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Commentary

AN International Convention on Transistors and Associated Semiconductor Devices has been arranged by the Institution of Electrical Engineers and will be held at Earls Court from 21 to 27 May. The Convention will be backed-up by an International Scientific Exhibition where the leading laboratories and manufacturers of the world will be displaying their latest developments in the transistor field, and which will be open during the period of the Convention. The Convention will be opened by Viscount Hailsham, the Lord President of the Council, who is the Minister responsible for scientific research and the introductory lectures will be given by Professor J. Bardeen, Dr. W. M. Brattain and Dr. W. B. Shockley who, only a little over ten years ago, in August 1948, jointly announced the invention of the transistor; the scientific importance of which was recognized by the award of the 1956 Nobel Prize for Physics.

In the comparatively short time since its invention the progress in the technology of the transistor and its application and, indeed, in the whole field of semiconductor physics has been enormous. Indeed, before the second World War few people had even seen a piece of germanium and it received its first real attention in the wartime scramble to develop better materials for microwave diodes. By the end of the war polycrystalline germanium, of what was then considered to be high purity, was available and it was during a basic research study of its surface properties, at the Bell Telephone Laboratories, that the transistor action was discovered.

In the intervening years semiconductor research has greatly increased and it is now a major scientific activity throughout the world. So far, almost all of this effort has been concentrated on germanium and silicon which, in some respects, is a little surprising since, by a conservative estimate, there are at least a thousand semiconductive substances. It is, however, fortunate that transistor action was first observed on germanium as it is one of the simplest of all solids, a factor which has contributed to the fact that it is, today, one of the most completely understood of all solid materials. Both germanium and silicon are now available as the purest and most perfectly formed crystals which scientists have had for their study and, even more important, this purity and structural perfection can be altered in a controlled manner by a number of techniques. It is the electrical effects of these alterations that are the distinguishing features of semiconductors.

Research into the basic properties of semiconductors has been backed up by sound progress in the manufacturing techniques necessary to make reliable transistors. To the production engineer the original point-contact transistor appeared as a product simple in structure, rugged and enclosed in plastic that would lend itself easily to mass production. Unfortunately, it did not turn out quite like that, for problems were soon encountered in terms of reliability, reproducibility and range of performance. The advent of the grown junction and alloying and diffusion techniques have all posed their considerable problems to the production engineer who has had to strike a balance between the present—to provide adequate facilities for existing techniques—and the future to be prepared for the new and better way in this rapidly evolving field. Indeed, so rapid has been its progress that it is estimated that in the last ten years or so some 800 different types of transistor have been developed while in Britain alone about twelve firms are engaged or about to engage in transistor manufacture.

With regard to the application of the transistor, while some of the early forecasts that it would almost entirely replace the thermionic valve have proved to be rather wildly optimistic, its impact has been significant and there is now a very large number of transistors in satisfactory use in a very diverse range of electronic equipment. The initial phase where a designer, or manufacturer, used transistors merely to be fashionable has, fortunately, largely passed and the transistor is now being used in applications where its virtues of reliability, small size, lower consumption and low operating voltages make it a more suitable circuit element than the corresponding thermionic valve.

The future of the transistor is difficult to forecast. It seems probable that for many applications germanium will be superseded by, perhaps, silicon which has an operating temperature range up to about 159°C as against some 50°C for germanium; or it may well be that both substances will be outmoded by another of the semiconductor materials, the properties of which have not yet been explored. It is, however, safe to say that as the basic knowledge of semiconductors expands and new and improved manufacturing techniques become available the operating range of both temperature and frequency and, consequently, the field application will increase in step.

It is on all counts an appropriate time to hold a Convention and Exhibition of the type planned and it should play a most useful part in consolidating the knowledge that has already been gained and in pointing the way ahead for future developments.

Printed Circuits Applied to Broad-Band Microwave Links

By R. Rowland*

This article describes how printed circuit techniques may be applied to the design of base-band and i.f. equipment for broad-band microwave links and discusses the advantages to be gained thereby. It is shown that, not only are the units cheaper to produce, but enhanced performance, stability and uniformity is obtained. Further, maintenance procedure is simplified and test equipment requirements are greatly reduced. Some typical printed circuit units are described.

(Voir page 312 pour le résumé en français: Zusammenfassung in deutscher Sprache Seite 317)

THE use of printed wiring or printed circuits in electronic equipment is often considered to be a method of reducing manufacturing cost or reducing space, and to those not familiar with the better examples of this technique it might be thought that these advantages were gained at the expense of quality in the resultant product. It is the purpose of this article to show that, far from this being the case, printing of wiring or circuits can result in considerable improvements in the quality of the product and better performance throughout its life, in a field in which the highest quality is considered essential. In the applications with which this article deals the technique was applied primarily to improve reliability and performance, and had it resulted in increased manufacturing cost it would still have justified its adoption. The fact that it has resulted in reduced manufacturing cost must be considered as a secondary though desirable advantage.

The term 'printed circuit' is herein used to describe an assembly in which some of the circuit elements, in addition to interconnecting leads, are formed by the conducting pattern of the printed board. Where the conducting pattern is used simply to interconnect conventional components the term 'printed wiring' is preferred. The equipment described comes into the first category.

Printed circuit boards may take several forms, the conducting pattern, formed by one of the many available processes, being on one or both sides of a board of insulating material. The type selected as being most suitable for the applications here described has a copper conductor pattern on one side of the board, formed from one-side copper clad stock board by a photographically controlled etching process. The reverse side of the board, on which associated components are mounted, has a legend printed on it indicating positions and identity of these components.

Application

Present-day broad-band microwave link systems designed for long distance transmission of multi-channel telephony and for television signals depend for their successful operation on the establishment and maintenance of critical circuit conditions at many of the stages in the transmission system. Notable among these stages are broad-band modulators, demodulators and intermediate frequency amplifiers, together with their associated base-band amplifiers, and it is with these units that this article primarily deals. The principles discussed are, however, applicable to other parts of the system, such as control and supervisory equipment, etc.

The standard of performance required from this equipment is shown in the following brief details of i.f. and base-band units which may form part of a 600-channel or television link.

Intermediate-Frequency Amplifiers

(1) Mid-band frequency — 70Mc/s
(2) 'Flat' bandwidth — 20Mc/s
(3) Uniformity of amplitude—frequency response: 60Mc/s to 80Mc/s — ±0·1dB
(4) Permissible slope of Amplitude characteristic 60Mc/s to 80Mc/s — 0·25dB
(5) Group-delay variation 64Mc/s to 76Mc/s — 1·0mμsec
(6) Input and output impedance — 75Ω
Modulus of reflection coefficient not worse than 2 per cent at 70Mc/s, not worse than 10 per cent at 60Mc/s and 80Mc/s.

Modulator plus Demodulator

(1) Intermediate frequency — 70Mc/s ± 50kc/s
(2) I.F. bandwidth to −0·25dB points — 20Mc/s
(3) Base-band — 25c/s to 7Mc/s
(4) Base-band response — 25c/s to 7Mc/s within 0·2dB
L.F. response not greater than 1 per cent tilt on 50c/s square wave.
(5) Overall intermodulation distortion not worse than −70dB.

Base-band Amplifiers (telephony)

(1) Bandwidth — 60kc/s to 3·2Mc/s
(2) Uniformity of amplitude/frequency characteristic within 0·1dB.
(3) Intermodulation distortion better than −80dB.

The design and construction of units fulfilling the above requirements does not present undue difficulty in the laboratory. Given time and the necessary test equipment the designer can produce critically-aligned complex circuit configurations which meet the requirements with ease. It is in the quantity production and maintenance in the field of such units that difficulties arise. Some of the difficulties are outlined in the following section.

Production and Maintenance Difficulties with Conventional Circuits

Taking an i.f. amplifier as a typical example, this may contain six or seven stages of amplification requiring the

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adjustment of twenty or more inductors in the interstage coupling networks. In order to achieve the required overall characteristics each stage has to be aligned with great precision. In quantity production such adjustments are carried out by a test department, and require the use of expensive equipment for the display and measurement of frequency response, group-delay variations, etc. In addition these operations call for a high degree of skill and technical ability on the part of test personnel and require extremely detailed instructions to be originated by the designer. Facilities must also be built into the equipment to enable the required measurements to be made. There is a limit to the complexity of alignment procedure which can reasonably be dealt with in this manner and the designer is sometimes forced to adopt less efficient circuit alternatives which lend themselves more readily to simple alignment.

Similar considerations apply in the case of maintenance in the field. In order to keep equipment operating at peak efficiency and to enable re-alignment to be carried out in

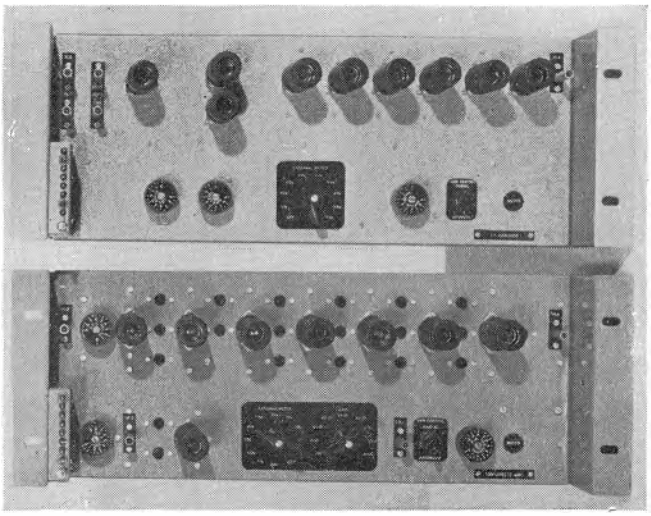

Fig. 1. Front view of the i.f. amplifier showing, on the bottom, a conventionally wired amplifier, and on the top, one employing printed circuits

the event of component replacement, test equipment similar to that described above must be available to the system operator. The cost of such equipment can represent a considerable portion of the total cost of a system.

It follows that any technique of equipment design and manufacture which results in a reduction of test equipment and testing time in manufacture and maintenance is worthy of serious consideration. If, in addition, the resulting equipment shows an improved performance at no extra cost, there occurs a combination of circumstances which represents a significant advance over conventional techniques such as those previously described. The technique of circuit printing offers such advantages.

Characteristics of Etched-Foil Printed Circuits

Early experimental work with etched-foil printed circuits revealed certain definite characteristics, among the most important of which are the following:

(a) *Repeatability*

Using a photographically controlled resist pattern, all etched boards produced from one particular master have such a high degree of similarity as to be virtually identical.

(b) *Ease of Manufacture*

The printed circuit process used in these applications has the essence of simplicity which makes it eminently suitable for quantity production of the type of equipment envisaged.

(c) *Reliability*

Assembly of components, soldering, etc., is made so simple that faults are virtually eliminated and wiring errors are virtually impossible.

(d) *Low cost*

The process is one which is economical for large or small quantities of similar units, and shows reduction in manufacturing costs compared with conventional wiring systems.

(e) *Testing*

By taking advantage of the repeatability characteristic, simplified test procedures are possible.

(f) *Replacement*

Printed circuits lend themselves to easy breakdown into sub-units suitable for complete replacement in the event of fault occurrence.

Experimental Investigation

In order to determine with what degree of accuracy printed circuits of this class could be produced, certain basic circuit units were dealt with. These included broad-band i.f. interstage coupling networks, i.f. output matching networks and i.f. input matching networks. It was found possible to reproduce inductors having values in the range $0{\cdot}1\mu$H to $3{\cdot}0\mu$H with a high degree of accuracy. These were printed in the form of a flat spiral.

Measurements made on tapped coils showed that it is difficult to achieve tight coupling between the sections of such coils, and it was therefore felt that circuits using separate untapped inductors with no mutual inductive coupling between them would be the most satisfactory. The type of interstage coupling network found most useful for broad-band i.f. amplifiers was the double-tuned transformer. This circuit may readily be converted to its equivalent T or π configuration which contains three separate inductors having no mutual inductive coupling. There are certain regions of primary to secondary impedance ratio where the T or π equivalent is not physically realizable for a given bandwidth. In such cases it is possible either to alter the impedance ratio by adding capacitance to primary or secondary, which reduces the gain $\times$ bandwidth product, or to use two circuits in cascade, each having an impedance ratio for which the elements are real. In the case investigated, however, the straightforward equivalent circuit was possible. The T configuration was chosen as this results in smaller values of inductance than are required for the π case.

Experimental printed boards were designed having two stages of amplification, an input matching circuit, one interstage network and an output matching network. Normal design methods were applied, with the alterations in configuration previously referred to. Certain of the required lumped capacitors were also printed initially in a single-sided interleaved form, but were inconvenient from the point of view of space requirements and were found to have appreciable positive temperature coefficients. As suitable conventional capacitors were available in small physical size these were used where additional tuning capacitance was required, and conventional capacitors were used for coupling and decoupling purposes.

A satisfactory method of alignment was evolved, which is described in a later section dealing with design procedure.

Having produced an experimental board with the required frequency response, matching, etc., copies were made under conditions simulating normal production with components subjected to the specified tolerances, etc. The resulting models were tested and their characteristics compared with those of the original. These comparisons showed that boards having identical characteristics could be produced without undue difficulty. It was necessary, in order to ensure good repeatability, to use components having close tolerances in certain cases such as for circuit damping, tuning, etc. High stability resistors with a tolerance of resistance value of $\pm$ 1 per cent were used for damping purposes and capacitors of similar tolerance were used where required for tuning purposes. Decoupling components were to normal tolerances of $\pm$ 10 per cent to $\pm$ 20 per cent.

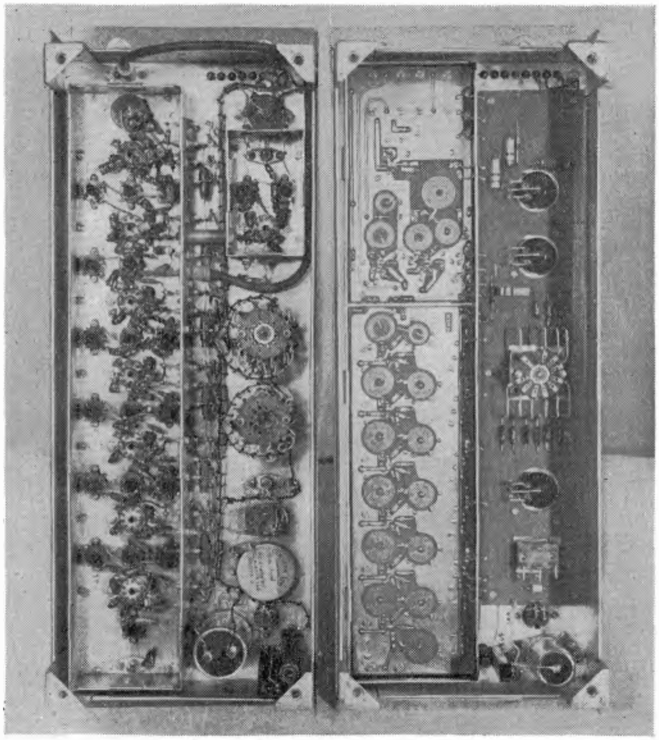

Fig. 2. Rear view of the i.f. amplifier

More complex boards having larger numbers of stages were then produced, eventually resulting in a six-stage amplifier. An important aspect which was examined at this stage was the effect on performance of valve changes. It was realized that for production purposes it would be impossible to select valves having very tight limits on interelectrode capacitance, mutual conductance, etc., and the effect of changing valves was carefully examined, using sets of valves selected to be just within upper or lower limits. The result of this examination showed that considerable changes can occur with straightforward circuits, resulting in the performance of the unit becoming unacceptable. Various methods were tried to reduce this effect and finally a satisfactory solution was found. It is not possible at the present time to publish details of the procedure adopted, but it is sufficient to add that it is possible to replace valves without selection and to maintain optimum performance.

