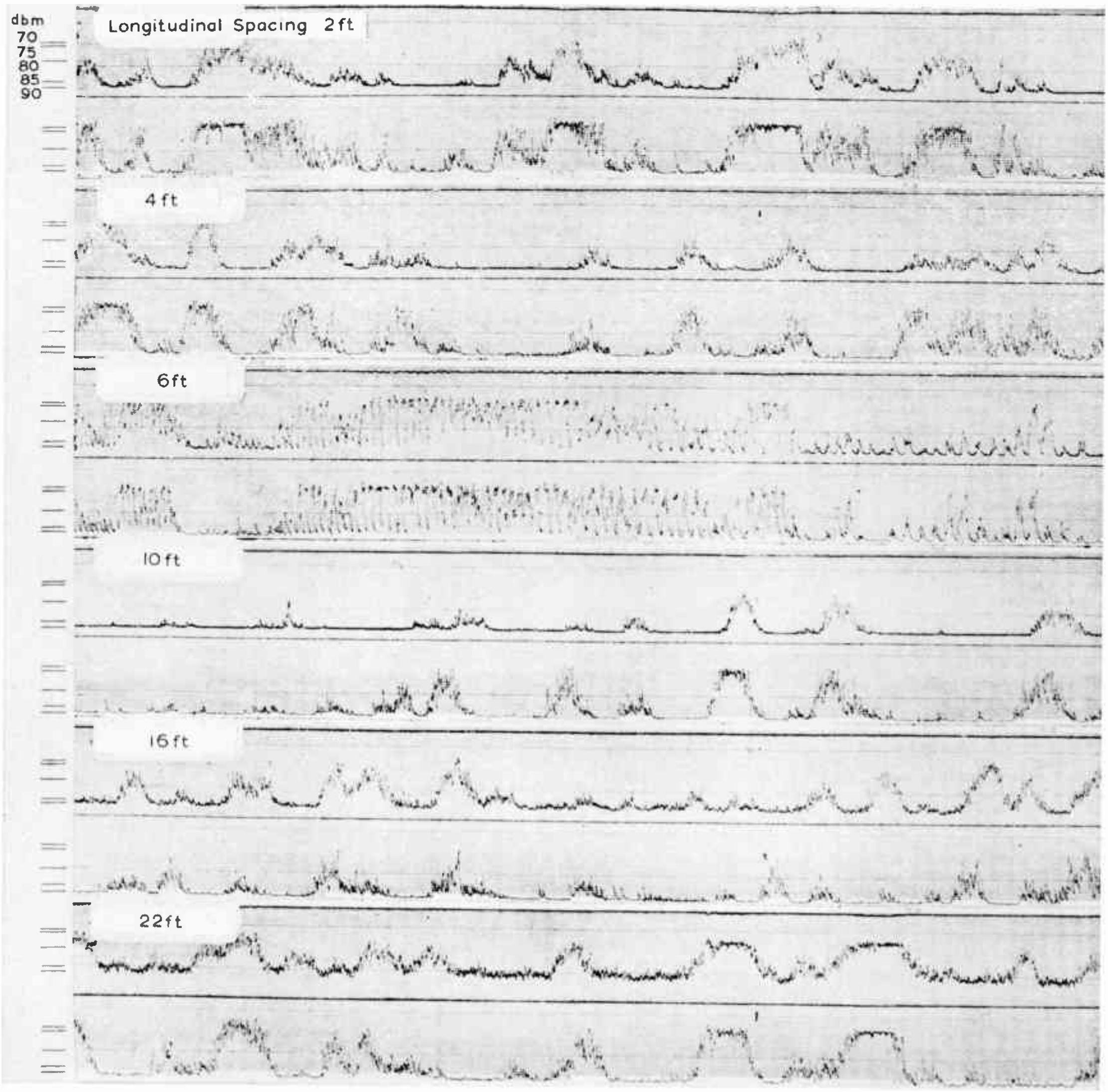
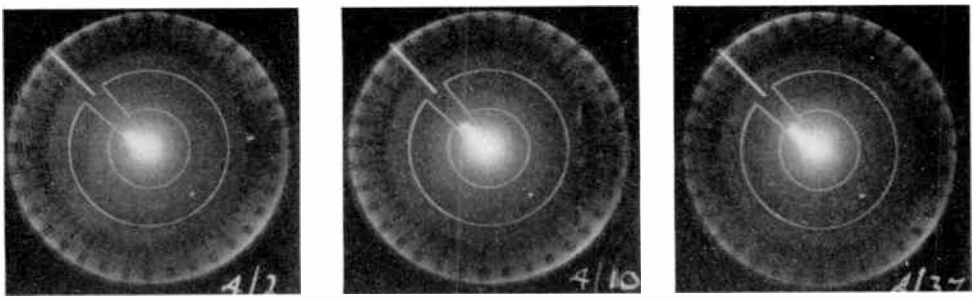


Fig. 4. Signal recordings for aerials spaced from 2 ft to 22 ft in the longitudinal direction.

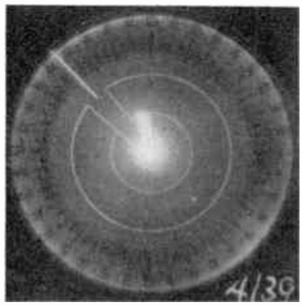
approximately 100 miles entirely over land in North America are discussed by Kurihara.[11]

The fading rates which he observed at 10 cm wavelength fall within the range 1–5 c/s and agree with those observed in the present study with an over-sea path of approximately the same length.[3] Having continuous signal records Kurihara calculated the correlation coefficients of signals received on aerials spaced horizontally and vertically. His results show a decrease of correlation with increased aerial spacing. The decrease is more rapid with the longer of the two paths and agrees qualitatively quite well with the present results. The fading depth observed by Kurihara and others[7] in the U.S.A. was, in general, less than the present results indicate. This can be attributed to the difference in local climate, and in particular to water vapour content, over the transmission paths.



2 ft 10 ft 18 ft

Fig. 5. Cathode-ray tube displays corresponding to Fig. 4.



80 dbm

3. Mean-Level Fading Characteristics

During October and November, 1955, a series of observations was carried out in the North Sea which provided some information on the mean-level fading. The observations were conducted over a period of three weeks at a time when atmospheric conditions were unfavourable for severe super-refraction, and the signal in the scatter region was virtually unaffected by refraction.

Using a 10 cm radar transmitter at Tantallon and a receiver on board H.M.S. *Fleetwood*, the ship made a number of opening runs from Tantallon. These are shown plotted together in Fig. 8. The transition from diffraction to scatter fields is clearly seen in the change of mean slope of the graphs from 2·6 db/mile to 0·20 db/mile. The variation in the mean scatter level is some 18 db for the three week period. It is reasonable to expect that the total annual variation in British coastal regions would be somewhat greater than this amount.

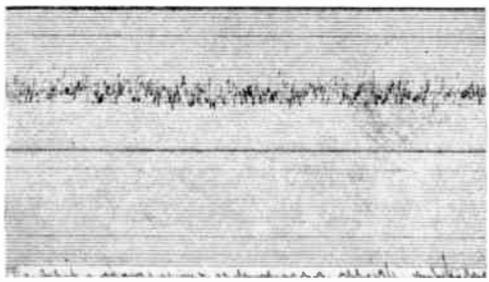


Fig. 7. Recordings of white noise.

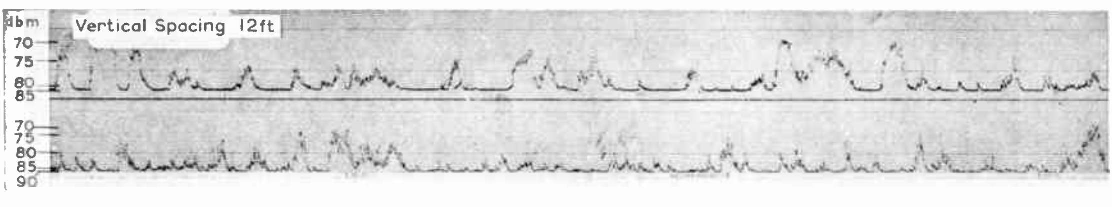


Fig. 6. Signal recordings for aerials spaced 12 ft in the vertical transverse direction.

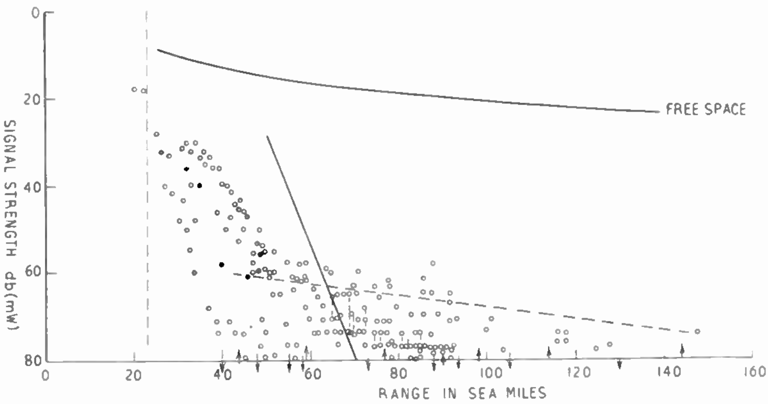


Fig. 8. Variation of signal strength with range during a three-week period in 1955.

Experimental investigation of Bullington et al[7] of the total annual variation of the mean scatter signal level at 4,000 Mc/s over a 150 mile coastal path in Newfoundland showed a variation of 14 db. This is less than the present results show for a limited period of three weeks and indicates, as would be expected, that the long term fading characteristics are likely to vary considerably with the local climate of the transmission path region. No other investigations of mean level fading carried out over extended periods in British coastal waters have yet been made. Until this has been done more complete comparison with existing measurements taken in the U.S.A. will not be possible.

4. Conclusions

Although, unlike super-refraction, the so-called atmospheric scatter phenomenon is semi-permanent in nature, it is nevertheless subject to large variations. Some figures to indicate the order of magnitude of these variations, both short-term and long-term, are given in this paper which does not, however, set out to represent the total of the phenomena. This would require observations over fixed links for a period of at least one year.

Present results indicate that, although the fading rate observations agree with those made in the U.S.A., the depth of fading, both slow and rapid, is greater than in the U.S.A. observations.