A further important consideration was the effect of adverse climatic conditions on the operation of equipment containing printed circuits. Various base materials are available for the printed boards, including resin-bonded paper, resin-bonded glass fibre and silicone-bonded glass fibre. Tests under extended tropical conditions showed that all currently available materials which could be considered suffered in some degree from moisture penetration. Various types of coating were therefore tested. Eventually a suitable coating resin was found which gave a high degree of protection when applied in the form of a thin film to the assembled printed circuit boards. The resistance to environmental conditions of a paper-based board so treated is several orders better than an untreated board of the best available materials. For the type of equipment herein described high grade paper-based laminates have been found very satisfactory.

As a result of the above investigations it was felt that units such as those specified earlier could be designed with printed circuits, having considerable advantages over conventional units. The following section describes a typical printed circuit i.f. amplifier which was designed to replace a conventional-wiring amplifier.

Printed Circuit Intermediate Frequency Amplifier

This amplifier forms part of a broad-band microwave link system and is used in conjunction with a low-noise preamplifier. The performance specification is as follows:

Mid-band frequency	70Mc/s
Amplitude/frequency characteristics	$\pm$0·1dB 60Mc/s to 80Mc/s $-$3dB 50Mc/s and 90Mc/s
Group-delay variation w.r.t. 70Mc/s	4·5mμsec at 60Mc/s and 80Mc/s 2·0mμsec at 64Mc/s and 76Mc/s
Maximum gain, nominal	42dB
Input signal level, nominal	200mV r.m.s.
Input signal range	$-$30dB to +10dB
Output signal level, nominal	700mV r.m.s.
Output signal variation over input signal range	$-$1·0dB to +0·5dB
Input impedance	75Ω
Modulus of input reflection coefficient	
Not worse than	5 per cent at 70Mc/s
Not worse than	10 per cent at 60Mc/s and 80Mc/s
Output impedance	75Ω
Modulus of output reflection coefficient	
Not worse than	10 per cent, 60Mc/s to 80Mc/s

The amplifier contains six stages of signal amplification, an additional buffer stage providing a separate output for branching or monitoring purposes, and an automatic gain control circuit. Three printed circuit boards are used, the first containing the amplifier proper, the second containing the buffer and a.g.c. circuits and the third the associated components, interconnecting wiring, and valve monitoring facilities. The boards are mounted in a 7in wide dished panel designed for mounting in a standard 19in rack. Valve screening cans are mounted on the front panel to provide good heat dissipation and certain components such as switches and potentiometers are mounted in such a way that they can be replaced without disturbing the printed circuit boards.

Photographs of the amplifier are reproduced in Figs. 1 and 2 which show the conventionally wired equivalent panel for comparison. Fig. 3 is a photograph of a typical circuit board and Figs. 4 and 5 show characteristics of the amplifier.

Outline of Typical Design Procedure

The design of such a unit would be undertaken in the following manner.

THEORETICAL DESIGN

This stage of the design follows normal methods except that circuit configurations suited to printing are used. In many cases circuit units previously developed in printed form can be utilized, with minor modifications where necessary. Previous experience gained from similar applications facilitates accurate estimation of stray capacitance, etc., enabling the designer to make a more accurate prediction of the final performance than is usually the case with conventionally wired circuits.

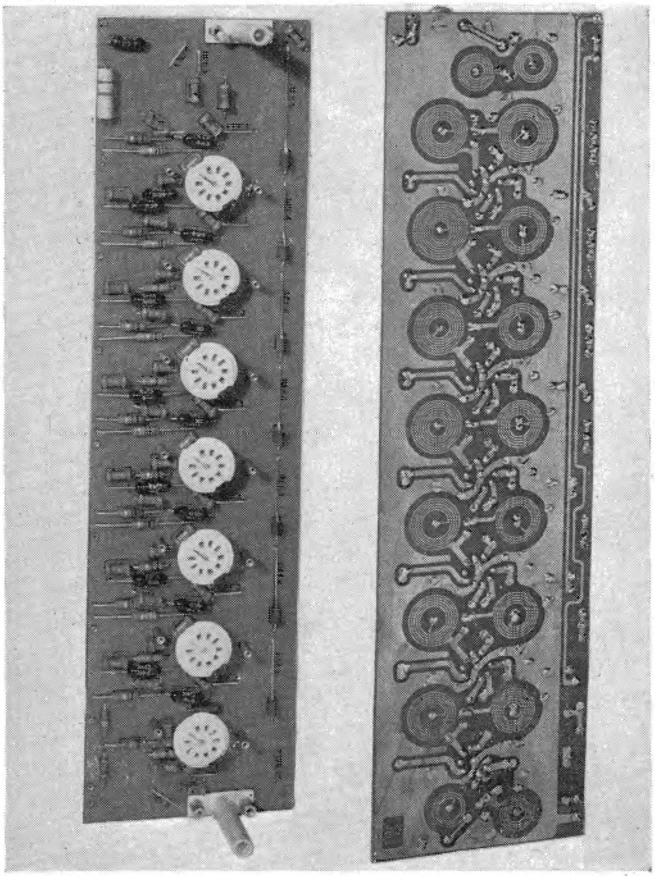

Fig. 3. Front and rear views of a typical circuit board

LAYOUT

At this stage 'know-how' acquired from previous experience is particularly valuable, eliminating much trial and error in arriving at satisfactory layouts, using suitable methods of mounting, avoiding undesirable coupling effects, etc. However, a clear idea of what is required from each circuit element and a liberal application of common sense will go a long way towards achieving neat designs meeting the required performance.

Layouts are made first in a rough stage, freehand, on squared paper at twice final size. Considerable advantages from a production point of view are gained by basing all connecting points on a grid or matrix. This enables manufacturing processes to be handled by automatic machinery, resulting in better repeatability and reliability, in addition to lowering costs. The preferred grid has a spacing of 0·1in in horizontal and vertical directions, all connecting points being made at grid intersections. The squared paper used for layout is therefore in units of 0·2in. Both pattern and components are indicated on the rough layout, suitable 'shorthand' symbols often being used to save time.

PREPARATION OF WIRING MASTER DRAWING

Master drawings are made at twice final size on a special dimensionally stable transparent plastic material. The material is fastened over a master grid sheet of high accuracy which determines the positions of all connecting points. Using the rough layout as a guide, all connecting points are drawn in as circular 'pads' using a specially prepared dense matt ink. Wiring patterns are then drawn where required, the width and spacing of these being determined by electrical requirements. As a general guide a minimum spacing and width of 0·05in is preferred, although in certain circumstances, such as for inductors, these dimensions may be reduced to 0·025in. These dimensions are final size, and are, of course, doubled on the master drawing. Outlines of earthed areas are next drawn and filled in. Inductor spirals, cut from standard black and white prints, are then fastened into position with adhesive; the number of turns required on each spiral is determined from curves relating number of turns to inductance for the particular spiral used. Finally the serial number and various production control markings are added.

An alternative method of constructing master drawings with adhesive tapes is used for some printed wiring work but is not suitable for the preparation of the more complex patterns of printed circuits.

PREPARATION OF LEGEND MASTER DRAWING

The function of the legend, which is printed on the component side of the board, is to indicate the positions and identity of all components.

The wiring master drawing is laid face downwards on the grid sheet and the sheet on which the legend is to be drawn is placed over this. Positions of all connecting points are then traced on the sheet from the wiring drawing. Outlines of components are then drawn in using the rough layout again as a guide. Component designations are then added, followed by the board reference number and alignment markings.

PREPARATION OF MASTER NEGATIVES AND COPIES

Master negatives of wiring and legend are made by normal photographic methods, with a plate camera set to give a two to one reduction from the master drawings. The master plates are thus at final size. Accuracy in setting up the camera is, of course, essential, both as regards dimensions and focus. The resultant negatives are checked carefully for correct dimensions and focus, the latter check being facilitated by a control pattern which is included in the drawing.

Copies of the master negatives are made by contact printing methods. This involves the intermediate preparation of contact film positives which are used later in the preparation of production drawings. High stability film is normally used for production negatives, although glass is used in some cases.

PREPARATION OF CIRCUIT BOARDS

This stage is essentially a copy of one of the production processes, and is carried out at a central plant set up for the purpose. Space does not permit a description of the process here.

Testing and Adjustment

The assembled board is now fitted to the appropriate chassis to undergo testing and adjustment. During the development stages it is often necessary to make adjustments to inductors, etc., and the usual sweep methods are used to indicate amplitude/frequency characteristics, etc. By the use of dust-iron and copper disk probes it is possible to make temporary alterations to inductors to determine the effect on characteristics, and when permanent alterations are required coils are pruned by peeling away turns from their centres. As it is obviously only possible to make permanent reductions in inductance it is arranged in the initial design to print all inductors somewhat larger than required, to permit precise adjustment. When a satisfactory performance has been achieved

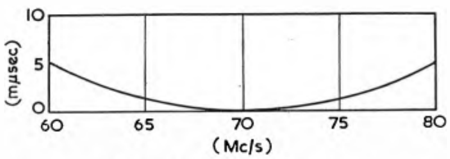

Fig. 4. Group delay variation

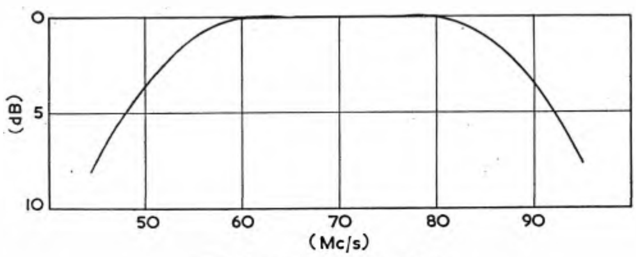

Fig. 5. Amplitude variation

the master drawing is altered as required, and new prints made. Where alterations are of a minor nature the glass negatives are altered, to avoid re-photographing.

Eventually the desired performance is achieved from a circuit without any alterations, and the negative from which this is produced is used as the master for production purposes. Protective finishes are applied to the satisfactory boards and final tests on valve replacement, component tolerance, etc., are made with the print in its final form.

Production Drawings

The number of drawings required for a printed circuit unit is considerably smaller than the number required for a conventionally wired unit. Complete manufacturing details of the circuit board assembly are contained on one drawing and a component schedule, together with the appropriate master negatives. Part of the circuit board drawing is made photographically by means of the film positives produced in the photographic process. Test specifications are compiled, and these, again, are simplified, as no circuit alignment is involved and a limited number of tests will suffice to indicate a correctly functioning assembly.

Brief Details of Some Other Units Successfully Printed

Besides the i.f. amplifier described above, several other units have been successfully designed with printed circuits. Among these are the following:

Demodulator for Multi-channnel Telephony or Television Signals

This panel includes limiters, discriminator, phase equalizer and base-band amplifier. Some details of its performance may be of interest and are given below.

Mid-band frequency	70Mc/s
'Flat' bandwidth	18Mc/s
Input signal level	350mV, r.m.s.
Input impedance	75Ω
Equalized group-delay variation	2mμsec, 64Mc/s to 76Mc/s
Deviation of input carrier	880kc/s r.m.s.
Base-band response	within 0·1dB 25c/s to 7Mc/s
Base-band output signal level	−13dBm (−26dBm per channel)
Noise power ratio (intermodulation + basic noise) with random selection of valves having average g_m	= 16·5mA/V

Frequency (kc/s)	Ratio (dB)	Equivalent Distortion Level (dB)
62	58	−72
1022	58	−72
2540	58	−72

Effect of valve changes on noise power ratio.—Limiter and amplifier valves replaced with valves having average g_m = 19·0mA/V (col. 2) and 14·0mA/V (col. 3), no adjustment.

Frequency (kc/s)	19·0mA/V (dB)	14·0mA/V (dB)
62	57	55
1022	58	57
2540	58	57

Effect of replacement of discriminator drive valve with valves having g_m = 19·0mA/V and 14·0mA/V. Without adjustment of discriminator linearity control (cols. 2 and 3) and with readjustment (cols. 4 and 5).

Frequency (kc/s	No adjustment		Readjusted	
	g_m=19·0 (dB)	g_m=14·0 (dB)	g_m=19·0 (dB)	g_m=14·0 (dB)
62	57	52	57	58
1022	58	53	58	57
2540	58	53	58	58

Effect of Temperature on Noise Power Ratio

Frequency (kc/s)	20°C (dB)	50°C (dB)
62	58	55
1022	58	55
2540	58	55

These changes are due to change in characteristics of the germanium diodes used in the limiter circuits.

The only adjustment provided in this unit is the linearity control which may be adjusted for maximum flatness of discriminator derivative using display methods, or alternatively for minimum intermodulation noise in a reference channel. Readjustment on changing valves is only required when the lowest possible intermodulation distortion is required.

A Low Noise Pre-amplifier for Use in Systems Using Travelling Wave Tubes

Brief details are as follows:

Midband frequency	70Mc/s
'Flat' bandwidth (within $\pm$0·1dB)	20Mc/s
Gain	20dB
Group delay variation 64Mc/s to 76Mc/s	1mμsec
Noise figure	9·0dB

Valves may be changed without adverse effect on performance.

Base-band Amplifier for Telephony or Television

This is a three-stage feedback amplifier having characteristics as follows:

Frequency response within 0·1dB 25c/s to 7Mc/s
Intermodulation distortion not worse than −80dB

Phase Equalizers

A series of fixed equalizers covering a range of values and suitable for equalization of group-delay/frequency variation in i.f. equipment has been made. These are available as plug-in units.

Supervisory Equipment

Various units of supervisory equipment such as pilot monitors and oscillators have been produced.

Proposed Maintenance Procedures

As a result of the initial requirement that no selection of valves shall be necessary in production, circuits which permit the changing of valves have been developed. In consequence of this valves can be changed in the field without affecting the performance of the equipment. The same considerations ensure satisfactory operation with valve ageing. Periodical testing can be applied to reveal valves which are reaching the end of useful life. This requires little in the way of test equipment, facilities for monitoring anode currents being built into the equipment. Apart from valve failures the only other faults which can affect performance are component or board failures. Such faults are readily revealed by abnormal valve currents or levels of signal at appropriate test points. Location of a board so affected does not, therefore, involve expensive test equipment. The recommended procedure to deal with a faulty board is replacement by a new board. Replacement is a simple matter and does not require a high degree of skill. Wherever possible the cost of a board is kept low by suitable unit breakdown, and it is normally economical to discard a faulty board. In this way test equipment requirements are kept to a minimum while equipment performance remains equal to new equipment. It would be possible, however, to repair faulty boards. This would involve location of the faulty component, replacement of the component with an exact equivalent, repair of the protective film on the board, refitting of the board and performance testing. To enable this to be done a skilled technician would be required, equipped with test apparatus suitable for overall performance testing. A repair kit would be required for the resin coating, with suitable curing facilities. Such arrangements could no doubt be provided by a large system operator at a central maintenance depot, but it is doubtful if the cost of such a scheme would compare favourably with the replacement scheme suggested. In the equipment at present in use, board replacement costs have been kept to a level of £10 or below and failures of boards in working equipment have been virtually nil to date. It is suggested that a system operator should be provided with spares of each board used. In a normal system each spare board would cover several units and the cost would be small. It is envisaged that a maintenance engineer would be able to visit an outlying station with virtually a replacement 'station' in a suitcase. With such a scheme a small system operator can enjoy the same standard of performance as a large operator at an economical cost.

Summary of Advantages of Printed Circuits

The advantages of the technique are conveniently dealt with in the following sections.

Improvement in Design

Improvements in design result from the fact that the designer is no longer restricted to circuits which are simple to align and test. Providing the circuits are suitable for printing it is no more difficult to print a highly complex circuit configuration than to print a simple one. Due to the particular physical form of the circuit improved stability, screening, etc., are achieved, the circuits tending towards two dimensional layout rather than three dimensional, with consequent reduction of unwanted coupling fields. Elimination of variations in wiring and component lead lengths results in better reproduction of the initial design, less allowance being required for production variations.

The final performance depends, as it should, upon the designer, and not upon the skill of test department personnel, circuit alignment after assembly being eliminated.

Simplification of Manufacture

The production of the printed boards is a straightforward procedure, ideally suited to large or small scale production. The facilities required are not extensive. Assembly of boards is a simple operation and can be assisted by semi-automatic machinery. As can be seen from the photographs, unit assembly and wiring is very much simplified.

Reduction of Manufacturing Cost

Reductions in cost result from the simplifications outlined above which reduce assembly time. In addition there is a great reduction in testing time. In the case of the amplifier described a testing time of about five minutes for the printed circuit version replaces a period of hours for alignment and testing of the conventional wiring version.

Reduction of Space and Weight

Printed circuit panels are ideally suited to compact module-type assemblies, giving reductions in space and weight. Even in the example dealt with, where the printed amplifier was designed as a unit replacement, a noticeable reduction in weight is achieved due to elimination of conventional coils, cableforms, etc.

Ease of Maintenance and Reduction of Maintenance Costs

As described previously, maintenance procedure is very much simplified and test equipment requirements are greatly reduced. This results in savings from the reduction in test equipment outlay and enables the necessary maintenance work to be carried out by a less skilled class of technician. The resultant economies in salaries and training costs can be considerable.

Acknowledgment

The author is indebted to the Engineer in Chief, Marconi's Wireless Telegraph Co. Ltd, for permission to publish this article.

Avalanche Transistors

An Appraisal of Their Properties and Uses

By R. C. V. Macario*, B.Sc., Ph.D.

The results from an investigation of the properties of some experimental alloyed junction avalanche transistors enables a review of their characteristics to be made. The characteristics are clearly related to the properties of the semiconductor junctions and this is advantageous when designing a device to a given specification. Design data are given and applications summarized.

(Voir page 312 pour le résumé en français: Zusammenfassung in deutscher Sprache Seite 317)

A NUMBER of papers have recently appeared describing various aspects of the properties and applications of alloyed junction avalanche transistors[1,2,3,4,5]. The device has many interesting characteristics and demonstrates in a novel way many of the properties of semiconductor pn junctions. The results of an experimental investigation of a number of samples show that the characteristics of the device are closely related to the properties of the junction suggesting that it should be possible to specify the characteristics in manufacture.

The most important of these characteristics are the voltage range and the speed of response. The maximum voltage excursions are somewhat limited, but an appraisal of their structural design suggests that extremely fast waveforms may be generated, and examples of this have already been described[6].

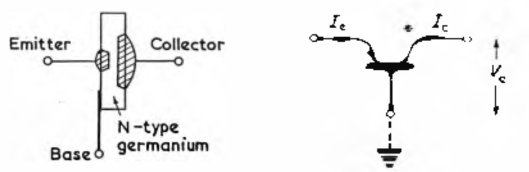

Fig. 1. Avalanche transistor structure and notation

Naturally, knowledge of the terminal and response characteristics of the device under transient signal conditions define the various circuit applications and examples of circuits which illustrate these conditions are given. The data given refer to room temperature, but the effect of temperature on the operation of the device is also described.

Attention is confined to germanium pnp structures, although in general the same considerations apply to the complementary structure and silicon devices, but the more important differences are pointed out.

Distinguishing Properties

The avalanche transistor[2] is a two-junction device which utilizes the avalanche multiplication mechanism of carriers crossing a reverse biased pn junction, with an emitting junction controlling the carriers being multiplied. The structure usually takes the same form as the r.f. alloy transistor as illustrated in Fig. 1 where a possible circuit symbol and associated notation is also shown. The emitter junction is usually made smaller than the (multiplying) collector junction in order to obtain a reasonable current transfer ratio.

At low voltages many of the characteristics are indistinguishable from those of a normal transistor, but as the collector voltage, and hence the multiplication factor M (= number of additional carriers of the same type generated for each carrier entering the depletion layer) is raised, they deviate markedly. The avalanche mechanism ultimately leads to junction breakdown, when $M = \infty$, and to use the device most effectively, the voltage characteristic must be limited by this breakdown voltage V_B rather than a Zener or punch-through breakdown. Limitation by the punch-through voltage V_P is possible[3,5], but only devices limited by body breakdown are considered.

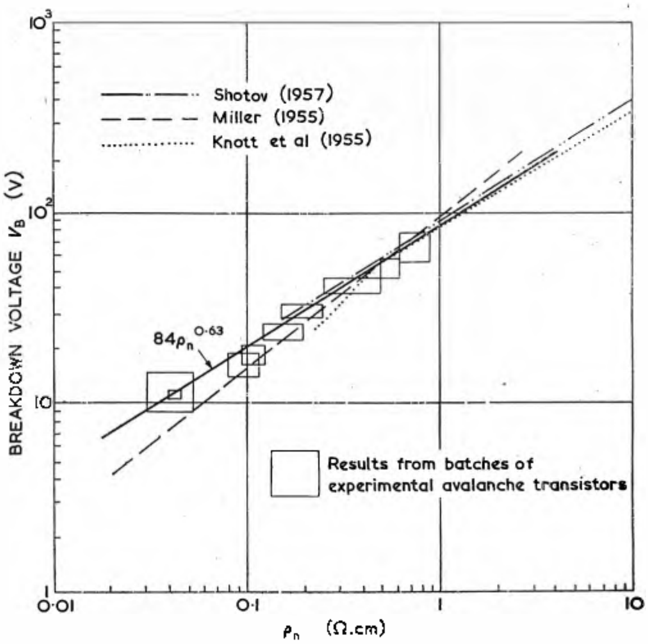

Fig. 2. Variation of junction breakdown voltage with base resistivity

The breakdown voltage of a reverse biased alloyed junction is mainly controlled by the resistivity ρ_n of the base material, and Fig. 2 collects together the available data[7,8,9] relating V_B to ρ_n (N-type base material). Thus avalanche multiplication, rather than a field effect which would obey a $V_B \propto \rho_n$ law, accounts for the breakdown voltage V_B, from approximately 10V to 50V where surface effects become important. A design equation is:

$$V_B = 84\rho_n^{0.63} \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (1)$$

Avalanche transistors therefore have base resistivities which are rather lower than usual, ρ_n usually being in the range from 0·05 to 0·3Ωcm.

Between 0 and V_B, the multiplication factor M increases uniformly from 1 to ∞ with increasing collector voltage V_c according to the well-known relation[9]

$$M = [1 - (V_c/V_B)^n]^{-1} \quad \ldots\ldots\ldots\ldots \quad (2)$$

n is a parameter which varies with resistivity and type of

* *Plessey Company Ltd., formerly A.E.I. Research Laboratories.*

base material[2]; for germanium pnp transistors $n = 3$. For germanium npn transistors n ranges from 4 to 6, while for silicon devices n is about 2.

With no emitter current, only the junction leakage current I_{co} is enhanced, becoming very large on approaching breakdown.

With an emitter current I_e, the collector current becomes $I_c = \alpha_0 . I_e . M$, and so alpha is a function of V_c, i.e.

$$\alpha_{ce} = M\alpha_0 \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (3)$$

Similarly

$$\alpha_{cb} = M\alpha_0/(1 - M\alpha_0)$$

where α_0 is the low voltage or normal alpha when $M = 1$. Thus the emitter-to-collector current gain α_{ce} may take all values from $\alpha_{ce} < 1$ to $\alpha_{ce} > 1$ being equal to unity when $M = 1/\alpha_0$, according to the value of the collector voltage V_c. The base-to-collector current gain α_{cb} will increase rapidly, pass through infinity, and decrease to unity as $V_c \rightarrow V_B$. There are thus two important collector voltages.

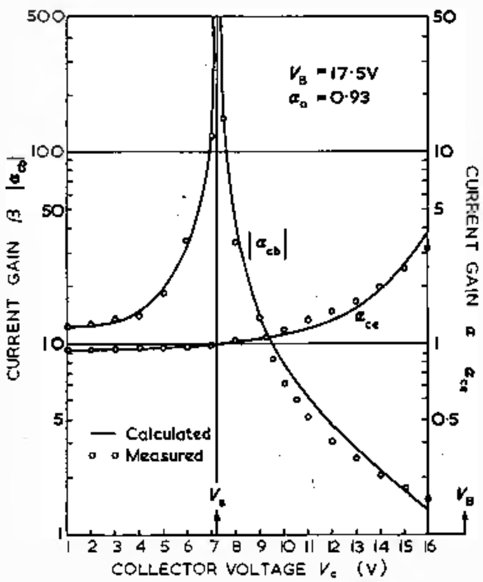

Fig. 3. Variation of current gain with collector voltage

(1) The collector-to-base breakdown, V_B, at which α_{ce} tends to ∞ and $\alpha_{cb} = 1$, i.e. collector current flows with no emitter current.

(2) The collector-to-emitter breakdown voltage, V_s, commonly called the sustain voltage[2], at which α_{cb} tends to infinity and $\alpha_{ce} = 1$, i.e. collector current flows with no base current.

V_s depends on V_B and the low voltage α_0 according to the relation[2]:

$$V_s = V_B . (1 - \alpha_0)^{1/n} \quad \ldots\ldots\ldots\ldots\ldots \quad (4)$$

V_s is therefore smaller, for a given V_B, the smaller the value of n and the larger the value of α_0. Fig. 3 shows that these equations in a practical transistor are obeyed quite closely.

V_s is a very significant voltage of the avalanche transistor. The base current changes sign at this voltage, and it defines the lower working voltage of the transistor as a bistable element. Fig. 4 illustrates the common emitter characteristics and the significance of V_s.

The distinctive property of the device is that in the region of collector voltage between V_B and V_s the transistor possesses a current gain greater than unity, and therefore is able to switch regeneratively between these two voltages.

For instance, any circuit which enables V_B to be approached before emitter current flows will switch to the lower voltage V_s should any emitter current begin to flow. One such circuit is the two terminal switch arrangement, Fig. 5, described by Miller and Ebers[2]. The voltage drop in R_b due to the eventual collector-to-base breakdown current causes emitter current to flow and the transistor relaxes to the lower voltage V_s. Very stable characteristics are observed, the precise negative resistance slope being determined by the value of R_b. By extending the discussion of

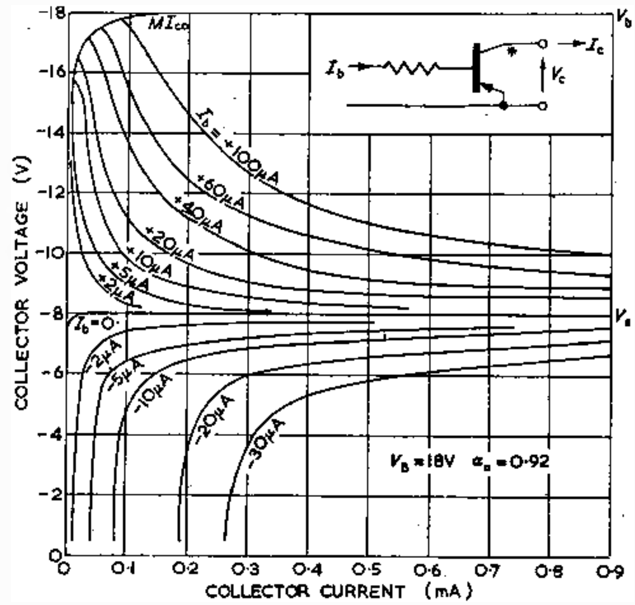

Fig. 4. Common emitter characteristics

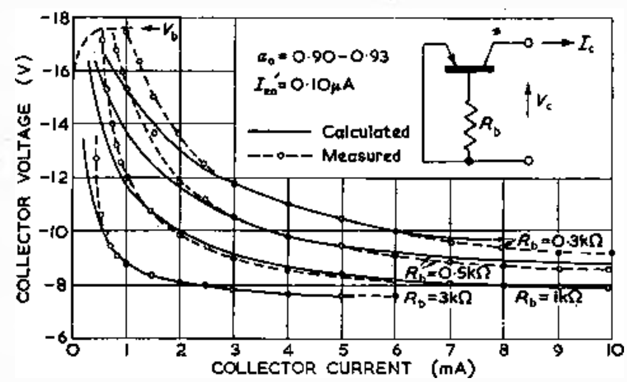

Fig. 5. Two terminal avalanche transistor switch

Miller and Ebers the characteristics can be calculated for different values of R_b when V_B, I_{eo}' and $\alpha(I_e)$ of the transistor are known. The comparison with the measured characteristics is shown in Fig. 5.

The characteristics of this circuit are a good guide to the properties of the device, and can be very easily displayed on a cathode-ray oscilloscope. The voltage ratio, V_B to V_s, is usually between 2 and 3 (depending on α_0) and this low ratio may have disadvantages, but the negative resistance characteristic is well defined and the 'off' impedance high.

Fig. 5 essentially represents the input characteristic, measured at the two terminals of the circuit shown. Other characteristics can be observed at other terminals or by changing the impedance within the circuit. For instance, adding an emitter resistance R_e to Fig. 5 will cause the characteristic to revert to a positive slope ($=R_e$) as V_c reaches V_s. Probably a better known characteristic is that of

Fig. 6 referring to the emitter; it is very similar to that of a point contact transistor. Similar considerations apply to the base input characteristics, but in this case the characteristic is short-circuit stable. The curves can be estimated and applied to switching circuits similar to those described by Logue[10].

Probably the best-known extension of the basic circuit of Fig. 5, is the relaxation oscillator configuration[5,6] of Fig. 7, where the collector current is supplied periodically by a

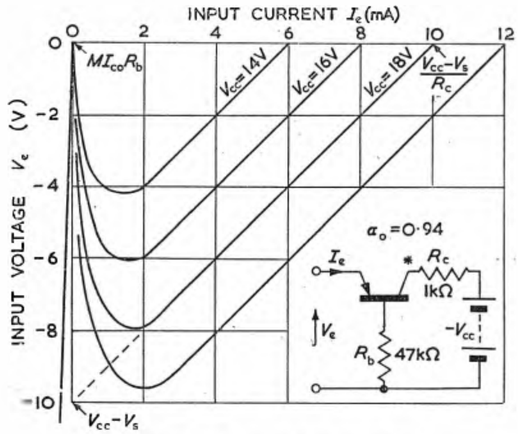

Fig. 6. Emitter input characteristics

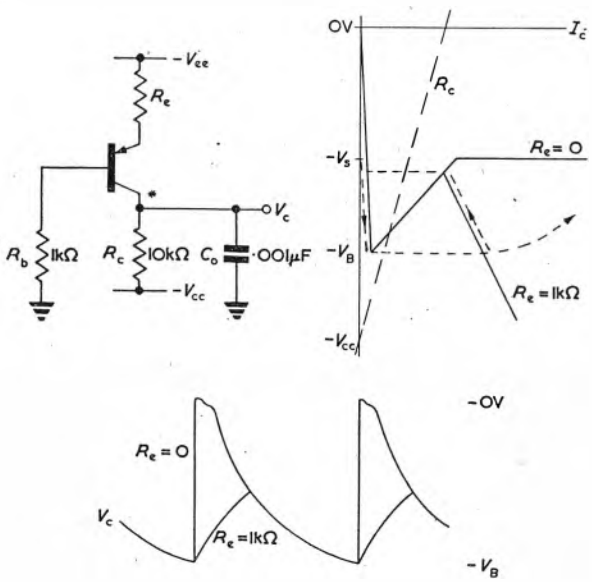

Fig. 7. Avalanche transistor relaxation oscillator circuit and waveforms

capacitance C_o. The transistor switches 'on' when V_o reaches V_B as R_c charges C_o. Two versions of the circuit are possible according to whether an external emitter resistance is added or not. Distinct waveforms are generated in the two cases, and this introduces a very interesting aspect of the avalanche transistor, namely, a carrier storage phenomena in the base of the transistor which enhances the switching action.