The significance of the sharp fall in signal correlation as the aerial spacing is increased in the two directions transverse to the direction of arrival is to be seen in aerial design. The transverse dimensions of broadside arrays designed to receive scatter signals will be limited by this lack of correlation. This will not necessarily be the case with end-fire arrays, as the correlation in the longitudinal direction is high and remains virtually unaffected as the aerial spacing is increased in this direction.

5. Acknowledgments

The senior author wishes to acknowledge helpful discussions with his colleague, F. A. Kitchen.

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7. Appendix: Pulse-Lengthening Network

The mean energy available from the i.f. amplifier when working within its linear range was insufficient for driving the multi-channel recorder. A pulse-lengthening circuit was therefore employed. The equivalent circuit is illustrated in Fig. 9 and the complete circuit diagram for a single channel in Fig. 10. In the design of the pulse-lengthener it was necessary to consider the detection efficiency, the ability of the circuit to follow rapid changes in pulse amplitude and its influence on the overall signal-to-noise ratio.

Let the pulse length be t_1, the spacing t_2, and let $\alpha = t_1/CR_1$ and $\beta = t_2/CR_2$, where R_1 is the charging resistance, R_2 is the discharging resistance, and C is the reservoir capacitance. Then if C is sufficiently large, the d.c. level to which an infinite train of pulses will rise is equal to V where

$$V = \frac{\text{pulse amplitude}}{1+\beta/\alpha} \quad \text{(see below).}$$

To maintain the signal-to-noise ratio, β/α must be small, i.e. $t_2/R_2 \ll t_1/R_1$. Thus a high detection efficiency implies that the amplitude following rate for a decaying signal is less than that for an increasing signal. The reservoir capacitance, C, must therefore be chosen to give a decay time-constant suitable for the following rate required from the circuit (see below).

If an estimate is to be made of the recording sensitivity of such a circuit it is important to determine the ratio of the d.c. output level, corresponding to the receiver noise, compared with the r.m.s. value of the noise voltage. A theoretical estimate of this ratio is 7·5 db.

The difference between a visually-detected tangential signal and the r.m.s. noise level is approximately 3 db. A 4·5 db loss of sensitivity is to be expected from using the recorder, as compared with a visual c.r.t. display.

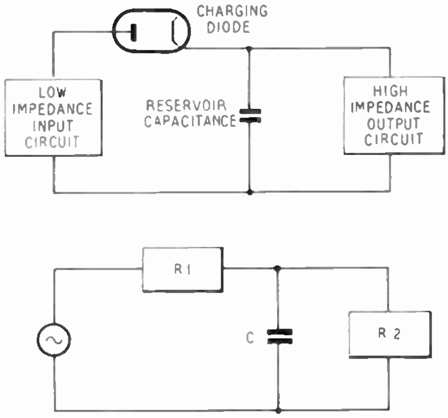

Fig. 9. Pulse lengthening circuit equivalents.
$R_1 = r_1 + r_2$ where r_1 = input circuit resistance
r_2 = diode forward resistance
$\frac{1}{R_2} = \frac{1}{s_1} \quad \frac{1}{s_2}$ where s_1 = output circuit resistance
s_2 = diode back resistance

Local interference is usually in the form of large, isolated, voltage spikes. The circuit will not respond to these if CR_1 is large. Unfortunately increasing CR_1 either reduces detection efficiency (R_1 large) or slows down the following rate (C large). It was found convenient to make C variable so that, when necessary, following ability might be sacrificed to obtain interference suppression.

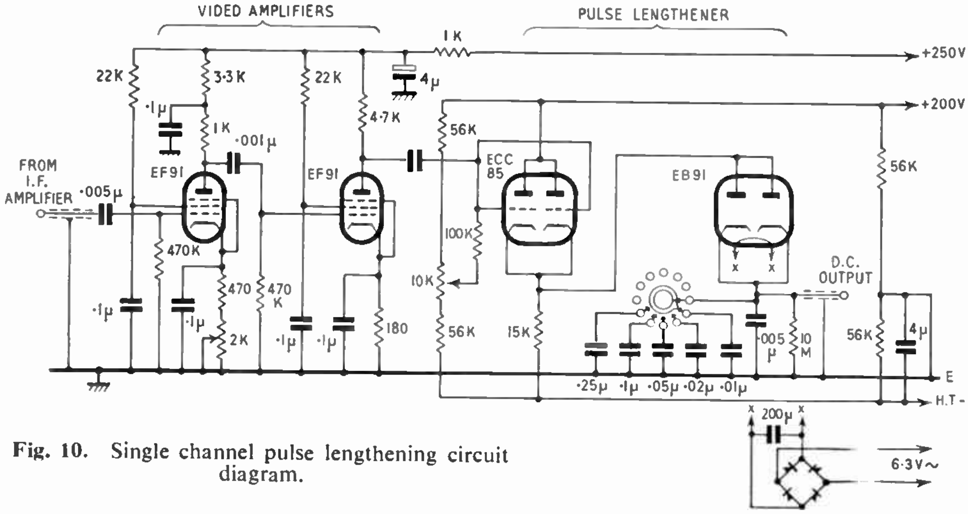

Fig. 10. Single channel pulse lengthening circuit diagram.

In practice R_1 was made 200 ohms by strapping a pair of cathode followers and a pair of diodes. Mains hum was a serious problem when R_2 was greater than 2×10^6 ohms, and even with this value it was found necessary to run the diode heaters from a separate d.c. supply. The capacitance C was 0·005 microfarads and the analysis below indicates that these values gives a detection efficiency of 0·8 and reasonable following of a 30 c/s modulation.

Figure 10 shows a single channel of the actual pulse-lengthening circuit used and Fig. 11 demonstrates the response of the circuit to 30 c/s modulation of 1 microsecond pulses with a p.r.f. of 500. The observed sensitivities indicated an improvement of 2 db for a change in p.r.f. from 500 to 3000 per second and a similar improvement for an increase in pulse length from 1 microsecond to 3 microseconds.

The output from the pulse-lengthening circuit was fed into a balanced d.c. amplifier which operated a standard pen recorder. The response time of this recorder was adequate for a 50 c/s signal and the chart speed used was 2 inches per second.

In order to make the most economic use of the recorder chart, most of the receiver noise was removed from the record by biasing the pulse-lengthening diode. This was effected by adjusting the cathode follower grid bias.

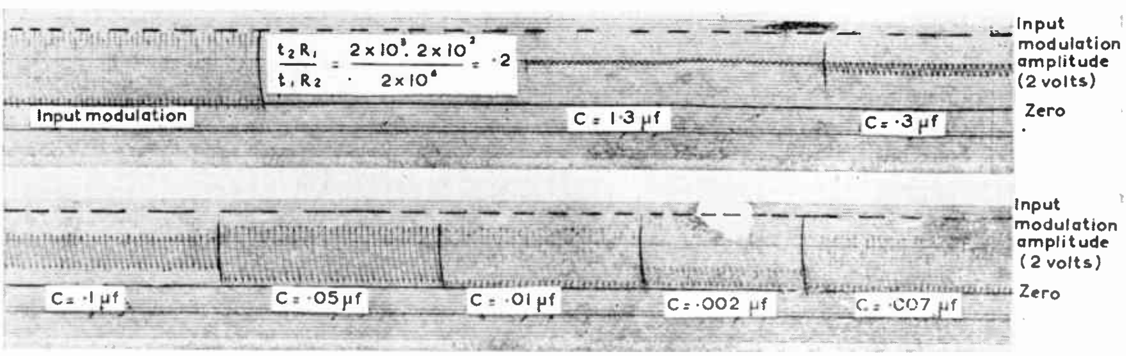

Fig. 11. Response of recording system to 1 microsecond pulses with superimposed 30 c/s amplitude modulation.

Circuit Analysis

(*a*) Efficiency:

Let t_1 = pulse length

t_2 = pulse spacing

R_1 = charging resistance

R_2 = discharging resistance

C = reservoir capacitance

ε_0 = pulse height

and let

$$\alpha = \frac{t_1}{R_1 C} \text{ and } \beta = \frac{t_2}{R_2 C}$$

After 1 pulse

$\varepsilon = \varepsilon_0 (1 - e^{-\alpha})$

After 1 pulse + 1 space

$\varepsilon = \varepsilon_0 (1 - e^{-\alpha}) e^{-\beta}$

After 2 pulses

$$\varepsilon = \varepsilon_0 (1 - e^{-\alpha}) e^{-\beta} + \varepsilon_0 [1 - (1 - e^{-\alpha}) e^{-\beta}] (1 - e^{-\alpha})$$

Whence after n pulses

$$\varepsilon = \frac{\varepsilon_0 (1 - e^{-\alpha}) \left[1 - \{ e^{-\beta} - (1 - e^{-\alpha}) e^{-\beta} \}^n \right]}{1 - \left[e^{-\beta} - (1 - e^{-\alpha}) e^{-\beta} \right]}$$

Let $n \to \infty$

$$\frac{\varepsilon}{\varepsilon_0} = \frac{(1 - e^{-\alpha})}{1 - e^{-\beta} \left[1 - (1 - e^{-\alpha}) \right]} = \frac{(1 - e^{-\alpha})}{1 - e^{-(\alpha+\beta)}} \simeq \frac{1}{1 + \frac{\beta}{\alpha}}$$

if α and $\beta \ll 1$

For efficient detection $\beta \ll \alpha$

Application:

Let $t_1 = 10^{-6}$ sec $\quad t_2 = 2 \times 10^{-3}$ sec

$R_1 = 200$ ohms $\quad R_2 = 2 \times 10^6$ ohms

$C = \cdot 005$ microsec

$$\alpha = \frac{10^{-6}}{200 \times 0{\cdot}005 \times 10^{-6}} = 1$$

$$\beta = \frac{2 \times 10^{-3}}{2 \times 10^6 \times 0{\cdot}005 \times 10^{-6}} = 1/5$$

α and β are small enough for the approximation to have some meaning.

Hence $\frac{\varepsilon}{\varepsilon_0} \cong 0{\cdot}8$ = detection efficiency.

(*b*) Following Rate:

The following rate is determined by the discharge time constant. The voltage falls to $1/e$ in $2 \times 10^6 \times 0{\cdot}005 \times 10^{-6} = 0{\cdot}01$ sec. The circuit will therefore follow modulations with periods of 0·02 sec reasonably well.