With an emitter resistance R_e the circuit operates as would be expected by superimposing the load line R_c on the d.c. input characteristic as illustrated in the diagram. The initial current which charges C_o as the transistor turns 'on' is in this case limited by R_e, but when $R_e = 0$, there is nothing to prevent a very large current charging C_o. The collector waveform therefore rises very sharply and moreover, though V_c should limit at V_s, it is found in fact that sufficient charge flows into C_o to take the collector voltage to zero. This has been shown to be due to the presence of carriers which are stored in the base as the very large initial current flows across the transistor[3,5]. The actual collector waveform edge, however, can be extremely fast. Rise-time between 1 and 100mμsec are typical.

Attention has been drawn to the two distinct modes of operation of the avalanche transistor as the applications fall into two such classes:

(1) Low-speed mode—as a bistable element.

(2) High-speed mode—as a pulse regenerative amplifier. Equivalent circuits for both modes can be constructed.

Other Effects of the Low-Resistivity Base Material

The lower than usual base material resistivity quite naturally affects other parameters of the transistor. Some of the changes are interesting and so are briefly noted, although, in the main, they are of little importance as far as applications are concerned.

Low Voltage Current Gain α_o

In general α_o is somewhat lower than usual because of the reduced emitter efficiency. This could be a disadvantage in bistable circuits, but if the design of the transistor is optimized[11] quite high alphas result even for avalanche transistors with very low base resistivities.

Collector Capacitance C_o

As the depletion layer in the more heavily doped material is more restricted a larger capacitance results for a given junction voltage. The usual relation

$$C = 1{\cdot}4 \times 10^4 (\rho_n \,.\, V_c)^{-1/2} \text{pF/cm}^2 \quad \ldots\ldots\ldots \quad (5)$$

is found to be obeyed for practical transistors. The capacitance is therefore quite large and tends to deteriorate the trailing edge of waveforms generated by avalanche transistors.

Leakage Current I_{co}

I_{co} proves to be of the order of 0·1μA at room temperature. Because I_{co} is so low, the true breakdown voltage is in general attainable in the transistors as expected from theory, and the full range of the characteristics utilized. A very low leakage current also suggests the transistor may have a very low noise figure of merit when used in a linear low voltage mode.

Base Spreading Resistance $r_{bb'}$

Measurements suggested that $r_{bb'}$ for the transistors examined could be conveniently represented by the relation

$$r_{bb'} \simeq 75 \,.\, \rho_n \text{ ohms} \quad \ldots\ldots\ldots\ldots \quad (6)$$

Thus, for avalanche transistors, this impedance is usually negligible compared to the external circuit resistances.

Small Signal Parameters

A comparison of the small signal hybrid h-parameters indicate the main changes.

Transistor	ρ_n(Ω/cm)	h_{11}(Ω)	h_{12}	h_{21}	h_{22}(mhos)
Avalanche	0·1	27	0·0002	0·965	4
r.f.	1·0	28	0·0006	0·98	0·6

The voltage feedback factor h_{12} is expected to be about one-third smaller due to its direct dependence on the depletion layer. On the other hand the collector conductance h_{22} is worse due to the marked dependence of α_{ce} on V_c.

An approximate expression for h_{22} is

$$h_{22} = 3 \,.\, M^2 \,.\, \alpha_o \,.\, I_e \,.\, V_c^2 / V_B^3 \quad \ldots\ldots\ldots \quad (7)$$

which indicates a rather poor collector impedance. However, in switching applications I_e is very small in the 'off' condition.

Equivalent Circuit Representation

LOW-SPEEDS

A simplified equivalent circuit[10] can be used as suggested in Fig. 8. The current which flows with the transistor 'off' is almost negligible; in the 'on' condition the current is limited externally, though a voltage V_s must be maintained across the transistor.

CONDITION	TYPICAL VALUE				
	r_e	r_b	r_c	I_c	I_b
EMITTER OFF	10MΩ	15Ω	10MΩ	$\simeq 0$	$\simeq 0$
EMITTER ON	$(25/I_e\Omega)$	15Ω	50kΩ	I_e	0

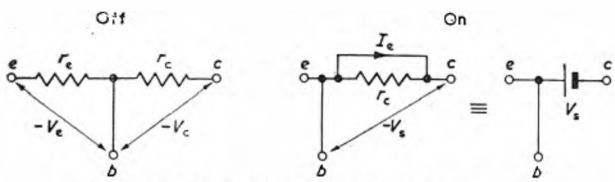

Fig. 8. Low-speed equivalent circuits

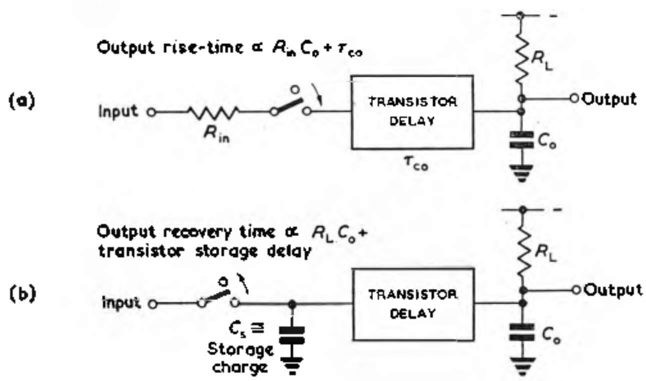

Fig. 9. Factors affecting the transient response
(a) Turn-on (b) Turn-off

HIGH-SPEEDS

In general the output waveform will be taken from the collector and the factors limiting the speed of the waveform are represented diagrammatically in Fig. 9.

In the case of low speed circuits, the time-constant R_eC_o dominates, although the effect of carrier storage (C_s) in the base of the transistor may tend to give a spike on the leading edge of some waveforms. However, in high-speed circuits this distortion is a very useful attribute as it extends the voltage range, cf. Fig. 7.

As the required speed of the waveform is raised, the internal delay and resistance of the transistor finally limit the leading edge of the waveform. The trailing edge is in general controlled by the external circuit. Thus Fig. 10(c) shows the effect of the emitter resistance R_e on the collector waveform of the circuit of Fig. 7 or Fig. 10(a). The emitter waveform is also shown and gives information about the current flow.

When V_c reaches V_B, in effect the emitter is short-circuited to the collector and $C_o + C_c$ would charge to zero volts via R_e, but for the presence of the transistor delay and internal impedance. The transistor is therefore conveniently represented as shown in Fig. 10(b), by r_s, L_b and C_c. As the transistor switches 'on,' the emitter voltage waveform rises at an initial rate $R_e \cdot V_B/L_B$ towards the value $V_B \cdot R_e/(R_e + r_s)$. The collector waveform rises at almost the same rate. If $R_e + r_s$ is very small the forward current is very large, of the order of one ampere, due to the enhancement by the avalanche mechanism and many carriers are stored in the base of the transistor[3,5]; it is this charge which is sufficient to charge $C_o + C_c$ to zero volts as indicated. The collector remains at zero volts until emitter current ceases to flow; this period depends on the initial current defined by $(R_e + r_s)(C_o + C_c)$, but can quite easily last a microsecond or so.

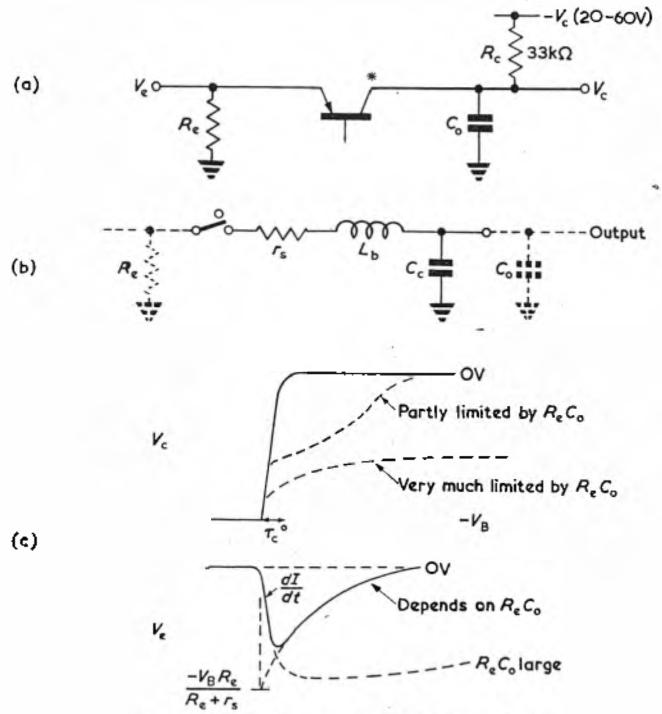

Fig. 10. Avalanche transistor switching action

If R_e is not small the initial current is limited and insufficient charge is accumulated in the transistor, consequently inhibiting the collector waveform. Eventually R_e may prevent the carrier storage effect altogether and V_c does not rise past V_s. The emitter waveform retains the initial sharp edge, but decays at a correspondingly decreasing rate. The effect of C_o on the operation is not so marked. A minimum value exists[5], however, partly determined by its ratio to C_c because sufficient charge has to be accumulated in the transistor prior to V_o reaching V_s for the collector to bottom. On the other hand, if C_o is made too large, the r_sC_o product may deteriorate the waveform. Practical values of C_o range from a few tens of picofarads to tenths of microfarads. Recently a detailed theory of the transients build-up in avalanche transistors has been published[14].

The rise-time of the collector waveform when only limited by the transistor delay is denoted by τ_c^o—this is related to the transistor base width, and is represented in the equivalent circuit by L_b.

The peak current as the transistor turns 'on' is limited by r_s and for some alloyed junction avalanche transistors this was observed to be about 2 to 3Ω.

ULTIMATE DESIGN CONSIDERATIONS

The speed at which an avalanche transistor will switch 'on' is related to the dispersion time taken by the carriers leaving the emitter to reach the edge of the collector depletion layer, but since the depletion layer retracts as the collector waveform rises, the total rise-time depends more on the actual transistor base width W. Actual observations of τ_c^o plotted against W calculated from observations of f_a suggest that τ_c^o is proportional to W^2 and this leads to an interesting design feature.

The optimum choice of base width is such that the transistor is just about to punch-through as the collector breaks down[5], or $V_B = V_P$; this gives an expression for the optimum choice of W versus base resistivity ρ_B. Expressions for the maximum alpha cut-off frequency $f_{a(max)}$ and the minimum response time $\tau_c^o{}_{(min)}$ can then be calculated[12]

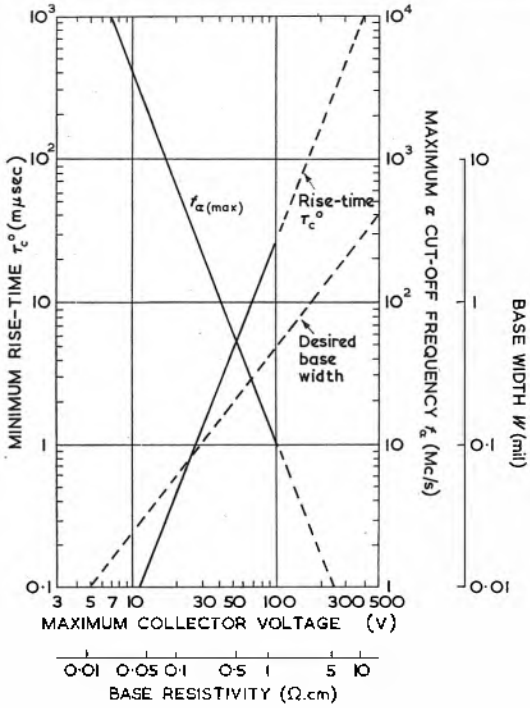

Fig. 11. Design data for avalanche transistor switch

as functions of the maximum collector voltage, now equal to V_B assuming that $\tau_c^o = W^2/D_P$ where D_P is the diffusion constant for holes in the base. These data are given in Fig. 11 and it seems possible that devices can be made with considerably higher speed of operation. It should be pointed out, however, that the rate of rise of the collector waveform will initially be faster than as suggested in Fig. 11 because the base width is effectively less.

Circuit Applications

Low-Speeds

Fig. 12 shows a bistable circuit configuration and the waveforms generated. The amplitudes can be calculated using the equivalent circuit of Fig. 8. It should be noted that the amplitudes obtained are the same whether the circuit is bistable, monostable or astable. The design of the circuits is similar to those with point contact transistors[10,13]. All the circuits have the disadvantage of small voltage outputs, combined with large ' on ' power dissipations. (Silicon devices would be better in this respect). Also, there is a spike on the leading edge of the collector waveform (due to carrier storage), while the trailing edge is rather sluggish.

These factors make it difficult to design avalanche transistor circuits which drive corresponding circuits, for instance, a single transistor binary stage.

On the other hand a twin avalanche transistor binary can be designed to operate satisfactorily, and to accommodate spreads in V_B and α_0 or V_s. Fig. 13 shows such a circuit together with a section of a ring counter arrangement that will operate in conjunction with the binary unit. A scale-of-ten can therefore be constructed using seven transistors. The

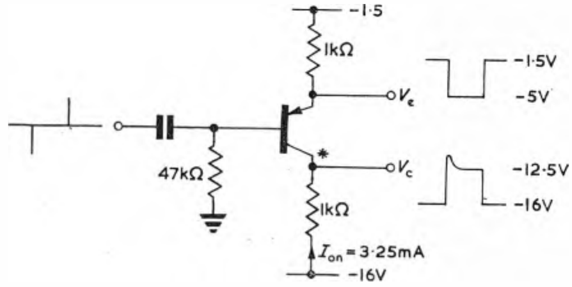

Fig. 12. Bistable circuit
(Transistor : $V_B > 16V$ $V_s = 8V$)

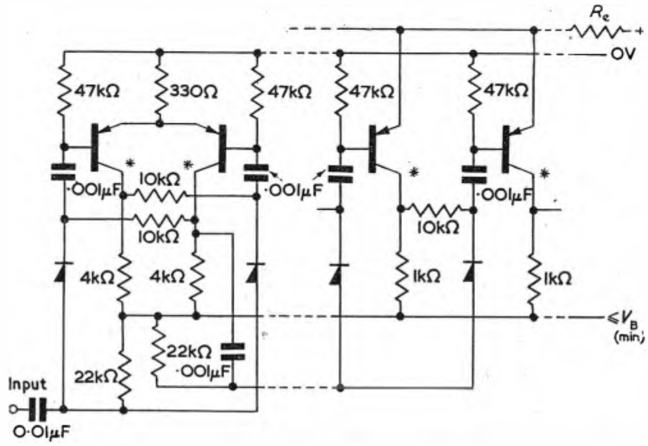

Fig. 13. Avalanche transistor binary stage and ring counter

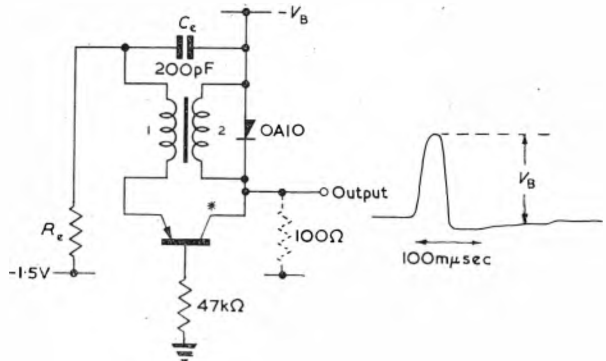

Fig. 14. Avalanche transistor blocking oscillator

tolerance on the transistors or the voltage supplies is not critical and the units operate up to a rate approaching 100kc/s.

High-Speeds

Here the application is to pulse regenerative amplifiers since very little power is required to trigger the transistor 'on' while the avalanche transistor will generate a very large current pulse very rapidly. The orders of voltage swing and rise-times involved are summarized in Fig. 11. The design of practical circuits, based on Fig. 7, have been described by Beale et al[5], and Chaplin and Owens[6] describe a method of generating extremely short waveforms.

The very fast collector waveform edge can also be utilized in a blocking oscillator arrangement as shown in Fig. 14. The waveform is determined by the inductance and so makes the waveform substantially independent of the transistor. The amplitude is V_B volts. The waveform repetition rate is mainly limited by transistor dissipation, which can be reduced to some extent by making C_e smaller since this capacitance effectively supplies the charge generating the collector waveform.

Effects of Temperature

Observations of V_B over a wide range of temperatures for a range of samples indicates that V_B obeys the law

$$V_B{}^T = V_B{}^\circ [1 + \gamma . T] \qquad (8)$$

over the range $-200°C$ to $+75°C$, where $V_B{}^\circ$ is the breakdown voltage at 0°C, T is the temperature in °C, and the temperature coefficient γ obeys the relation

$$\gamma = 0{\cdot}12 \, . \, (\rho_n{}^{0\cdot 70}/V_B{}^\circ) \mathrm{V}/°\mathrm{C} \qquad (9)$$

This last relation is particularly interesting as it suggests that γ is almost independent of ρ_n, because of equation (1), which is not an unexpected result since the breakdown phenomena is an effect which takes place almost exclusively within the depletion layer. V_B therefore increases with temperature; however, the effect is partially offset by the increase of leakage current I_{co} with temperature which increases at the expected rate.

The other significant parameter is α_0, which tends to increase with temperature. Thus V_s decreases, and so the net effect is for the bistable properties of the transistor to improve as the temperature is raised.

Conclusion

A review of the factors which are most likely to enter into the design and application of avalanche transistors has been attempted. In the main, the device is capable of generating extremely fast waveforms over a large voltage range, but there is a carrier storage effect. It appears possible, moreover, that the device can be designed to meet specific applications.

Acknowledgments

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Radar for Small Craft

A new low-power radar equipment—the 'Consort'—has been designed by the Marconi International Marine Communication Co. Ltd. for use in small vessels where space and electrical power are limited. The new equipment is suitable for tugs and harbour craft of all types, fishing and pilot vessels, lifeboats and yachts; and its introduction opens up a new field of application for marine radar equipment.

The complete equipment comprises a display unit, transceiver unit and aerial unit; the use of transistors, printed circuits and new manufacturing techniques having resulted in an extremely compact installation. The 'Consort' operates from a 24V battery or by conversion from other voltages, and has a maximum range of 14 miles —sufficient to be of real value under fog conditions, and to provide navigational assistance ample for the needs of small craft.

The display unit is designed for mounting on a deckhead, bulkhead or table. A 5in cathode-ray tube is employed, and a magnifying lens incorporated in the detachable visor increases the effective diameter of the display to 8in.

Two minutes after switching on, the 'Consort' reaches a 'stand-by' condition, and it is then ready for immediate use. A 'press-to-view' switch is provided on the display unit, and when this is depressed the scanner starts and the equipment becomes fully operational for two minutes. It will then revert to the 'standby' condition until the 'press-to-view' switch is again depressed; but this switch can be overridden to provide continuous viewing if conditions require it. This arrangement ensures instant readiness without waste, battery consumption being reduced from 10A when the equipment is operating to 4A in the 'stand-by' condition.

The tubular case of the display unit is sealed with rubber gaskets to give a strong and watertight assembly, and hand grips are fitted on the mounting bracket to enable the observer to steady himself when using the equipment in bad weather. The whole assembly measures 14in × 11in × 25in long including the detachable visor, and weighs only 14lb. Bearings are shown at 5° intervals on a graticule engraved on the Perspex screen; and range is measured by range rings equally spaced on the face of the display.

The transceiver is housed in a steel cabinet measuring 26in × 16in × 10in and weighing 62lb. It can be fitted in the smallest wheelhouse, and installation costs have been kept to a minimum by the use of a made-up cableform 10ft long, which is attached to the transceiver and plugs into the display unit. A panel on the front of the transceiver carries the range selector switch, a brilliance control for the range rings, and gain, sea clutter and rain clutter controls. The power supply is controlled by a mains switch; and focus, brilliance and fine tuning controls—which are seldom used—are mounted on the rear of the display unit. Five ranges are provided, covering 0·6, 1·5, 4·0, 8·0 and 14·0 miles.

The normal supply voltage is 24V d.c. provided by batteries with or without float charge. A stabilized power unit in the transceiver makes fluctuations in supply voltage from 20·5 to 32V permissible.

The scanner assembly incorporates a 3ft slotted waveguide aerial in a waterproof fibreglass radome. It is designed for mounting directly on the roof of a small wheelhouse; but if extra height is needed to clear obstructions, a light alloy tripod mast can be supplied.

A Rapid Response Recording Cardiotachometer

By A. W. Melville*, A.M.I.E.E., and J. B. Cornwall*, Grad. I.E.E.

The instrument described has been designed for use under theatre conditions. It incorporates excellent discrimination against interference, an effective pulse amplitude control circuit and a rate recorder with alternative time-constants, one of which provides a rapid response suitable for the evaluation of fast acting stimuli.

(Voir page 313 pour le résumé en français; Zusammenfassung in deutscher Sprache Seite 318)

A NUMBER of instruments for recording the human pulse rate have been described in the literature. The one to be described is the outcome of several years' experience with an earlier model and incorporates several important improvements. Our particular requirement has been for an instrument that will produce a satisfactory continuous record of pulse rate under normal theatre conditions, on a patient who may be conscious or anaesthetized and who, in the latter case, may be undergoing major thoracic or abdominal surgery. This calls for the maximum possible freedom from artifacts produced by voluntary and involuntary movements of the subject, or by electrical equipment in the theatre. There should also be rapid response to rate changes, so that the short-term effect of fast-acting drugs may be observed accurately and the onset of a crisis be apparent immediately.

The recorder should operate with a reasonable variation of input pulse amplitude without readjustment. Several methods of pulse detection are available:

(a) Heart or arterial pulse sounds may be used, but the difficulty in eliminating interference and airborne noise and various friction noises makes this method unattractive.

(b) The blood pressure pulse may be utilized in two ways. If a vessel is already cannulated for recording blood pressure with an electromanometer, electrical impulses from the apparatus may be used directly for counting. This method restricts the general use of the instrument but has been used by Wyatt[1].

In an alternative method the pressure pulse may be detected as an opacity pulse in accessible tissue, the most suitable being the pinna of the ear. This method has been used by Bekkering and Van Dal[2] and is advantageous when the subject is required to carry out controlled exercise during the recording period.

(c) The heart action potentials may be amplified directly. In spite of the apparent difficulty in discriminating between several phases of the complex electrocardiac cycle and against other muscle potentials, this method has been retained because of the simplicity of connexion to the subject.

Description of Equipment

The complete instrument can be considered in two main sections:

(a) An amplifier with negligible response at the mains frequency and the maximum possible differentiation between the *R* wave of the electrocardiac complex and all other electrical signals present and providing a constant amplitude pulse to the counter.

(b) A stable, fast response rate recording circuit.

The Amplifier (Fig. 1)

The authors' experience has shown that even when an efficient mains frequency rejection filter is used, the discrimination obtained from a differential input amplifier justifies the additional complication. In the present equipment a two stage differential pre-amplifier employing 6SL7 twin triodes, provides a rejection ratio of 10 000 to 1. The first stage is preceded by a three section low-pass filter which eliminates interference arising from r.f. diathermy in adjacent rooms. Interference from diathermy in use on the patient is not entirely eliminated, but amplifier blocking is prevented.

The second differential stage is followed by a twin-*T* rejection filter, which terminates in the gain control. The filter is tuned to the mains frequency. Stage three is a straight high-gain amplifier and stage four is tuned to 8·5c/s by a twin-*T* network in the negative feedback loop. The effective *Q* of this stage is controlled by a resistor in series with the twin-*T* to prevent excessive 'ringing'. Boyd and Edie[3] have shown that while the spectrum of the e.c.g. rises steeply in amplitude down to below 5c/s that of muscle potentials falls at an increasing rate below 20c/s. The choice of pass-band is necessarily a compromise and for some years a median frequency of 8·5c/s has been used successfully. The overall frequency response is shown in Fig. 2. A diode clipper at the grid of stage four removes unwanted negative going signals and results in a significant reduction of interference. Stage four is followed by a pulse amplitude control circuit. The output of V_4 is fed to an attenuator consisting of a Philips thermistor type B832015P/330K with a nominal resistance of 330kΩ at 65°F, and a 10kΩ resistor. The output of V_{7a} is rectified by V_6 and the smoothed d.c. control signal is fed to the control valve V_{7b}. The insulated heater of the thermistor Th_1 carries the cathode current of V_{7b}. An increase in signal amplitude at V_{7a} anode reduces the current through the thermistor control heater with a consequent increase in thermistor resistance. A second similar control circuit is used at the input to the gating stage V_5, and the cathode circuit of V_{7b} also includes a Philips thermistor type B832003P/2K2 as a further control element. This control circuit will maintain the output pulse amplitude with ±5 per cent even when the input increases or decreases by a factor of 3.

The gating stage V_5 has two available recovery periods. These periods correspond to maximum pulse rates of 110 and 240 per minute which ensures that no signal is admitted to the relay circuit, V_8, for a period of 1/110 or 1/240min after the reception of a pulse.

The period corresponding to the lower rate limit will be used whenever possible so that the maximum protection

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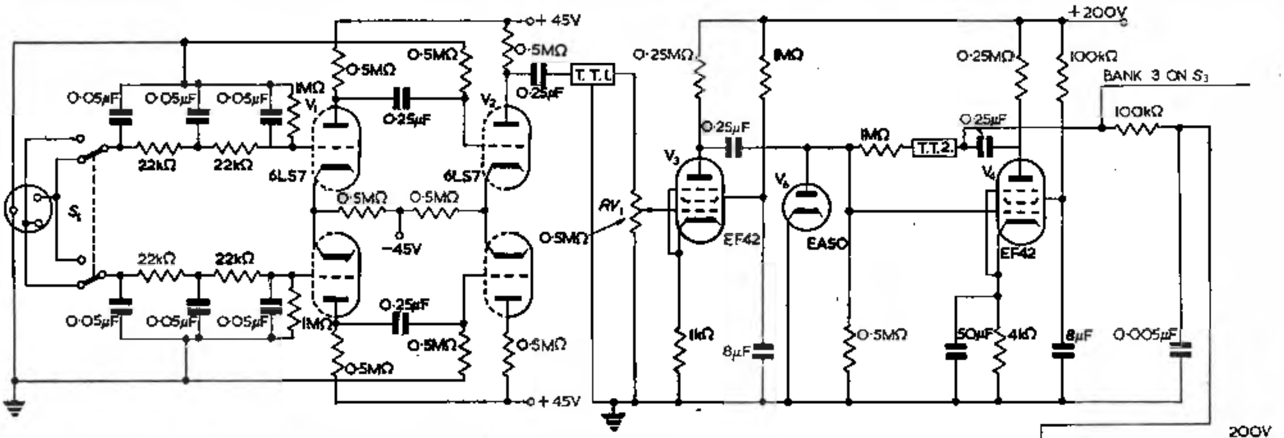
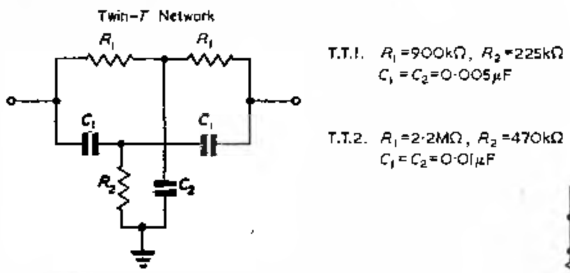




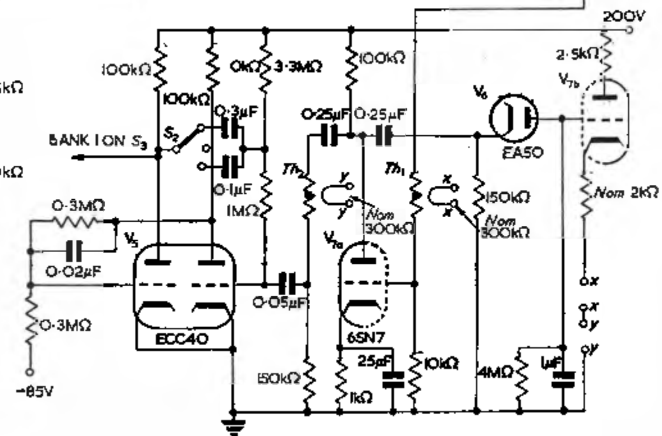


Fig. 1. Amplifier and gating circuits

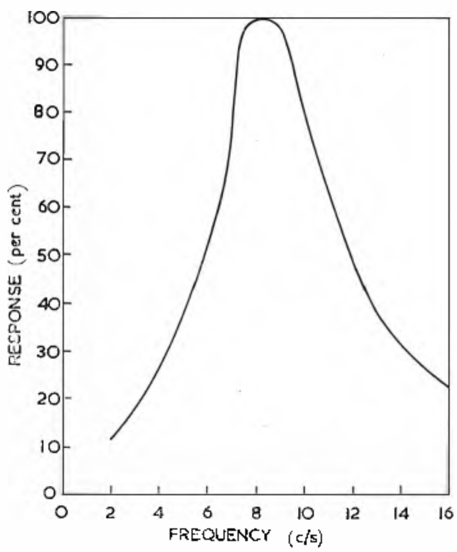


Fig. 2. Frequency response of amplifier

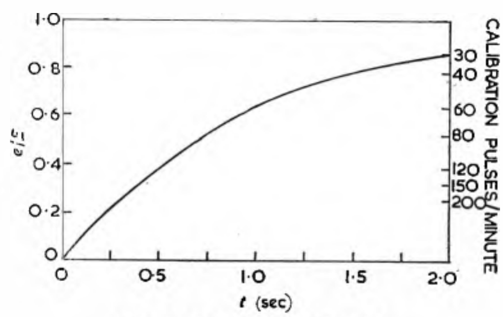


Fig. 3. Plot of scale calibration

against interference signals is obtained from the gate since its effectiveness approaches 100 per cent as the pulse period approaches the gate recovery time. The possibility of developing a circuit in which the gating period is controlled by the pulse rate is most attractive, but has not yet been found feasible. The square wave output from the gating circuit is differentiated to provide a constant-amplitude short duration pulse.

The Rate Recorder

This is an adaptation of a circuit developed by the authors for use in a drip rate recorder for intravenous solutions[4]. If a capacitor is allowed to charge through a resistor from a constant voltage source for the period between consecutive pulses and the final charge reached during each period recorded, the record can be calibrated in pulse rate.