INCREASING TECHNICAL MANPOWER

Speaking on Technical Education to the Southern Regional Advisory Council for Further Education, Lord Hailsham, the British Minister of Education, commented on the output of scientists and engineers:

"This is at present under 10,000 a year, but most be stepped up to 17,000 a year in 10 years, and 20,000 a year as soon as possible. These figures, impressive as they are, may represent a maximum potential increase, but as a requirement are on the modest side.

"Further Education is a must for the multitude, and while I am Minister by far the most important task which I shall set myself will be a drive to push up the numbers of those receiving formal education of all kinds after the ages of fifteen and sixteen, both at school and after school.

"The White Paper envisaged an increase in a year's output of advanced students to fifteen thousand. This means an increase of 5,500 from the existing figure of 9,500—an increase of 20,000 in training of whom 14,000 will be taking sandwich courses. This in turn requires about 2,000 more teachers capable of taking advanced technological work. You will remember the total output of graduate technologists in any one year is only 2,300.

"Part-time advanced courses must be provided within reasonable travelling time of home and work. There must be a good spread of Higher National Certificate courses over the country, and for a considerable part of the expansion of advanced work, we shall be relying on these courses in the 150 or so area colleges. But at present many of these courses have very small numbers of students—our first objective, therefore, it seems to me, should be to strengthen these courses.

"When it comes to advanced full-time or sandwich courses which are after all of university or near university standard, we shall have to expect the students to go to the teachers rather than vice versa. That is why we take the view that the bulk of these courses should be concentrated in colleges of advanced technology and regional colleges.

"It is the greatest possible mistake to suppose that only advanced work counts. Work at the lower levels is just as important, and if of proper standards, should command as much respect.

"There also must be the greatest possible measure of freedom for students to attend courses in areas of authorities other than their own. What I am asking the Local Authorities to do is to agree to give automatic consent to "out-county" attendance of any student whom a college of advanced technology wishes to admit to an advanced course.

"In the White Paper the Government propose to feed £85M into technical education in the next five years, in order to provide buildings and equipment. The authorised programme for the next three years already amounts to over £40M and we intend to add a further £15M for 1959-60 very soon.

"This at first sight may seem a very small sum. But it must be remembered that we already have another £17M under construction. And in any case capital will represent a small part of the increase. The salary bill alone will have to be doubled by the 1960's.

"It is inevitable that technical education should be viewed primarily as an opportunity for material advance only: personal, because of the careers which it opens for the individual; national, because of the opportunities it offers of new markets and material development. I do not so regard it, nor am I one of those who because they see the overriding importance of the moral in human affairs, play down technical education as something which needs to be counterbalanced by an adequate number of arts students, trained to supply the spiritual needs of the community. Such a dichotomy in our national life would almost certainly be fatal.

"On the contrary, I feel that we should value technical advance not only for its own sake, but for the vital contribution it can make in its own right to the spiritual values of our national culture. I cannot but regard mathematics as one of the main gateways to philosophy as well as to science. I cannot regard science as other than a mental discipline in honesty and objectivity which can only enrich and enliven the culture of those who come into contact with it."

PRINCIPLES OF DESIGN OF BATTERY OPERATED FREQUENCY MODULATION RECEIVERS*

by

R. A. Lampitt (Member)† and J. P. Hannifan (Associate Member)†

First presented before the North Western Section in Manchester on 2nd February, 1956.

Read before the Institution in London on 30th January, 1957.

In the Chair: Mr. F. G. Diver, M.B.E. (Member).

SUMMARY

The special requirements of battery operated receivers generally include low running cost and the maintenance of adequate performance at reduced battery voltages. The designer, in attempting to achieve these results, is faced with several major problems not encountered in the design of the mains operated counterpart arising from the inherent features of the limited range of 1·4 V filament valves. Following a discussion of the above, principles and design of the individual stages of an a.m./f.m. receiver are dealt with, including the mixer stage, i.f. amplifier and demodulator circuit. The remainder of the receiver follows standard practice and receives only a qualitative treatment.

1. Introduction

The general principles of battery receiver design, whether for a.m. or f.m. operation, follow quite closely those for the design of similar mains-operated receivers, but there are, of course, a number of special problems involved which may be tabulated as follows:—

(1) *Running cost* must be kept as low as possible consistent with required performance.

(2) *Performance at low voltages* must be kept as high as possible in order that battery life shall not be unnecessarily limited. This has a distinct connection with the problem of running cost.

(3) *Low gain per stage*, resulting from low values of mutual conductance in the valves, makes it necessary to use a larger number of stages for a given performance compared with that required in a mains receiver. This also is reflected in running cost and initial cost.

(4) *Power output* is limited and special attention must therefore be paid to the design of the final stages of the receiver in order that the efficiency shall be as high as possible.

(5) *Valve types* are limited to quite a small range, all of which are in the 1·4-volt series, including the DC90 triode, DK96 heptode, DF96 and DF97 H.F. pentodes, DAF96 diode-pentode a.f. amplifier, DL96 output pentode and a tuning indicator type DM70. This small range somewhat limits the choice of the set designer who has therefore to take particular care in designing the associated circuits.

(6) *Circuit limitations* are also imposed due to the fact that all battery-driven valves necessarily have directly heated cathodes.

2. Important Features of Battery Receiver Design

2.1. *Running Cost*

The running cost of any electrical equipment (neglecting cost of maintenance) is directly proportional to the rate of power consumption, and so far as battery-operated receivers are concerned, rate of consumption is of prime importance. It has several side effects which control cost in different ways. For example, a given receiver chassis may be powered by very small capacity batteries, enabling a portable form of cabinet to be used because of the low weight and small dimensions. These batteries, however, may only have a useful life of say 100 hours, and therefore require frequent

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replacement. Apart from the inconvenience caused, this is a costly process as the initial cost of a battery is not in direct proportion to its size. By increasing the capacity of the batteries, however, so that their life becomes increased to 300 hours, replacement then becomes only one-third as frequent, although the initial cost of the larger batteries is only about one and a half times greater than before. Thus the running cost is halved.

A typical a.m./f.m. receiver having good sensitivity and quality of reproduction would cost approximately 1·3 pence per hour to operate from a medium-size "portable" type battery. With accommodation for a battery of twice the volume, however, the running cost would be decreased to 1·0 pence per hour.

2.1.1. Physical size of battery

The larger physical dimensions of the battery may then make it necessary to increase the size of the cabinet and may even limit its use as a "portable," and instead result in its being admitted to the class of table receivers. This could increase the cost of the cabinet and so increase slightly the initial cost of the receiver.

The object of the designer is, therefore, to see that the efficiency of the receiver is as high as possible, i.e. that the power consumption is at an absolute minimum consistent with the performance required.

2.1.2 Performance at reduced voltages

Unlike mains power, battery power is not constant, the l.t. and h.t. voltages beginning to fall from the moment the receiver is put into operation and continuing to do so until the performance of the receiver falls to a point at which it becomes necessary to replace the batteries. This is called the "end-point," and the time taken for the batteries to reach this point is known as the "useful life of the batteries."

The battery life is governed not only by the rate at which discharge takes place (the consumption of the receiver) but by the performance of the receiver at reduced voltages. This latter point is important as a more rapid fall in sensitivity, for example, would result in the receiver becoming unusable much sooner, and the batteries would then require replacement before they had reached a reasonable state of discharge.

The correct functioning of the local oscillator at reduced voltages is even more important than the falling-off in sensitivity, and in fact, it is mainly upon this that the end-point finally depends. Special attention should be paid to this part of the circuit, components should be of high quality and oscillator coil and circuit designed to ensure a continuation of oscillations to as low a level of battery voltage as possible, the principle being that so long as the oscillator continues to function, reception *is* possible under certain conditions.

The correct balancing of the capacities of the l.t. and h.t. batteries also plays an important part in the establishment of a satisfactorily low end-point. The maintenance of oscillations in the local oscillator is obviously a function of mutual conductance in the frequency changer, and at low voltage levels, this is affected by the relative values of h.t. and l.t. voltages. It is, therefore, necessary to arrange that the sizes of the batteries are so balanced that the l.t. always remains proportionately higher than the h.t. voltage. In the case of two separate batteries, if the h.t. unit is replaced before the l.t. it will then produce a situation where the voltages are in the wrong proportion, and the oscillator may cease to function after a very short period, in addition to which, the filaments of all the valves may be damaged. Thus a false end-point is created and running cost then increased.

A much more satisfactory set of conditions is produced if a properly-balanced combined l.t. and h.t. battery pack is used, and is designed to suit all other characteristics of the receiver in addition.

2.1.3. Valve performance

The valves themselves have reached quite an advanced stage of development, and the latest "96" range of 1·4-volt types draws a filament current of only 25 mA, except for the output-pentode which takes 50 mA at 1·4 V or 25 mA at 2·8 V. This is half the consumption of the valves in the previous range, and it is because of this that set-makers are now able to produce comparatively large multi-valve battery receivers which may be operated within a reasonable cost.

As a result of the low filament power, the maximum permissible cathode current is low, and therefore figures of mutual conductance are of the order of 750 microamperes per volt compared with 6 mA per volt for mains-operated types. This naturally results in the necessity for more stages in a battery receiver for a given performance compared with those required in a mains receiver. Since the addition of another stage has the two-fold effect of increasing initial and running cost, it is important that each stage is operated under optimum conditions, and that the receiver is designed to suit a particular application and no more.

One particular item of performance in which the battery-operated receiver cannot, at present, hope to compete with its mains-operated counterpart, is the power output. The normal output available from the average mains driven output pentode is between about 3 and 4 watts. Compared with this, the output from a single 1·4-volt battery pentode seems very puny at 150 milliwatts. However, provided that sufficient attention is paid to the efficiency both in the circuit and in the loudspeaker, quite a volume of reasonable quality sound can be produced—certainly enough for comfortable listening in an average size living-room.

Where higher quality of reproduction is required, such as for a good class table model receiver, and more particularly if frequency modulation is included, two pentodes (type DL96) may be used in class AB1 push-pull. Under these conditions, 400 milliwatts may be obtained, and if a well-designed phase splitter is used for the penultimate stage, a very good frequency response is also possible.

In addition to the fact that the limited range of valves puts the designer at a disadvantage when dealing with special types of receivers, other restrictions are imposed due to the very nature of the valves themselves.

For operation at the required current levels, directly heated cathodes are necessary, and it is this fact which is mainly responsible for the circuit problems which arise. For example, some of the most popular circuits are made impracticable by virtue of the fact that the cathode is automatically "tied" to earth through the filament battery. These include such circuits as cathode-coupled phase splitters, delayed a.g.c. circuits and, most important in the case of f.m. reception, grounded-grid and cascode r.f. amplifiers.

Because of the importance of reducing oscillator radiation from the aerial terminals of an f.m. receiver, a signal-frequency amplifier is desirable, and where mains operated receivers are concerned, this may be achieved conveniently by the use of a grounded-grid triode circuit feeding into the neutral point in the oscillator grid coil. For the combined amplifier and mixer, twin triodes have been adopted and may be built into several varieties of circuit at very little inconvenience.

For battery operation, the theoretical advantages to be gained by the use of an r.f. amplifier should be weighed against entirely different considerations, and this invariably leads to its omission from the receiver altogether.

2.1.4. V.H.F. techniques

The system of frequency modulation necessarily involves the adoption of v.h.f. techniques. The design of the "front end" of the receiver requires special attention, and in this connection, chassis layout and screening are particularly important in order to reduce oscillator pulling and radiation, in addition to which, very careful circuit design is necessary. As radiation is much more pronounced at v.h.f., correct balancing or neutralizing of the signal circuit is also important, so that during manufacture and servicing, adjustments must be made with considerable care.

The design and adjustment of the intermediate-frequency amplifier likewise calls for special consideration as bandwidth and a linear response within the pass-band, play an important part in the preservation of good quality reproduction in a frequency modulation receiver.

The detector also is different and operates according to a totally different principle from that of a detector for amplitude modulated signals. It must be capable not only of demodulating the carrier but also of rejecting any unwanted a.m. which may also be present. It is a more complicated circuit and requires very much more care in design and adjustment.

In the earlier forms of f.m. receiver, amplitude modulation rejection was achieved by the addition of a special "limiter" stage operating in conjunction with a Foster-Seeley phase discriminator-type of demodulator. With this combination, amplitude modulation rejection was almost perfect above the "threshold" level, but because of the extra cost and the low gain of the limiter stage, efforts were concentrated on the design of a self limiting demodulator which would result in a more economical use of components. A modified form of Foster-Seeley phase discriminator called a ratio detector, was developed and has since proved to be very popular among set designers generally. The degree of amplitude modulation rejection at high signal levels is almost as high as with the separate limiter, but it has the added advantage that limiting continues to levels far below those at which the previous circuit ceased to function.

Because of the 50-microseconds pre-emphasis imposed at the f.m. transmitter, a de-emphasis network is required in the audio frequency section of the receiver, and if full use is to be made of the superior quality of reproduction, made possible by the f.m. system, special attention should be paid to this section of the receiver, but other than this, no further comment is necessary here.

3. Details of Design

With regard to battery operated f.m. receivers, certain points may be dealt with in some detail.

In view of the circuit complications involved in the use of an r.f. amplifier together with what has been written regarding battery drain, etc., the receiver "front end" reduces to a self oscillating mixer stage, a circuit of which appears in Fig. 1.

3.1. *The Mixer Stage*

This is the only type of circuit which is practical at the present time.[10] A triode is chosen for its low noise properties, its simplicity and high conversion gain. This particular valve is a DC90[11] and was developed for this purpose and to supplement the 91 series of 1·4 volt valves which incorporated 50 mA filaments.

The circuit consists of a simple parallel-fed anode-tuned oscillator consisting of L3 and L4 coupled to which (via C4) is the aerial transformer L1 and L2. The point at which

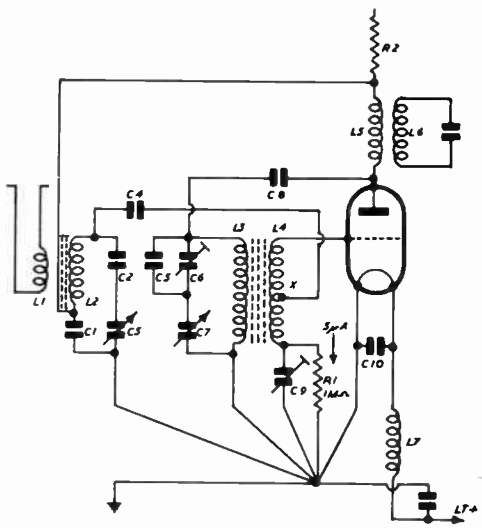

Fig. 1. Self oscillating mixer.

C4 is tapped into L4 is very important as it is upon this, together with the value of C9 that the oscillator radiation mainly depends. The two halves of L4, the grid–filament capacitance of the valve and C9 form a bridge across which is developed the oscillator voltage E_{osc} being fed to the grid. Fig. 2 shows this more clearly and shows that by correct positioning of the tap X and the adjustment to the value of C9, the bridge may be balanced so that the point X appears as a point of zero oscillator potential with respect to earth.

This is a convenient point at which to inject the signal voltage E_s so that theoretically no

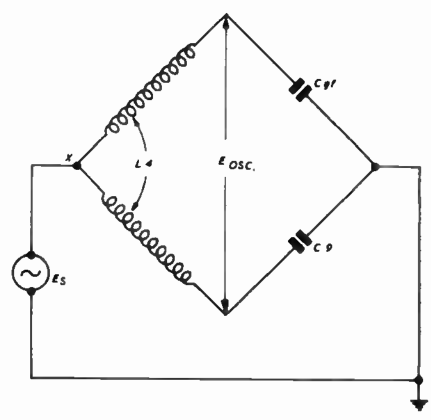

Fig. 2. Equivalent circuit of mixer grid impedance.

oscillator voltage should appear across the aerial circuit L1, L2. In practice, however, this ideal situation is never quite achieved, but nevertheless, oscillator feedback is reduced very considerably. C10 and L7 decouple the filament and prevent radiation and interstage coupling via the l.t. battery circuit.

The anode coil L5 is tuned by the capacitance appearing between anode and filament by virtue of C8 and the oscillator circuit associated with it.

Since the conversion conductance of the DC90 is smaller than its mains counterpart, it is important that the operating conditions of the valve should be selected carefully so that the conversion gain obtained is as high as possible. In order to obtain a high gain, it is essential that the impedance of the anode load be kept high compared with that of the valve. When viewed from the secondary (L6), R_a appears in parallel with L5 and as the DC90 is a triode, heavy damping of L5 will occur, in addition to which i.f. negative feedback will occur through the anode–grid interelectrode capacitance C_{ga}. Thus gain is limited.

In Fig. 1 a resistor R2 is shown connected in series with L5, and the junction is connected to earth via C1 which appears inside the r.f. tuned circuit L2, C2, C3. This forms a positive feedback loop at intermediate frequency, by means of which negative feedback through C_{ga} is neutralized, damping reduced and conversion gain increased. The principle involved is made clear by reference to Fig. 3. The impedance of the tuned circuit around L3 is small at intermediate frequency so that C8 appears effectively between anode and filament forming with C1, a voltage divider across the coil L5. C4 appears between R2 and the grid by virtue of the fact that the aerial tuned circuit also offers a low impedance, resulting in the formation of a capacitance bridge, which by the correct choice of value for C4, may be balanced so that the grid voltage relative to the filament becomes zero. Negative feedback can thus be completely neutralized and, in fact, very slightly over compensated for so that anode damping of L5 is also virtually eliminated, resulting in a considerably increased gain. By this means, the effective value of R_a may be increased to approximately 300,000 ohms.

The value of C1 is critical and should be of close tolerance in order that consistent results may be obtained in production. It should also be of high quality since it forms part of the tuning capacitance for L2, but as its value is comparatively high, it does not limit the effective overall coverage of the tuning capacitance.

The total amplification of the circuit from the aerial terminals to the control grid of the first i.f. amplifier, consists of the aerial transfer gain multiplied by the conversion gain of the mixer. With a dynamic impedance of 2,000 ohms for the aerial tuned circuit and an aerial impedance of 300 ohms, the aerial transfer gain is $\sqrt{(2000/300)} = 2{\cdot}6$, which can be more than doubled by matching into a 60-ohm aerial instead.

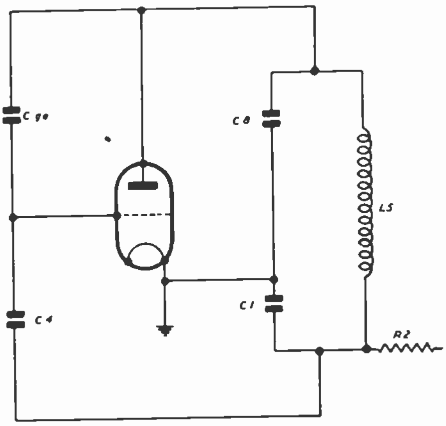

Fig. 3. Equivalent i.f. feedback network around the oscillator mixer.

An average i.f. transformer operating at 10·7 Mc/s has a transfer impedance of approximately 15,000 ohms, so that with a conversion conductance of 300 microamperes per volt, the mixer gain is approximately 4·5, making an overall amplification of 26 times. This value can be obtained in practice quite easily, and can, in fact, be improved by using a well designed i.f. transformer with a transfer impedance of 22,000 ohms, in which case the gain would be 38.

Consideration may be given to the position and value of the grid leak. By connecting it in parallel with C9 and raising its value, input damping is reduced and aerial gain is increased by virtue of the fact that a higher transformer ratio is obtained. In addition, its value is important from a noise point of view. The equivalent noise resistance of a self oscillating triode decreases with increasing values of conversion conductance, and increases with grid current so that it is important that the latter should be kept low either by a reduction of oscillator drive or by increasing the value of grid leak. A reduction of oscillator drive would limit oscillator performance at reduced battery voltage, and in turn would limit battery life, in addition to which, conversion conductance would be reduced. In this case, conversion gain would fall and noise would tend to increase. The obvious solution, therefore, is to raise the grid leak within the limit of stability to as high a value as possible, and in Fig. 1 this is shown as 1 megohm.

An increase in the aerial transformer ratio resulting from a reduction of input damping on the grid of the mixer, together with the choice of a low aerial impedance of 60 ohms further reduces the local oscillator voltage appearing across the aerial terminals.