In such a charging circuit the instantaneous voltage e across the capacitor C, when expressed as a fraction of the applied voltage E, is given by:

$$e/E = 1 - e^{-t/RC}$$

and Fig. 3 illustrates the plot of e/E for t=2sec, R=1MΩ and $C = 1\mu$F. The projection on to the voltage co-ordinate of points on this curve corresponding to the periods of pulse rates from 40 to 200 per minute illustrates the relationship between e/E and pulse rate when this system is used. This calibration is seen to be the superimposition of a reciprocal and an exponential curve, resulting in a scale which, while not linear, is quite satisfactory for the purpose.

The basic switching circuit is shown in Fig. 4.

The reception of a pulse operates the relay contacts shown as S_1, S_2, and S_3 in the following sequence.

S_1 opens, S_2 closes, S_2 opens, S_3 closes, S_3 opens, S_1 closes.

Thus the charge received by C_1 during the period between pulses is distributed between C_1 and C_2 by S_2 when the next pulse is received, the source of voltage being removed at S_1. The parallel connexion is then broken at S_3 so that

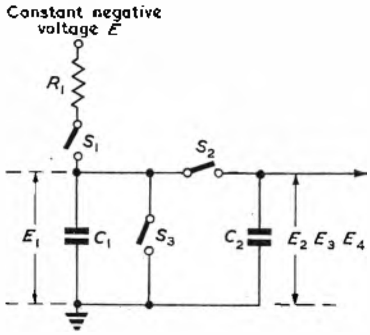


Fig. 4. The basic switching circuit

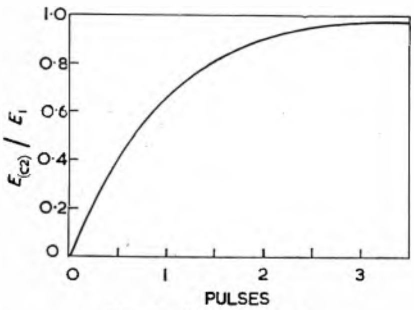


Fig. 5. Response time

C_1 may be discharged at S_3 and then reconnected to the voltage source at S_1. C_2 retains its charge between pulses, and the voltage on C_2 is recorded by a simple valve-voltmeter.

If the initial charge on C_2 is zero, and the relay operation is at a constant rate the conditions are:

Charge given to C_1 between pulses $= C_1E_1$.

When this charge is distributed $C_1E_1 = (C_1+C_2)\,E_2$

$$\text{or } E_2 = \frac{C_1E_1}{C_1 + C_2}$$

where E_1 is the voltage on C_1 initially and E_2 is the voltage on C_1 and C_2 in parallel and is the voltage retained by C_2.

Similarly when the second pulse is received the total charge $= C_1E_1 + C_2E_2$ and when this is distributed

$$C_1E_1 + C_2E_2 = E_3\,(C_1 + C_2)$$

$$\text{or } E_3 = \frac{C_1E_1 + C_2E_2}{C_1 + C_2}$$

where E_3 is the voltage retained by C_2 after the reception of the second pulse.

If $C_1 = 1\mu$F and $C_2 = 0{\cdot}5\mu$F and E_2, E_3, E_4 .... are plotted against constant rate pulses with the initial voltage on C_2 zero, the curve shown in Fig. 5 is produced.

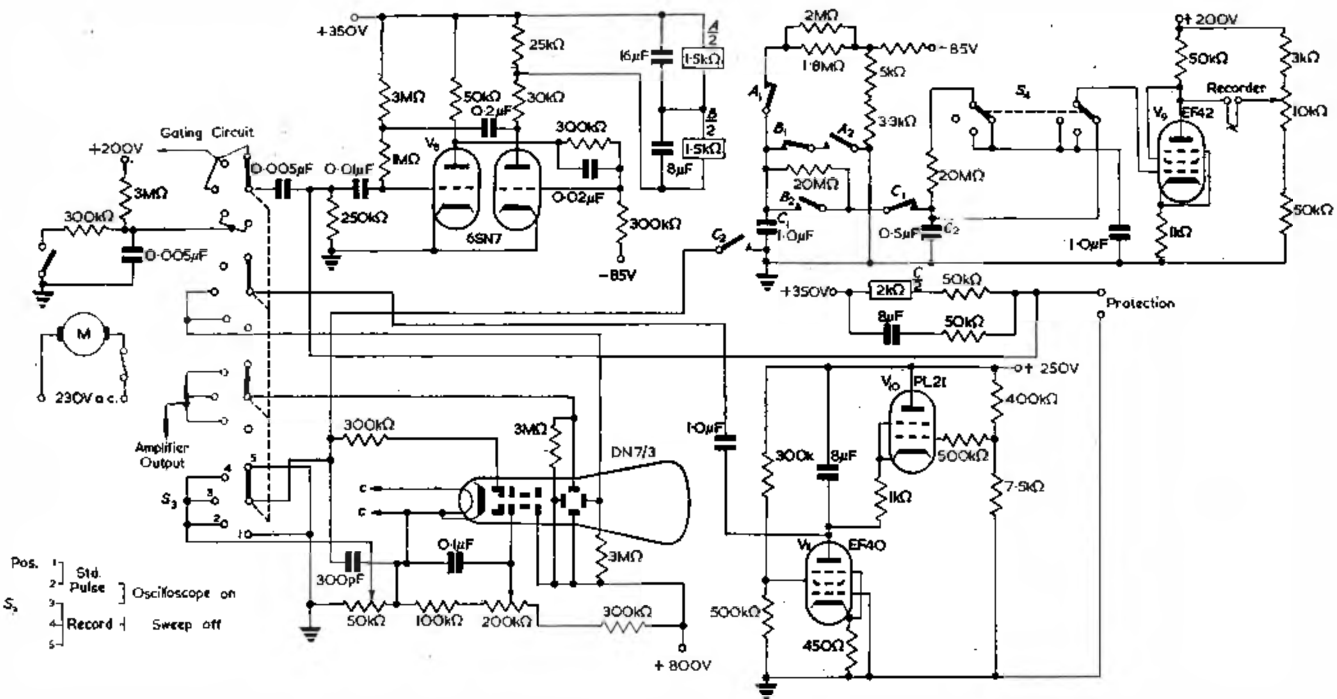
From this it is seen that the voltage on C_2 will have reached 0·96 of its final value when three pulses have been received, and this is true regardless of the actual pulse rate, the magnitude of the change and whether it is an increase or decrease.

In the complete circuit, Fig. 6, the switching cycle is performed by a pair of relays in series with second anode of a triggered relay valve V_8. As previously mentioned, the trigger voltage is the differentiated output of the gating circuit. The electrolytic capacitors shunting the relay coils increase the energizing times, reduce audible noise and also modify the de-energizing time to provide the correct switching sequence.

A simple valve-voltmeter V_9 will drive a 0 to 1mA recording milliameter, the one used in this case being an Esterline-Angus with a coil resistance of 1·2kΩ. A zero set control is provided and a change of the valve V_9 does not effect the calibration providing the zero is set accurately. The scale is calibrated from the curve shown in Fig. 3. A calibration checking pulse is provided by a synchronous clock motor with a final shaft speed of 24rev/min and fitted with a four lift cam operating a pair of contacts, which provide pulses at a repetition rate of 96 per minute.

The rapid response time indicated by Fig. 5 is particularly useful when effects of fast acting stimuli are to be recorded.

Fig. 6. Oscilloscope and counter circuit



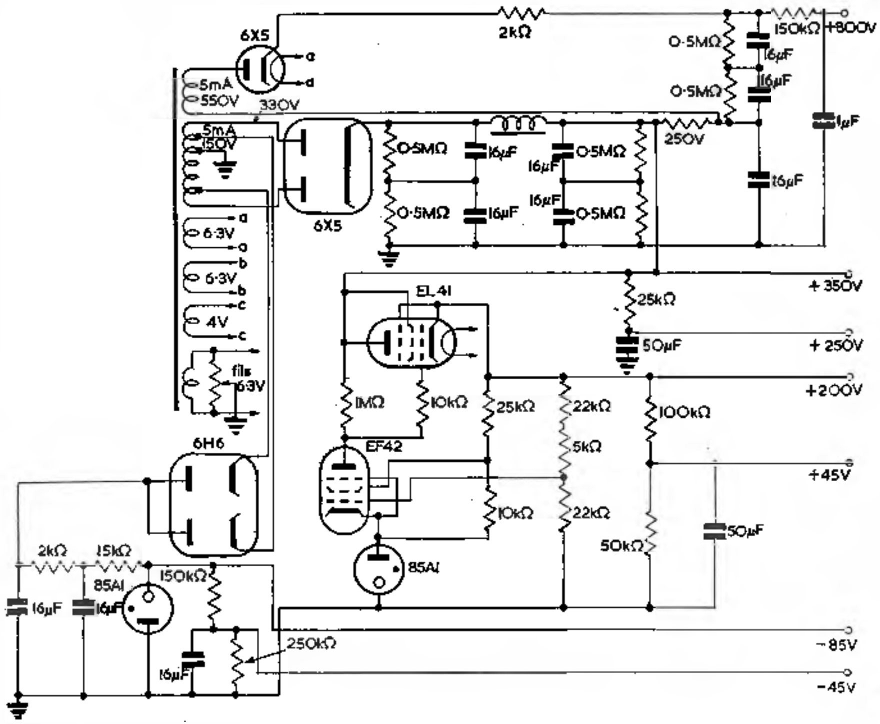


Fig. 7. Power supplies

Setting up Adjustments

To facilitate the setting up, a c.r.t. type DG7/3 is included. A linear time-base with a sweep time of 4sec may be switched in so that the polarity and waveform of the pulse may be examined. In most cases the electrocardiographic lead 2, connected between right arm and left leg will provide an adequate pulse. In difficult cases, the three conventional leads may be connected and the best combination chosen by observation on the c.r.t.

In cases of thoracic and abdominal surgery, artifacts are usually considerably reduced if an earthed electrode plate is located beneath the operating region, and this may also act as the indeterminate electrode for the diathermy equipment. Fig. 8 shows a simultaneous record of the unfiltered electrocardiograph on the top channel and the waveform at the anode of V_5 on the bottom channel, during a period of "full knee bend' exercises. One error has occurred at the point indicated by the arrow, when the gating circuit has been returned to its initial state prematurely by an intensive muscle artifact. Fig. 9 illustrates on channel 2 the waveform at the anode of V_{7a} when the e.c.g. as recorded on channel 1 had a superimposed line frequency interference signal.

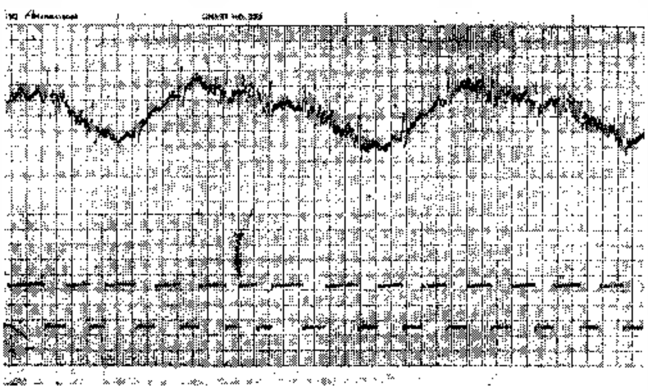


Fig. 8. Illustration of muscle potential elimination

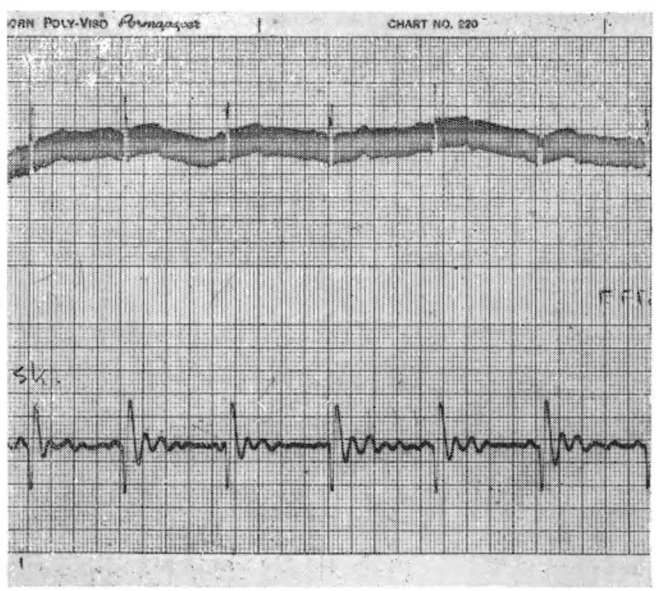


Fig. 9. Illustration of 50c/s interference elimination

There are circumstances when a longer time-constant is an advantage. Switch 4 provides for the addition of a time-constant of 20sec to the counting circuit and this arrangement is used when recording long-term variations and particularly when there are occasional unavoidable artifacts.

To ensure that the record is not distorted by transients from a diathermy machine which may be connected directly to the subject, a relay is installed in the diathermy machine itself. A normally open contact is connected to the terminals marked 'diathermy' so that when the machine is used, these terminals are closed. This circuit is extended to the cardiotachometer where simultaneously it earths the signal line between V_5 and V_8 and operates a relay which disconnects the rate-counting circuit, and leaves the recorder indicating the rate at the instant of interruption. The de-energizing of this relay is delayed for approximately 3sec so that the rate counting circuit is once more operating normally before the recorder is switched back into circuit.

The circuit of the power supplies is given in Fig. 7.

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Transistors and Cores in Counting Circuits

By F. Rozner*, B.Sc.(Eng.), and P. Pengelly†

The square loop magnetic core and the transistor are now well accepted circuit elements in digital computers. In this article it is shown how the two elements can be combined to form non-critical reliable counting circuits capable of handling input frequencies up to 750kc/s. A modification to allow the circuit to be used as a shift register is also described.

(Voir page 313 pour le résumé en français: Zusammenfassung in deutscher Sprache Seite 318)

THE square loop magnetic core is now a well established memory element in digital computers. As it possesses many advantages over other switching elements, efforts have been made to extend its applications to other kinds of circuits, e.g., gates, counters and shift registers. Several different types of shift registers have been described and one such circuit, using one core per bit, is shown in Fig. 1(a).

The operation of the circuit can be briefly explained as follows. Assume that the polarity of the shift pulse is such that it changes the state of '1' cores to '0'. Its effect upon '0' cores is negligible. If a core contains '1' then, during switching '1' → '0', it acts as a transformer. The voltage appearing across the winding W_2 charges the capacitor and drives a current i_c through R and W_3. Now i_c acts in opposition to the shift pulse current i_s and since the latter is dominant it cancels out the effect of i_c. The windings are here assumed to have equal number of turns. When the shift pulse terminates, i_c (now supplied from C) switches the following core '0' → '1'. The overall effect of one shift pulse is to shift the pattern of digits one place to the right.

Three points can be made in connexion with this circuit: (a) all losses are supplied by the shift pulse, (b) the current charging the capacitor flows effectively in parallel with the core magnetizing current, and (c) power is lost in R. The power carried by the shift pulses is considerable and it may become an embarrassment if a transistorized source of shift pulses is used. To reduce the demand upon such a source the circuit in Fig. 1(a) is modified as shown in Fig. 1(b). Here, R is replaced by a diode which is non-conducting during the shift pulse. When the latter terminates, the full voltage from C is available to switch the following core '0' → '1'.

It is a simple matter to arrange a shift register to act as a counter[2]. Both circuits in Fig. 1 can be converted by connecting the last stage to the first and then arranging that all but one core are in the '0' state. Each shift pulse will then shift the '1' around the ring.

Core-Transistor Circuits

As already mentioned all the losses in the circuits in Fig. 1 are supplied by the shift pulses. Apart from the strenuous requirements upon the driving source there is the difficulty of supplying many stages whose power demand varies. It is desirable, therefore, to limit the function of the shift pulses to triggering and to supply the energy losses from elsewhere. These considerations have prompted the search for ways to incorporate a power amplifier in the basic circuit, transistors being particularly suitable for this purpose.