Although oscillator radiation is comparatively high, due to there being no pre-amplifier stage, much can be done to lessen the effects by a wise choice of oscillator frequency.[12] (See Fig. 4.)

If the oscillator is designed to operate at a frequency above that of the incoming signal, coverage with an i.f. of 10·7 Mc/s will be 98·2–110·7 Mc/s, the second harmonic of which will cover almost all of the upper half of band III. Television channels 10, 11, 12 and 13 are therefore likely to suffer interference if the f.m. receiver is tuned to any one of about 20 out of the 30 proposed f.m. stations.

With a low oscillator frequency, the fundamental will lie in the band 76·8–89·3 Mc/s which will be quite clear of band I, but will, as in the previous case, just overlap band II. The second harmonic, however, between 153·6–178·6 Mc/s only falls on the lower edge of band III, but as there are no f.m. broadcast stations above 95 Mc/s, any interference caused at all would only be due to a momentary swing of the f.m. tuning control to the extremity of its traverse.

As there is less likelihood of interfering signals entering the receiver, the problem of radiation appears more important, and so far as this is concerned, a low oscillator frequency seems favourable.

So far as it affects design, there are also important points in its favour, namely,

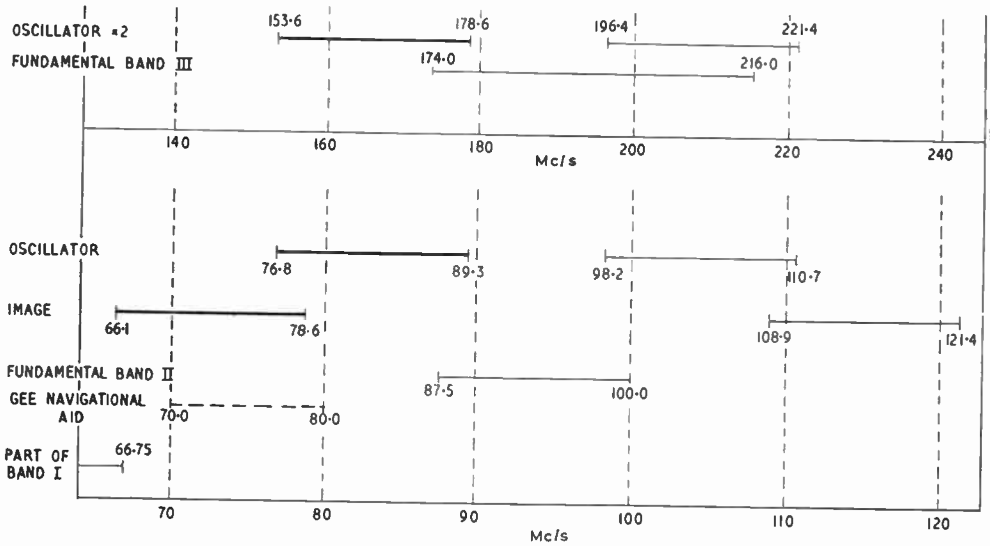

Fig. 4. Frequency spectrum chart.

(*a*) oscillator drift is a little easier to control,

(*b*) image ratio may, theoretically, be better,

(*c*) with a given value of stray capacitance, it becomes necessary to use larger values of inductance in the oscillator tuned circuit. (Where battery receivers are concerned, this is of considerable importance, because by this means, it is possible to increase oscillator grid drive and so maintain oscillations down to lower values of reduced battery voltage. This, of course, increases the effective battery life, and together with the other points raised, provides an argument in favour of the lower frequency.)

Results in practice agree very well with calculations, and a table model receiver has been designed using a triode-connected pentode type DF97 as a self oscillating mixer drawing 25 mA of filament current instead of the 50 mA three electrode valve type DC90. With an intermediate frequency of 10·7 Mc/s and an oscillator operating at the lower of the two possible frequencies, stability proves to be good, and the oscillator has a cut-off at or below 50V h.t. and 0·95V l.t.

3.2. *The I.F. Amplifier*[1,3,9]

In a receiver of the type under discussion, that is, one having a low gain at radio-frequency, the i.f. amplifier provides a large proportion of the total amplification and therefore has the main control over the receiver sensitivity.

From a knowledge of the overall performance required, the necessary gain of the amplifier may be determined readily. For the type of a.f. amplifier suitable for a typical high quality table model receiver, it was found that a signal of 160 mV at intermediate frequency deviated $\pm 22 \cdot 5$ kc/s at 1,000 c/s was required at the grid of the ratio detector driver stage in order to produce a power output of 50 mW in the loudspeaker. Under these conditions amplitude limiting was sufficient to produce a usable signal/noise ratio of the order of 35 db. Taking the measured gain of a typical self-oscillating mixer stage as 38, and assuming that 10–12 microvolts at signal frequency is required at the aerial terminals to produce 50 milliwatts output, an amplification of $(160{,}000/10 \times 38) = 420$ is required at intermediate frequency.

At a frequency of 10·7 Mc/s, this is approximately the maximum gain practicable from a two-stage amplifier using valves of the type available, giving a gain per stage of 20·5. Where higher sensitivities are necessary, such as for a portable receiver which is required to operate from a telescopic dipole almost at ground level, a third i.f. stage would be essential, in which case, further consideration would have to be given to battery consumption, etc.

The main problem to be solved in the general design of the i.f. amplifier is to obtain the highest possible overall gain with a frequency response having a level pass-band and rapidly increasing attenuation beyond it. The level part of the response is found to be a more stringent requirement than freedom from instability, due to the following considerations.

In any tuned amplifier there are the following sources of feedback: coupling between input and output circuits; common impedances in h.t. and l.t. voltage supplies; and grid to anode interelectrode capacitance.

Suitable shielding and decoupling and careful circuit layout take care of the first two of these, which brings one to the basic fact that the limit to maximum overall amplification is set by feedback through the anode–grid inter-electrode capacitance.

It can be shown that this feedback is negative when the anode circuit is capacitive, and positive when it is inductive, so that a valve having a parallel tuned circuit as its anode load will produce a distorted frequency response curve within the pass-band, with increased amplification below carrier frequency (as the load becomes inductive) if the feedback is of sufficiently large proportions.

Energy transfer from anode to grid may be looked upon as having the same effect as the presence of resistance across the grid circuit. The input resistance of the valve must be limited to a value much greater than the dynamic resistance of the grid tuned circuit, as obviously, if it approaches that value, instability will result if the input resistance is negative. Some fixed relationship between the magnitudes of the two resistances must be decided upon, and in practice this may be taken as being greater than 5 to 1.

The minimum values of the input resistance, which is varying throughout the pass-band, are found to occur at the 70·7 per cent. response points and are

$$R_g(\text{min}) = \pm \frac{2}{\omega C_{ga} A_0}$$

where A_0 = amplification at resonance for a single tuned load and is equal to E_0/E_g (Fig. 5).

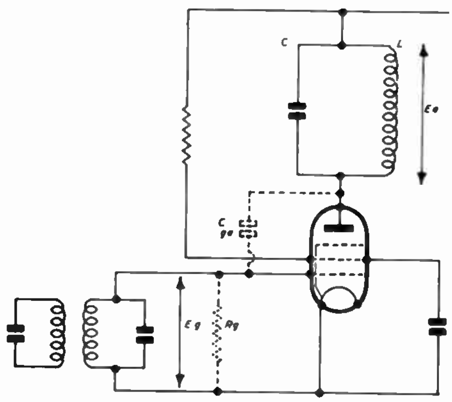

Fig. 5. Grid-anode capacitance feedback in i.f. amplifier.

For a pentode having a large internal resistance, the amplification A_0 may be written

$$g_m R_D$$

where R_D = dynamic resistance of the load.

When a second and identical tuned circuit is critically coupled to the anode coil such that $Q_p = Q_s$, the load resistance at resonance is then halved, so that the equation becomes

$$R_g(\text{min}) = \pm \frac{4}{\omega C_{ga}\, g_m\, R_D} \qquad \ldots\ldots\ldots(1)$$

In order to determine the value for R_D, it is first necessary to consider a further factor, a factor governing the shape of the frequency response curve.

From considerations of the system of modulation, the bandwidth required falls in the region of 200 kc/s at −3 db, and is given by the formula

$$\delta = K_c f_0 \sqrt{(2)} \left\{ \left(\frac{1}{m^2}\right)^{1/n} - 1 \right\}^{\frac{1}{4}} \qquad \ldots\ldots\ldots(2)$$

where

$K_c = 1/Q$ = critical coefficient of coupling

f_0 = resonant frequency

n = number of band filters (assumed identical)

$$m = \frac{\text{actual output voltage at } \tfrac{1}{2}\delta \text{ cycles off resonance}}{\text{output voltage at resonance}}$$

Transposing equation (2) and writing

$f_0 = 10{\cdot}7$ Mc/s

$\delta = 200$ kc/s

$n = 3$

$1/m = 1{\cdot}413$ (3 db)

then $Q = 54$

With the object of obtaining as high a L/C ratio as possible for the tuned circuit, the value of C must be kept as low as possible having regard to stray capacitances also present in the circuit. With the valve type in question (DF96) this is usually in the region of 10 pF, and in order to keep drift within reasonable limits, this should represent not more than about 60 per cent. of the total tuning capacitance so that a fixed value of 5 pF may be added making a total of 15 pF.

With this value

$$R_D = Q/\omega C = 54{,}000 \text{ ohms.}$$

Returning to the question of input resistance and substituting the value of R_D in equation (1), the minimum value of grid resistance

$$R_g(\text{min}) = \pm 147{,}000 \text{ ohms.}$$

Using identical critically-coupled transformers in the grid and anode circuits of the valve at 10·7 Mc/s, where $Q_p = Q_s = 54$ and $C_p = C_s =$ 15 pF, a transfer impedance of $\frac{1}{2}R_D = 27{,}000$

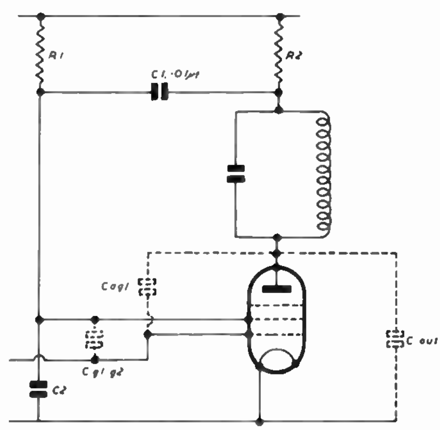

Fig. 6. The screen neutralized i.f. amplifiers.

ohms is obtained in both cases and it is in parallel with this that the input resistance of the valve appears. The ratio between the transfer impedance presented to the grid and R_a(min) is 147,000/27,000 = 5·44.

This is above the specified limit which experience has proved to be satisfactory, and as a result, very little distortion of the response curve occurs.

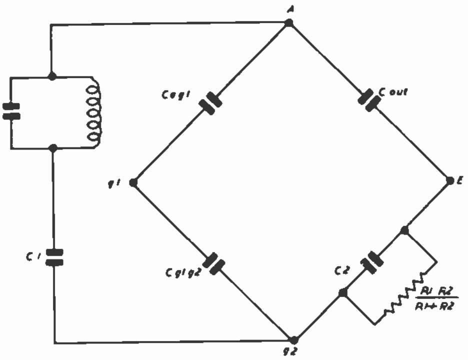

Fig. 7. Equivalent circuit of the neutralized i.f. amplifier.

The amplifier gain per stage $= g_m R_T$
where R_T = transfer impedance of load
and $g_m = 750$ μA/V.

Hence the gain is 20·25, a near approach to that required.

This figure may be increased slightly by reducing the effect of anode–grid coupling through C_{ag} and to do this, a convenient system of screen neutralizing may be used as shown in Fig. 6.

The value of R2 is of the order of 1,000 to 2,000 ohms and is connected in the "earthy" end of the tuned load in order to raise that point above earth potential at intermediate frequency. C1 is large and effectively connects R1 and R2 in parallel so that the small proportion of the total signal output appearing across R2 is applied across C2 and thence between the screen and filament of the valve.

The simplified diagram of Fig. 7 shows the various interelectrode capacitances together with C2 forming a capacitive bridge with g1 and the filament at the points of balance. The parallel combination of R1 and R2 appears across C2, and unfortunately, adversely affects the phase relationship between the voltages in opposite arms of the bridge. Notwithstanding this, however, the negative feedback through the capacitance C_{ag1} can be reasonably well balanced by a correct choice of component values, and stage gain increased by about 20 to 30 per cent.

It is found in practice that component values are not very critical and that fixed values may be used with safety. Thus, there are no circuit complications except for the addition of a single resistor and capacitor.

3.3. *The Detector*[14, 2, 5, 6, 9, 17]

In the application under discussion it is found advantageous to use the ratio detector since, for a given number of stages, greater intermediate frequency amplification may be obtained with a reasonable immunity from amplitude variations in the incoming signal, than from a limiter-phase discriminator system.

The operation of the ratio-detector as a frequency modulation demodulator having inherent amplitude modulation rejection, has been dealt with in the literature, and therefore is not included here.

There are two types of circuit in common use. The first is the balanced or symmetrical type, and Fig. 10 shows how such a detector may be incorporated in a typical receiver. The rectification process is performed by two germanium diodes, the use of which is dictated by the fact that a double-diode valve, having separate cathodes of the directly heated type, even if available, could not be incorporated for obvious reasons.

The second type of circuit is the unbalanced arrangement, which, as shown in Fig. 8, has a single load resistance, one end of which is earthed. There is some economy in the use of this form, but in general, the balanced type will give better amplitude modulation rejection at low frequencies and is preferred in this case.

It is found that output voltage against signal input voltage is not constant for the ratio detector for a given frequency deviation, but this effect can be reduced to a large extent by operating the amplifiers immediately preceding the detector as partial limiters.

With this object, a.g.c. is not applied to the i.f. amplifiers, but instead, high value bypassed resistors are inserted in the grid circuits. By

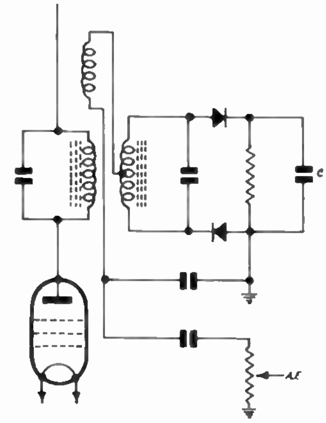

Fig. 8. Unbalanced ratio detector using germanium diodes.

this means, amplification of weak signals is maintained at a maximum, whilst at higher signal levels limiting bias is developed.

The negative voltage appearing across the stabilizing capacitor C varies as the receiver is tuned, and is applied to the grid of the DM70 triode tuning indicator where this is included in the receiver.

4. The Complete Receiver

The first class service area around a B.B.C. v.h.f. f.m. transmitter is stated to be that area in which the field strength exceeds 1 millivolt per metre at a height of 30 ft. above ground and the second class service area 250 microvolts per metre under the same conditions.[18]

For a receiver to serve a useful purpose on the home market, it should have a sensitivity sufficient for it to operate, when used with an efficient aerial, well into the second class service area. It should also be designed to develop a signal/noise ratio under these conditions of greater than 35 db in order to provide a reasonable degree of noise-free entertainment.

Some liberties in aerial installation may be taken with mains-operated receivers as high sensitivities may be obtained quite readily, but

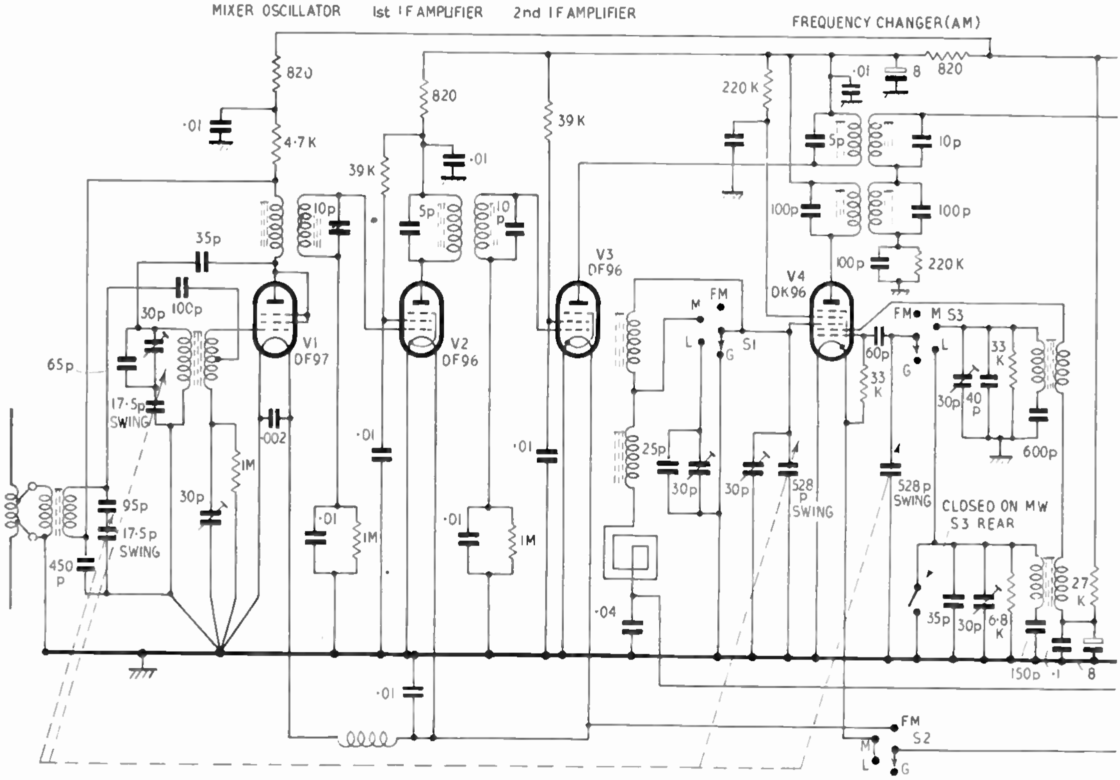

Fig. 10. Theoretical circuit diagram of a 9-valve transportable receiver.

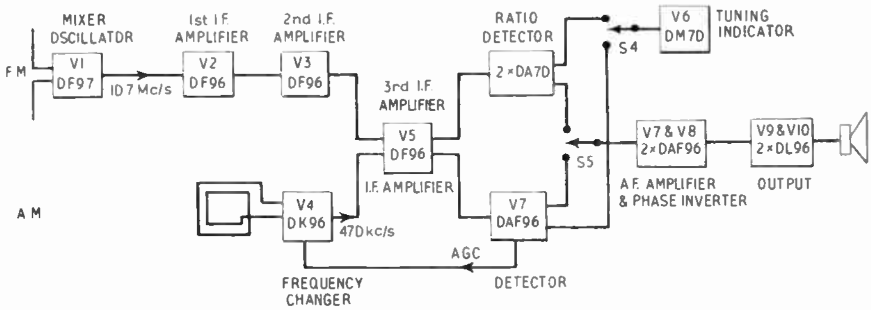

Fig. 9. Block diagram of a 9-valve transportable receiver.

for an economical battery-operated receiver where sensitivity is at a premium, greater attention to the aerial problem is justified, and it is considered necessary to make use of a well elevated multi-element array for reception beyond the 1 millivolt per metre limit.

4.1. *Table Receiver*

For a table model receiver operating from an external aerial, it would be reasonable to assume that a signal of some 400 microvolts would be induced into the aerial so that using a valve line up as indicated previously and shown in Figs. 9 and 10 (giving a usable sensitivity of 20–25 microvolts) would give a fairly wide margin between maximum sensitivity and available signal to cater for falling sensitivity due to battery discharge during operation.