Since the shift pulses no longer supply the circuit losses, counting by means of shift registers connected in a ring ceases to be economical. Great saving in components and power consumption is achieved if the counter consists of a number of cascaded bistable circuits. It is to fill the need for a reliable core-transistor binary that the circuit described below was developed.

Fig. 2(a) illustrates the use of a capacitor in series with the winding instead of in parallel as in Fig. 1. Assume that the core is in state '0' and C is discharged. A positive rectangular pulse applied as shown will switch the core '0' → '1'. If the width of the pulse is sufficient C will charge up to the peak voltage. When the pulse terminates C discharges via the winding and R, switching the core back '1' → '0'.

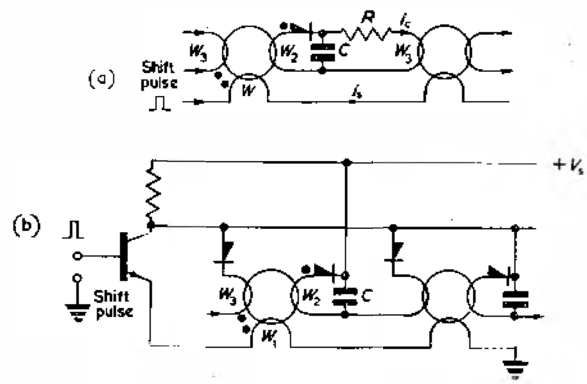

Fig. 1. Shift registers using one core per bit
(a) Delayed transfer
(b) Gated transfer

The core passes, therefore, through the complete hysteresis cycle every time a pulse is applied. Two strokes can be distinguished, viz. '0' → '1' and '1' → '0'. In order to make it a scale-of-two circuit it is necessary to limit the switching to one stroke per pulse. The circuit in Fig. 2(b) shows the way it is achieved. It represents a three-stage counter, with one core and one transistor per stage. There is, in addition, one transistor, X_0, driving the chain.

Let core 1 be in state '0'. A positive pulse applied to winding W_{11} switches the core '0' → '1' and fully charges C_{11}. The voltage across W_{12} brings X_1 into conduction. Due to regeneration the collector current i_c builds up very quickly and a positive pulse is applied to W_{21} charging C_{21}. When the collector voltages of X_0 and X_1 begin to fall, C_{11} and C_{21} discharge via W_{11} and W_{13} respectively. The directions of the discharge currents are such that the magnetic fields in core 1 cancel each other out and the core remains in state '1'. Consequently, the next pulse applied to W_{11}, while charging C_{11}, does not switch the core. X_1 remains non-conductive, and on termination of the input pulse the discharge current of C_{11} switches the core '1' → '0' without the presence of any inhibiting action.

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† *Ferguson Radio Corporation Ltd.*

Thus the necessary condition for bistability, viz, one switching stroke per input pulse, is satisfied.

When core I switches '0' → '1' the applied field is quite large resulting in a switching time which is short compared with the duration of the input pulse. The voltage pulse appearing across W_{12} is correspondingly narrow and it must be broadened to ensure that the output pulse of X_1 is sufficiently broad to charge C_{21} fully. It is desirable in fact, to maintain similar driving conditions all down the counting chain. The combination $R_{12}C_{12}$ acts as a simple pulse stretcher. The presence of these components obviously slows down the rise and fall of the pulse. This effect is negligible on the front edge when the presence of regeneration ensures a very rapid build up of current in X_1. It is more significant on the back edge which occurs without the core switching back, i.e. without regeneration.

analysis of the circuit in Fig. 2(b) is both difficult and unprofitable. Useful design formulae can be established by making some approximations. Although errors are introduced the circuit is sufficiently non-critical in operation not to be significantly affected by them.

The switching time of the core is governed by the law

$$\tau_m^{-1} = KN(I_m - I_c) \quad \ldots\ldots\ldots\ldots\ldots \quad (1)$$

where K is a constant dependent on core material and dimensions,

N is the number of turns in the winding,

I_m is the amplitude of the current pulse,

I_c is the current which would cause the core to switch in infinite time and is related to the knee of the curve.

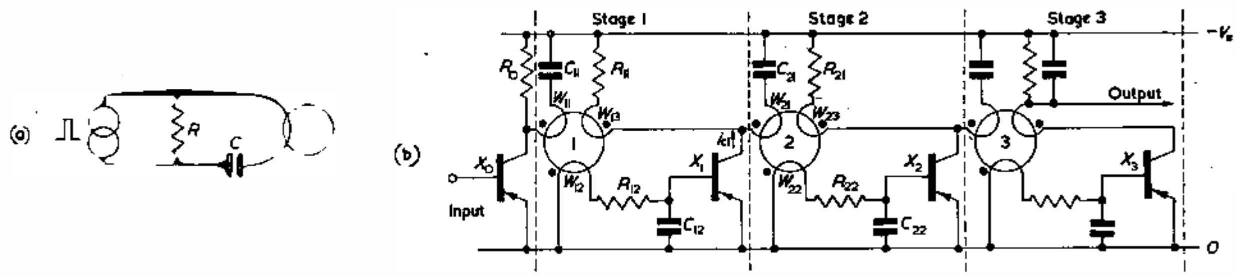

Fig. 2. Core-transistor scale-of-two circuits
(a) Switching the core with C in series
(b) Scale-of-eight counter using series C

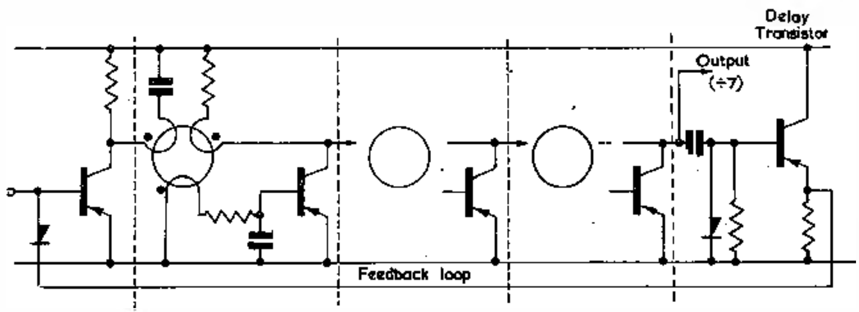

Fig. 3. Scale-of-seven showing feedback arrangement

The filter $R_{12}C_{12}$ has an additional function. The hysteresis loop of the core is not ideally rectangular. Consequently, a small voltage spike appears accross W_{12} when an input pulse is applied to W_{11} even if the core does not switch. The energy content of this spike is small, however, and the filter effectively prevents the appearance of spurious pulses in the output of X_1.

When the count of an n stage circuit is to differ from 2^n it is a standard practice to feed back some pulses and so modify the count. The feedback pulses must be suitably delayed so that they do not coincide with the proper input pulses. Fig. 3 shows an arrangement in which a transistor X_4 is used to delay the feedback pulses. An output is obtained from the last stage immediately after all cores switch '0' → '1'. The feedback pulse suitably delayed is applied to the selected core so as to switch it '1' → '0', and thus modify the count. The count of the circuit in Fig. 3 is 7.

Design Considerations

By virtue of the non-linear behaviour of the core accurate

It can be deduced from this law that the switching time varies inversely with the excess amplitude of the current pulse and that there is a threshold current below which the core will not switch.

If in the circuit in Fig. 2(b) the collector current of X_0 rises very rapidly and its maximum amplitude is sufficient, then the core will switch '0' → '1' in a time which is short and nearly independent of C_{11} and R_0. This is, in fact, what happens. When the core switches '1' → '0' the time taken is much longer and it depends upon C_{11} and R_0. It may form the main limitation of the counting rate and must be considered in some detail.

Fig. 4 shows four current curves: (a) represents the inverse exponential discharge of C_{11} which would take place if X_0 (Fig. 2(b)) cut off instantaneously and the core did not switch. Curve (b) allows for the finite decay time of X_0 collector current. Curve (c) shows the discharge curve when core 1 switches '1' → '0'. Curve (d) is the approximation to (c) by a rectangular current pulse. The areas under all four curves are the same and equal to the maximum charge stored in C_{11}, viz. $C_{11}V_s$, where V_s is the supply voltage.

From curve (d) the mean switching current I_m and the overall switching time τ_m can be defined, such that

$$I_m = (V_s C_{11}/\tau_m) \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (2)$$

τ_m will normally be greater than the core switching time τ_c, but for maximum speed the two values should be very close.

R_o can be calculated with the help of the equivalent circuit in Fig. 4 which is valid for the discharge of C_{11} when the core is not switching, i.e., curve (b) in Fig. 3(a). The generator current is assumed to decay linearly according to $i_c = (V_s/R_o)\ (1 - (t/\tau_1))$, where τ_1 is the decay time.

The loop equation is,

$$i_c R_o + i_m R_o = V_s - (1/C_{11}) \int_0^t i_m dt$$

Solving for i_m,

$$i_m = (V_s C_{11}/\tau_1)\ \left\{ 1 - \exp\left(-(t/C_{11}R_o)\right) \right\} \ldots\ldots \quad (3)$$

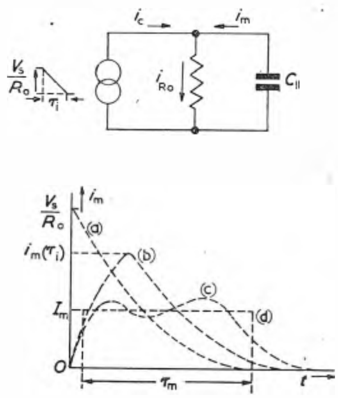

Fig. 4. Discharge curves for the series C

(a) Idealized discharge curve
(b) Actual discharge curve with the core not switching
(c) Actual discharge curve with the core switching
(d) Approximation to curve (c)

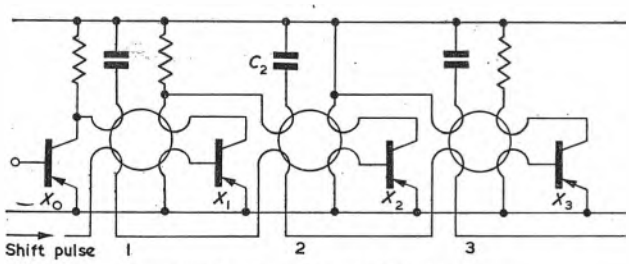

Fig. 5. Three-stage shift register

Maximum i_m in the circuit will flow when $t = \tau_1$. Expanding the exponential term up to the second power and putting $t = \tau_1$,

$$i_m(\tau_1) = (V_s/R_o)\left(1 - \frac{\tau_1}{2C_{11}R_o}\right) \ldots\ldots\ldots \quad (4)$$

This equation is valid only for $\tau_1 < C_{11}R_o$.

$i_m\tau_1$ denotes the peak current in curve (b), Fig. 4.

If curve (b) is assumed to be a triangle with base τ_m, then the area under (b) $= \frac{1}{2}\, i_m(\tau_1)\tau_m =$ area under (d) $= I_m\tau_m$, i.e. $i_m(\tau_1) = 2I_m$.

Solving the above equation for R_o and substituting for $i_m(\tau_1)$ yields,

$$R_o = \frac{V_s}{2C_{11}I_m}\left[C_{11} \pm \sqrt{(C_{11}^2 - 4C_{11}I_m\tau_1 V_s^{-1})}\right] \ldots \quad (5)$$

With the aid of equations (1), (2) and (5) the circuit can now be designed.

The following data apply to the circuit in Fig. 2(b).

Cores: Mullard FX1502, Ferroxcube, 2mm o.d., 50 turns per winding.

Transistors: Mullard OC44, average $\beta = 50$, average $f_\alpha = 10$Mc/s.

Supply: $V_s = 10$V.

The information contained in equation (1) is more often supplied as a curve. For the cores used the value 760mA-turns, giving τ_m of about 2μsec was selected. Hence, substituting in equation (2),

$$C_{11} = (I_m\tau_m/V_s) = 0{\cdot}76/50 \times (2/10)\mu\text{F} = {\cdot}003\mu\text{F}.$$

The decay of the input current pulse was measured to be about 0·5μsec. Substituting in equation (5), gives $R_o = 166\Omega$, but in fact, the value is not critical, and both $R_o = 180\Omega$ and $R_o = 150\Omega$ are suitable.

Although the components evaluated above refer to the first stage of the counter, all stages can be identical. Thus, in Fig. 2(b), $R_o = R_{11} = R_{21}$, etc., $C_{11} = C_{21} = C_{31}$, etc., and so on.

Other components are non-critical, $R_{12} = 180\Omega$ and $C_{12} = 0{\cdot}003\mu$F give satisfactory results. Similarly, provided that the transistors have a $\beta > 40$ and $f_\alpha > 7$Mc/s their characteristics do not matter.

The maximum input repetition frequency which the circuit can handle depends on the switching time and on the permissible collector dissipation of X_o and possibly X_1. As mentioned earlier the circuit was designed to have '1' → '0' switching time of 2μsec. However, the '0' → '1' time is much shorter, 1μsec being sufficient. It follows that the circuit will count 1μsec pulses at 330kc/s repetition rate. In accordance with equation (1) the switching time can be cut by increasing the current. To achieve this result both C_{11} and R_o must be reduced as shown by equations (2) and (5). In fact, the highest input frequency achieved was 750kc/s with $C_{11} = 0{\cdot}001\mu$F, $R_o = 120\Omega$ and $V_s = 12$V.

Shift Register

The circuit described so far can be modified to act as a shift register. For this purpose it is necessary to add an extra winding for the shift pulse and to remove the inhibitory action which occurs as a result of the capacitor discharge current flowing through the winding of the preceding stage. Fig. 5 shows a three-stage shift register. The direction of the windings is such that the positive shift pulse switches the core '1' → '0' and so does each transistor. The capacitor discharge current switches the core '0' → '1'.

Assume that the contents of the register is 1, 0, 1, in cores 1, 2 and 3, respectively. The incoming shift pulse will have no effect upon core 2, but it will switch cores 1 and 3. The shift power required to do so is quite small because the switching is aided by the regeneration in transistors X_1 and X_3. Pulses are emitted by X_1 and X_3 but none come from X_2. Consequently only C_2 is charged. The state of the cores is now 0, 0, 0, with C_2 fully charged. Upon the termination of the shift pulse C_2 discharges switching core 2 '0' → '1'. Hence the state of the register is now 0 1 0, i.e., the pattern of digits has been shifted one stage to the right.

Acknowledgment

The authors are indebted to the Directors of Ferguson Radio Corporation Ltd for permission to publish this article.

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The Charge Storage in a Junction Transistor during Turn-off in the Active Region

By R. S. C. Cobbold*

Through the solution of the one dimensional diffusion equation for a junction transistor, equations are derived for the emitter and collector currents that exist under conditions of minimum turn-off time. These equations are shown to be in fairly good agreement with the practical results.

(Voir page 313 pour le résumé en français: Zusammenfassung in deutscher Sprache Seite 318)

CONSIDER a pnp transistor that is conducting in the active region. If a reverse base current is applied, the collector current will fall in a manner that is dependent on the magnitude of this reverse current. Three particular cases that arise can be considered.

Suppose that the magnitude of the reverse base current is less than the emitter current that passes in the steady state. Then holes will continue to be injected into the base region after the reverse base current is applied, which results in a comparatively long turn-off time.

When the reverse base current is equal to the emitter current then the net emitter current will be zero, and only the collector will act as a sink for holes.

If the reverse base current is just greater than the emitter current, then both emitter and collector will act as sinks, and the turn-off time is shortened.