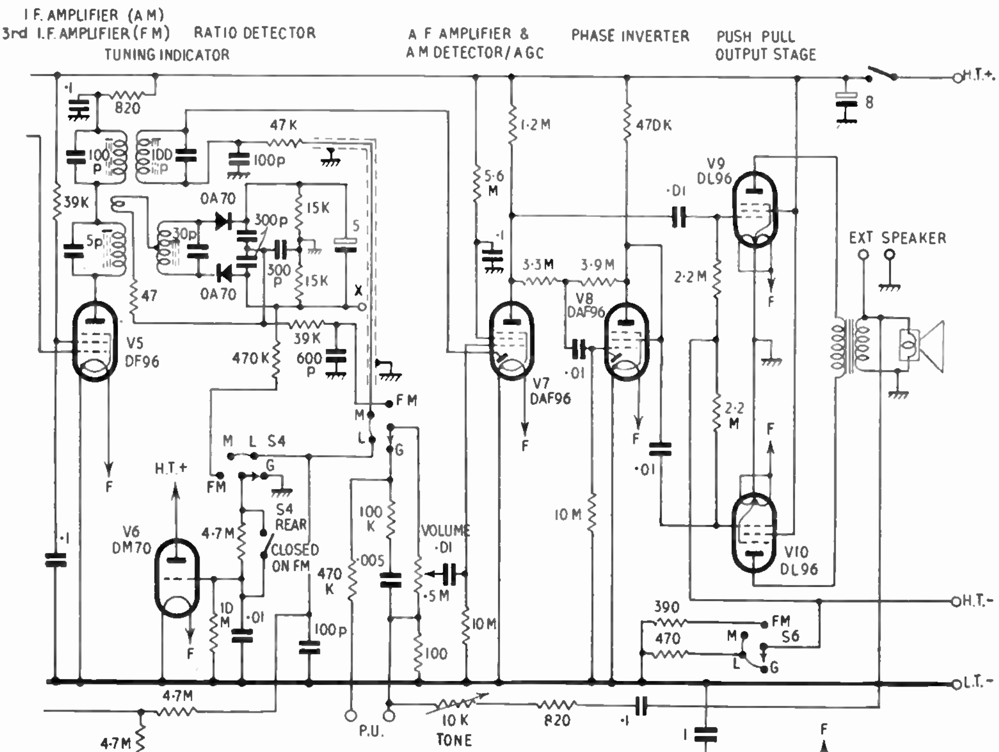

Fig. 10 (*contd.*).

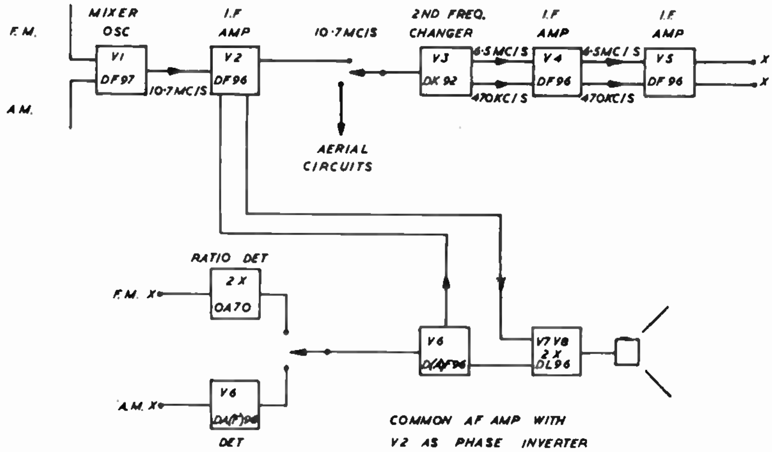

Fig. 11. Block diagram of 8-valve double superheterodyne incorporating reflex phase inverter. Suitable for portable receiver.

This receiver has been shown to be economically possible with a.m./f.m. facilities and with push-pull output. On a.m. the heptode frequency changer (DK96) is switched in and operates with one combined i.f. amplifier at 470 kc/s, whilst on f.m. the DK96 becomes inoperative and the remaining valves form the f.m. receiver complete. When not in use, valves are rendered inoperative by switching off their filaments, a means by which economies are made in battery power.

Performance is such that in areas of field strength exceeding 20 millivolts per metre, good reception may be obtained from an inbuilt contracted dipole mounted on the back of the cabinet, and satisfactory results extend down to 1 millivolt per metre by the use of a full-length picture-rail dipole instead.

4.2. *Portable Receiver*

When the portable variety of receiver is considered, however, the situation becomes automatically much more difficult, as the aerial must be self-contained. It is apparent that the efficiency of a contracted dipole such as is fitted to table model receivers, would be unsuitable for general use, and that a pair of telescopic rods forming a full-length V-shaped dipole makes the most successful compromise. Again making an estimate of the performance required in terms of sensitivity, the actual signal voltage induced into the aerial may be taken as 40 microvolts for usable results.

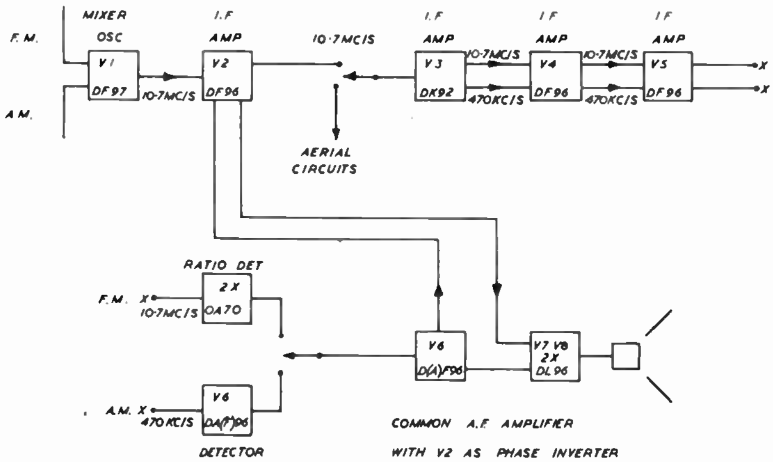

Fig. 12. Block diagram of 8-valve superheterodyne incorporating phase inverter. Suitable for portable receiver.

The problem of producing a higher sensitivity suitable for a portable receiver necessitates the use of an extra stage giving an increase in gain of about ten times. This, of course, can only be achieved at the expense of a further 25 mA of filament current unless economies can be made elsewhere.

If, however, the receiver is also to have a push-pull output stage (a desirable feature in a high quality receiver), then instead of operating the output stage from a two-valve self balancing phase-splitter as is usual, the 1st i.f. amplifier can be used to function also as a reflex a.f. phase-inverter with the a.f. voltage developed on the screen. This system has been found quite satisfactory and results in the extra stage of i.f. amplification being available without sacrificing the advantages of a push-pull output stage, and without the burden of another valve.

H.t. consumption, however, has become a little higher, and as battery space is normally limited in a portable receiver, valuable economies in both l.t. and h.t. power may be made by the simple addition of an "economy switch" made to isolate one half of each of the output valve filaments from the l.t. supply. This, of course, reduces sensitivity and maximum available output power, which in certain circumstances may be sacrificed in favour of battery life.

Various combinations of circuits may be used, one of which is the double superhet as shown in Fig. 11. By this means, the self oscillating mixer, together with the 1st i.f. amplifier, may operate at 10·7 Mc/s as before, after which the frequency may be converted to about 6 Mc/s and at this lower frequency, higher amplification is possible.

Figure 12 shows a further eight valve combination using four i.f. amplifiers all operating at 10·7 Mc/s, together with a reflex a.f. amplifier and push-pull stage.

Both these receivers are capable of good quality reception well into the second class service area when operated at ground level and are therefore extremely satisfactory for portable use.

5. Acknowledgments

The authors wish to thank members of the staff of the Ever Ready Co. (G.B.) Ltd. for their assistance, and the Directors of the Company for permission to publish this paper.

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NEW BRITISH STANDARDS

The British Standards Institution has recently issued the following new and revised Standards. Copies may be obtained from the B.S.I. Sales Branch, 2 Park Street, London, W.1.

B.S.415 : 1957 **Safety requirements for electric mains supplied radio or other electronic apparatus for acustical or visual reproduction.** Price 6s.

The requirements deal with the design, construction and electrical performance from the safety aspects only; they include protection of the user from electrical shock, and from flying glass in the event of the implosion of a cathode ray tube in the apparatus. Requirements limiting the temperature rise in the event of a component failure are also specified. Attention is drawn to possible danger to the health of the user from X-radiation from high-voltage equipment—for example, from cathode-ray tubes and from rectifier valves.

The standard requires tests to be supplied under "normal" conditions and also under "fault" conditions. The introduction of "fault" condition tests is an innovation in British Standards, although it has for some years been a feature of the national regulations of some other countries. Experience has shown that, in a complex piece of equipment such as a radio or television receiver, there are some components in which an insulation failure cannot be considered to be impossible. In anticipation of such possible failures the standard requires the application of short-circuit tests and makes compliance with it conditional upon there being no danger of fire or shock when any combination of certain specified artificial short-circuits is applied.

B.S.530 : 1948 **Graphical symbols for telecommunications.** Suppt. 4 : 1956 **Miscellaneous recommendations and symbols (including transistors.** Price 3s. 6d.

The Inter-Service list of symbols has previously differed from those in B.S. 530 but the Services have agreed to use, instead, B.S. 530 (and its supplements) together with an addendum listing these differences. Supplement No. 4 has been drafted mainly with a view to removing these differences.

Further guiding principles for the preparation of circuit diagrams are given and also new or modified symbols which reflect advances in technique. Symbols for transistors and allied devices form an important part of the Supplement. In drafting these, attention has been paid to American practice.

B.S.530 Suppt. 5 : 1957 **Graphical symbols for telecommunication. Functional symbols for switching diagrams.** Price 2s. 6d.

This Supplement gives symbols suitable for indicating the functional operation of working electronic circuits without giving details of all the components and without reference to notes describing the circuit operation.

B.S.1000A : 1957 **Universal Decimal Classification—Abridged English Edition.** Price 42s.

Preceding the General Introduction, there is a new Note on the Dewey D.C. and U.D.C. which explains the resemblances and differences between these two related systems; the General Introduction itself has been recast as a concise guide for the less experienced classifiers and as a simple explanation as possible for the uninitiated and curious. Equally important is the freshly compiled Alphabetical Subject Index, which now runs to well over 20,000 entries, as against some 2,000 in the 1948 edition.

In both the Main and Auxiliary Tables, the terminology has been extensively revised, and significant "Extensions and Corrections to the U.D.C." authorized to date have been incorporated. Though slightly extended, these abridged tables are still about one-tenth of the length of the full edition; they are therefore sufficiently detailed for most general use whilst providing the essential "background" for specialists.

B.S.R.2 : 1956 **Basic characteristics of radio service selection and intercommunication systems for civil aircraft.** Price 3s.

Specifies the inter-relating characteristics of aircraft radio receivers, transmitters and audio amplifiers that will enable them to be operated in conjunction with a mixing system or directly with the types of headphone and microphone in general use. The audio-frequency output of a number of receivers, transmitters and amplifiers may be routed to any member of the aircraft crew, and microphones and telegraph keys routed to the transmitters. Input and output powers and loads, methods of volume control and limits of interference are specified, and an Appendix gives the preferred arrangements of microphone/telephone connections for different handsets and headsets.

A selection of abstracts from European and Commonwealth journals received in the Library of the Institution. All papers are in the language of the country of origin of the journal unless otherwise stated. The Institution regrets that translations cannot be supplied.

621.372:621.385.16

The attenuation-mode of the single wire transmission line. G. SCHIEFER. *Archiv der Elektrischen Ubertragung*, **11**, pp. 35-40, January 1957.

On the basis of the strip model devised by Sensiper the helical wire line is thought of as an inhomogeneous delay line, along which an infinity of partial modes propagates, rather than a single mode. For the two limiting cases of a "narrow" and a "wide" strip helix the attenuation constant is calculated taking all partial modes into account. For the complicated sums of Bessel functions appearing in this computation practical approximations are given. Some results deviate rather considerably from previous solutions where only the dominant mode is considered. The solutions are discussed by reference to measured values so that detailed diagrams are finally available for determining the attenuation constant in the entire range of technical interest.

621.373.5

On crystal-controlled oscillators. G. BECKER. *Archiv der Elektrischen Ubertragung*, **11**, pp. 41-47, January 1957.

A great part of the customary crystal oscillator circuits can be reduced to basic arrangements with an underlying general condition for oscillation. Two categories of circuits can be distinguished and prove to be series-resonant and parallel-resonant circuits respectively.

621.375.121

Two-terminal video couplings. DHARMAJIT GUPTA SARMA. *Journal of the Institution of Telecommunication Engineers, India*, **3**, pp. 21-31, December, 1956.

Critically damped two-terminal video interstage coupling networks have been studied. Some infinitely complex two-terminal networks having good monotonic transient response are derived and a number of practical networks of low orders of complexity suggested. Third and fourth order coupling impedances for optimum transient response have been obtained.

621.383

Properties of very thin leaves of glucinium and their applications to the modulation of light and to television. M. AUPHAN. *Onde Electrique*, **36**, 1040-1045, December 1956.

By evaporation in vacuum it is possible to obtain leaves of glucinium, with both surfaces free and having a thickness between 200 and 2,000 angstroms. The process is to deposit the glucinium by evaporation on to a temporary support which can be sublimed, and which, after evaporation, leaves the leaf free. Gratings can be used during evaporation and interesting structures may result; in particular, a leaf of glucinium can be stretched parallel to a glass support and at a very small distance from it. This distance can be maintained constant by a number of separators. If an electron flux impinges on such a structure, it passes through the leaf, which is deformed by the electrostatic force caused by secondary emission. In this way a light modulator is obtained which can be used in monochrome or colour television.

621.385:621.375.2

The life and reliability of valves. K. RODENHUIS, H. SANTING and H. J. M. VAN TOL. *Philips Technical Review*, **18**, No. 7, pp. 181-192, December 1956.

The authors define "reliability" as the reciprocal of the failure rate, i.e. the percentage failures in a batch of valves after 1,000 hours operation. They regard the end of a valve's useful life as the moment at which the failure rate begins to increase. A useful life of 10,000 hours and a failure rate of 0·5 per cent. per 1,000 hours have already been reached with "Special Quality" type valves; it is hoped to be able to reduce the failure rate to 0·1 per cent. per 1,000 hours. Valve failures may be either gradual or sudden. The causes of both categories of failure are discussed separately and a description is given of the measures taken in recent years to reduce the incidence of valve defects. The spread of characteristics and shock and vibration resistance are also discussed. In conclusion the desirability of close cooperation between manufacturer and user is underlined.

621.385:621.3.032.213.6

A pulse method for measuring the intermediate layer in oxide cathodes. A. LIEB. *Nachrichtentechnische Zeitschrift*, **10**, pp. 88-89, February 1957.

The paper describes a method for measuring the electric values in intermediate layers which grow in the cathodes of amplifier valves.

621.385.1:621.373.4

Design considerations for travelling-wave tubes for output stages in radio links. M. MULLER. *Nachrichtentechnische Zeitschrift*, **10**, pp. 11-15.

Simple relations between design data and operational data for travelling-wave tubes are developed. Under the assumption of a constant magnetic focusing field of 600 gauss and of discrete voltages for the operation, a list of data for travelling-wave tubes covering 2–10 Gc/s and 1–200 W is calculated. This list has the nature of a proposal for standardization. Reasons and justifications are given which show that an input drive and a level of ½ mW for output travelling-wave tubes is possible.

621.391

The information contained in noise sounds.—R. EMKEL. *Nachrichtentechnische Zeitschrift*, **9**, pp. 493-498, November 1956.

The proof of a correlation between the information contained in speech and the values in a field for sound produced by speech has a practical significance for telecommunication engineering. The noise sounds are of particular interest because of their wide frequency spectrum and their low level. The importance and the dynamics of consonants in the audio transmission of information is investigated by means of hearing tests.

621.391

Geometrical interpretation of some results obtained from calculations of channel capacity. CLAUDE E. SHANNON. *Nachrichtentechnische Zeitschrift*, **10**, pp. 1-4, January 1957.

The channel capacity of discrete channels not containing memory devices can be investigated by geometric means. This approach leads to a number of new convexity properties and to a certain knowledge concerning the nature of transmission rate and capacity.

621.396.11.029.53

The path attenuation in ionospheric propagation of hectometric radio waves (550-1600 kc/s). M. SCHOLTZ. *Nachrichtentechnische Zeitschrift*, **10**, pp. 6-11, January 1957.

A summary from original papers constitutes a report on night-time propagation in the medium wave-band. Curves for attenuation (as a function of frequency) which have internationally been acknowledged in the last 22 years are compared with each other. The fading phenomena are represented by means of parameters: statistical scattering, frequency of fading and systematical variations.

621.396.69

The 'asmodulor" process—a new technique of construction for electronic apparatus. L. PATERNAULT. *Onde Electrique*, **36**, pp. 1031-1039, December 1956.

Electronic apparatus can nearly always be split into a number of homogeneous sub-assemblies, and this is behind the "asmodulor" process which enables an electronic assembly to be realized by an assembly of a number of identical mechanical units carrying the components. By reason of the mechanized production and small labour cost, this leads to a reduction in price, and it favours mechanized test methods more than conventional type equipment. Big developments in this process are particularly foreseen in the military and aeronautical fields. The article describes in detail the process which is based on the American "Project Tinkertoy," discusses its applications and gives some information on the special components employed.

621.396.822:621.396.96

Measurement of noise factor in centimetric radar. N. N. PATLA. *Journal of the Institution of Telecommunication Engineers, India*, **3**, pp. 32-37, December 1956.

Two practical methods for measurement of noise factor in centimetric radar receiver are described. The first method applies to the i.f. amplifiers (30-60 Mc/s), and the second method applies to the system consisting of the mixer followed by the i.f. amplifier (3000-10,000 Mc/s). It has been found that results obtained by these methods have been consistent to about 2 per cent. with exceptions due to traceable causes. Detailed calculation of the overall noise factor of the crystal mixer followed by an i.f. amplifier has also been presented with a critical discussion.

621.396.933:629.13.018

The application of radio frequency predictions in aeronautical communications. A. FOXCROFT. *Proceedings of the Institution of Radio Engineers, Australia*, **18**, pp. 7-12, January 1957.

This paper reviews briefly the frequency prediction data which is available in Australia, describes certain frequency prediction problems peculiar to aviation and considers their relationship to the frequency allotment plans of the aeronautical mobile service. The application of predictions to the planning of aeronautical communications is discussed having regard to maximum and minimum usable frequencies and limitations imposed by the allotment of frequencies. Finally the day-to-day use of prediction data is described and illustrated by examples.

621.397.611.2:778.53

A flying-spot scanner for televising 35 mm film. F. H. J. VAN DER POEL. *Philips Technical Review*, **18**, No. 7, pp. 193-201, December 1956.

In a flying-spot scanner for televising 35 mm film, the film moves with constant speed and this complicates the problem of interlaced scanning. The problem is solved by using two optical systems. The two objective lenses have to be closely identical and should be as free as possible from distortion and vignetting. Since there is always some shrinkage in a film and since the drive is effected in the normal way via a sprocket, the film does not in fact travel at a strictly constant speed. A detailed explanation is given of why this is not troublesome.

621.397.62.029.62

Television receiver for metric waves. V. BIGGI. *Onde Electrique*, **36**, pp. 1021-1030, December 1956; **37**, pp. 55-67, January 1957.

The author examines the problem of designing a receiver intended for long distance direct reception of a television signal from a main transmitter radiating on metric waves. The quality of the signal given by the receiver, both video and sound, must be good enough to modulate a satellite transmitter. After having given particular attention to the design of the input circuit in order to give a low noise factor, and to the selectivity requirements imposed by asymmetric sideband reception, the author suggests an original method of automatic gain control for positive modulation. An application of this method illustrates how direct reception provides a simple means for the transmission of the signal to a point for re-radiation.

621.397.813:621.397.331.2

Defects in picture-quality in the image orthicon television camera tube. R. THEILE and F. PILZ. *Archiv der Elektrischen Ubertragung*, **11**, pp. 17-32, January, 1957.

The image orthicon is the most sensitive all-purpose television camera tube. There are, however, characteristic defects in the picture quality, especially when televising high-contrast scenes. In addition to the discussion of the well known quality parameters, the article mainly deals with these typical and characteristic spurious signals, which appear in the picture as white border lines, loss of definition on horizontal white-to-black transients and geometrical distortion (blooming). It was found that the white border line effects are due to capacitive coupling between adjacent picture elements which are more pronounced when the target capacitance is small. On the other hand, the edge distortion and the geometrical errors can be explained by additional deflection of the scanning electrons in front of the target by the potential pattern of the stored charges. Summarizing the results, possibilities are indicated how these typical spurious signals can be reduced by suitable construction and operation of the tube.