The limiting value of turn-off time will be obtained when the reverse base current is sufficiently great so that the emitter current is limited only by the diffusion time of holes in the base region. Under these conditions there are perfect hole sinks at both the emitter and collector junctions, and the turn-off time will be independent of variations in the reverse base current.

It is this limiting case which is the subject of this article. This case is particularly interesting since, under these conditions, the minimum turn-off time of a junction transistor is obtained.

Formulation of Problem

To solve the problem, recourse must be made to the physical phenomena involved, rather than to the equivalent circuits, which apply only for times greater than several transistor time-constants.

Assume that the emitter and collector can both act as perfect hole sinks, and that the effects of collector capacitance and base width modulation can be neglected.

If the short-circuit current gain in the grounded base connexion, α_0, tends to unity, then the initial hole density in the base will be a linear function of distance; hence one can write

$$\Delta P = mx + q$$

where ΔP = excess minority carrier density

x = distance from emitter

m and q are constants determined by the base width $2W$, the diffusion constant D_p for holes in the base region, and I_{c1} the initial collector current.

The one-dimensional diffusion equation for holes in the base can be written as

$$\partial\Delta P/\partial t = -(\Delta P/\tau_p) + D_p(\partial^2\Delta P/\partial x^2) \quad \ldots\ldots \quad (1)$$

where τ_p = minority carrier life-time in the base region.

Initially, for

$$0 < x < 2W,$$
$$\Delta P = mx + q$$

Also, at the boundaries,

$$\Delta P = 0 \text{ at } x = 0 \text{ and } x = 2W$$

Solution

It is shown in the Appendix that, subject to the initial conditions, the solution to equation (1) can be written in the form

$$\Delta P = \sum_{n=1,3,5\ldots}^{\infty} 4/\pi \left[\frac{q + Wm}{n} \sin(n\pi x/2W)\right] \exp\left[-(t/\tau_p)\left(1 + \frac{n^2\pi^2 L_p^2}{4W^2}\right)\right] - \sum_{n=2,4,6\ldots}^{\infty} 4/\pi \left[\frac{Wm}{n} \sin(n\pi x/2W)\right] \exp\left[-(t/\tau_p)\left(1 + \frac{n^2\pi^2 L_p^2}{4W^2}\right)\right]$$

where $L_p^2 = D_p\tau_p$

In a junction transistor the base width is usually very much less than the diffusion length L_p, thus

$$\frac{n^2\pi^2 L_p^2}{4W} \gg 1$$

Now in the active region, $m = -(q/2W)$, so that equation (1) can be rewritten in the form

$$\Delta P = -\sum_{n=1,2\ldots}^{\infty} 4/\pi \left[\frac{Wm}{n} \sin(n\pi x/2W)\right] \exp\left[-(t/\tau_p)\frac{n^2\pi^2 L_p^2}{4W^2}\right] \quad \ldots\ldots\ldots\ldots \quad (2)$$

Equation (2) was computed for various values of $(t/\tau_p \,.\, \pi^2 L_p^2/4W^2)$ and the results are shown in Fig. 1. The y axis is $\Delta P \,.\, \pi/4Wm$, and the x axis is the distance from the emitter. The quantity $t/\tau_p \,.\, \pi^2 L_p^2/4W^2$ can be expressed in terms of t/τ_n, where τ_n is the transistor time-constant.

It can be shown[1] that $\tau_n = \dfrac{4W^2}{2{\cdot}43\, D_p}$

Now $D_p = (L_p^2/\tau_p)$,

$$\therefore\ t/\tau_p\,(\pi^2 L_p^2/4W^2) = (t/\tau_n)4{\cdot}06$$

Hence the exponent can be written $(-(t/\tau_n)) \,.\, 4{\cdot}06 \,.\, n^2$

The Collector and Emitter Currents

The collector current I_c is given by

$$I_c = e D_p \left(\frac{\partial\Delta P}{\partial x}\right)_{x=2W} \text{ where } e = \text{charge of a hole.}$$

* *Defence Research Board, Canada*

From equation (2),

$$I_c = -e\,D_p\,.\,2m\left\{\sum_{n=1}^{\infty}\cos(n\pi x/2W)\exp\left[-(t/\tau_p)\frac{n^2\pi^2L_p^2}{4W^2}\right]\right\}_{x=2W}$$

Now $\cos n\pi = (-1)^n$

$$\therefore I_c = +e\,D_p\,.\,2m\sum_{n=1}^{\infty}(-1)^{n+1}\exp\left[-(t/\tau_p)\frac{n^2\pi^2L_p^2}{4W^2}\right]$$

The series is convergent for $t \gg 0$, and for $t \to 0$ the limit is 0·5 . I_c has been determined for various values of t/τ_p $\pi^2L_p^2/4W^2$ and the results are presented graphically in Fig. 2. For large values of t the first term in the series predominates.

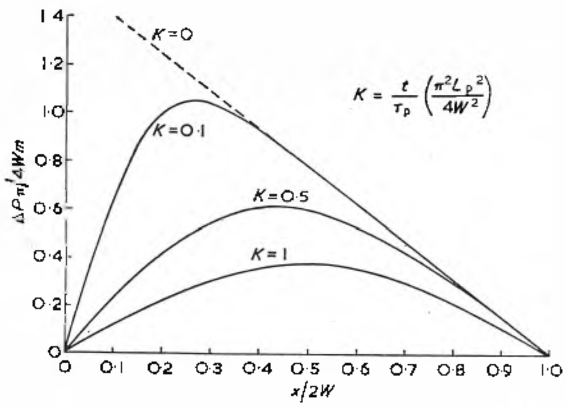

Fig. 1. Excess hole density in the base region

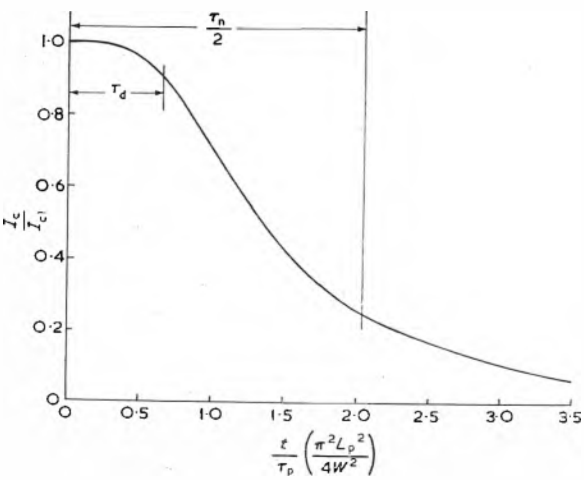

Fig. 2. Decay of the collector current

Thus for $t > 0{\cdot}4\ \tau_n$

$$I_c \sim 2e\,D_p\,.\,m\,.\,\exp(-4t/\tau_n)$$

A delay time can be defined, that is the time taken for the collector current to change by 10 per cent of its initial value. This is given approximately by:

$$\tau_d = 0{\cdot}16\ \tau_n$$

In a similar manner it can be shown that the emitter current is given by

$$I_e = 2e\,D_p\,m\sum_{n=1}^{\infty}\exp\left[-(t/\tau_p)\frac{n^2\pi^2L_p^2}{4W^2}\right]$$

The sum of the series tends to infinity as $t \to 0$, Also, for $t/\tau_p\,.\,\pi^2L_p^2/4W^2 > 1{\cdot}5$, the collector and emitter currents become approximately equal in magnitude, and this is shown graphically in Fig. 3, where I_{e1} is the steady state emitter current.

Measurements

Measurements were made on a number of different transistors to determine how closely the theoretical prediction approximated the practical results.

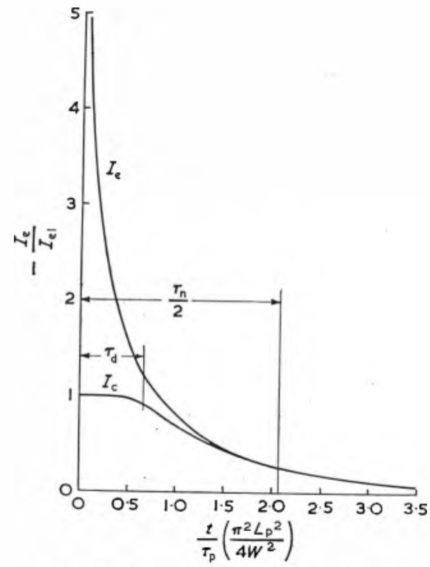

Fig. 3. Decay of the emitter current

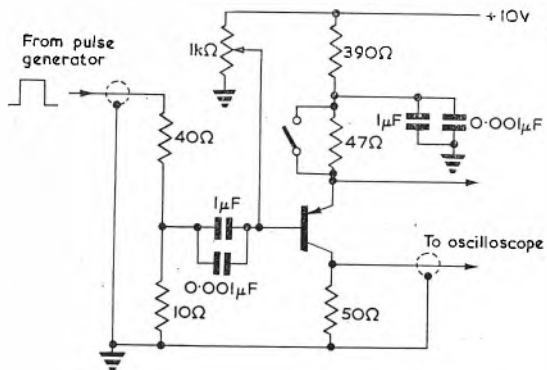

Fig. 4. Circuit for measuring decay times

Fig. 4 shows the circuit for measuring the collector and emitter currents when a large reverse current was applied. A Western Electric mercury relay provided an energizing pulse that had a rise-time of less than 1mμsec, but as the oscilloscope had a rise-time of 13mμsec reliable measurements of transistors with f_α greater than 3Mc/s could not be made.

The maximum initial reverse base current was limited by the extrinsic base resistance, r_b', and the amplitude of the voltage pulse available at the base terminal. Nevertheless, this was greater than 150mA for the values of r_b' encountered.

Variation of the pulse amplitude showed that for sufficiently large reverse base currents the emitter and collector decay times became independent of the change in reverse base current. It was under these conditions that measurements of the decay times were made.

The effect of the collector capacitance was to produce a large initial increase in the collector current, thus obscuring any plateau resulting from the delay time τ_d. An additional effect[3] was the reduction of minority carrier density resulting from the variation of f_α with the collector to base voltage. As this voltage increases so the effective base width decreases, resulting in an extra flow of minority carriers out of the collector. The effect of this last phenomenon on the collector current has not been determined, though for high frequency transistors it may be quite significant.

Figs. 5 and 6 show the emitter and collector current waveforms for two pnp transistors whose alpha cut-off frequencies are 0·1Mc/s and 0·68Mc/s respectively.

Measurements of the decay time of the collector and emitter currents to 10 per cent of their initial values indicate an agreement to within 30 per cent of the calculated values.

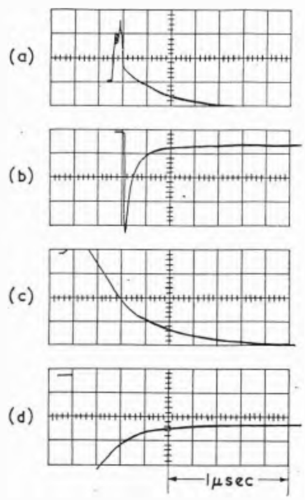

Fig. 5. G.E.V.T.D. 112 transistor, $\tau_n = 1{\cdot}6 \cdot 10^{-6}$sec

(a) Collector current 15mA/Division
(b) Emitter current 35mA/Division
(c) Collector current (expanded scale) 4·7mA/Division
(d) Emitter current (expanded scale) 8mA/Division

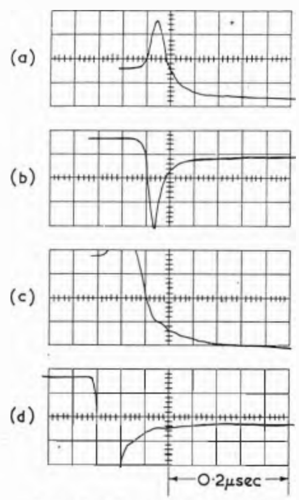

Fig. 6. Texas 301 transistor, $\tau_n = 0{\cdot}234 \cdot 10^{-6}$sec

(a) Collector current 15mA/Division
(b) Emitter current 22mA/Division
(c) Collector current (expanded scale) 4·5mA/Division
(d) Emitter current (expanded scale) 10mA/Division

This is surprising in view of the gross approximation made in the theory.

Conclusions

From a simplified one-dimensional model of a junction transistor, expressions have been derived for the decay of the collector and emitter current that occurs when a very large reverse base current is applied. From these expressions it has been shown graphically that both the collector and emitter currents decay to 10 per cent of their steady state values in approximately three-quarters of a transistor time-constant.

By neglecting the effect of collector capacitance, it is found theoretically that a collector current delay time of $0{\cdot}16\ \tau_n$ occurs, but in practice this effect is masked by the collector space charge capacitance.

Experimental measurements of the collector and emitter currents for several types of pnp transistor show a close agreement with the theoretical predictions. It thus seems reasonable to assume that the results will be applicable to any type of alloyed junction transistors, and will prove to be of use in the design of circuits where a very fast turn-off is required.

Acknowledgments

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Appendix

Solution of the Diffusion Equation

$$\partial\Delta P/\partial t = -(\Delta P/\tau_p) + D_p\,(\partial^2 P/\partial x^2)$$

Assume a solution of the form $\Delta P = T(t)\,.\,X(x)$ then $\Delta P = [C_1 \cos ax + C_2 \sin ax]\exp(-t/\tau_p)\exp(-D_p a^2 t)$

Now at $x = 0$ $\quad \Delta P = 0$

$\therefore C_1 = 0$

Also, at $x = 2W$, $\quad \Delta P = 0$

$\therefore a = (n\pi/2W)$ where $n = 1,2,3$---

Hence,

$$\Delta P = \sum_{n=1}^{\infty} C_n \sin a_n x \exp\left[-(t/\tau_p)(1 + a_n^2 L_p^2)\right] \quad \ldots. (4)$$

using the relation $L_p^2 = D_p\tau_p$

At $t = 0$, $\Delta P = mx + q$

Thus, expanding $mx + q$ in a series

$$mx = \frac{m\,.\,4W}{\pi}\sum_{n=1}^{\infty}\frac{(-1)^{n+1}}{n}\sin(n\pi x/2W)$$

valid for $0 < x < 2W$

$$q = \frac{q\,.\,4}{\pi}\sum_{n=1}^{\infty}\frac{\sin(2n-1)(x\pi/2W)}{2n-1}$$

$$mx + q = (4/\pi)\Big[(q + Wm)\sin(x\pi/2W) - (Wm/2)\sin(2\pi x/2W) + \frac{q + Wm}{3}\sin(3\pi x/2W) - \text{-----}\Big] \quad \ldots\ldots\ldots\ldots (5)$$

Comparing the terms in equations (4) and (5)

$$C_n = (4/\pi)\,.\left(\frac{q + Wm}{n}\right) \text{---- for } n = 1,3,5,7,\text{----} \quad \ldots. (6a)$$

$$C_n = -(4/\pi)\,.\,(Wm/n)\text{---- for } n = 2,4,6,\text{----} \quad \ldots. (6b)$$

Substituting equations (6a) and (6b) into equation (4)

$$\Delta P = \sum_{n=1,3,5\cdots}^{\infty} 4/\pi\left[\frac{q + Wm}{n}\sin(n\pi x/2W)\right]\exp\left[-(t/\tau_p)\left(1 + \frac{n^2\pi^2 L_p^2}{4W^2}\right)\right] - \sum_{n=2,4\cdots}^{\infty} 4/\pi\left[\frac{Wm}{n}\sin(n\pi x/2W)\right]\exp\left[-(t/\tau)\left(1 + \frac{n^2\pi^2 L_p^2}{4W^2}\right)\right] \quad \ldots. (7)$$

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Rectangular Hysteresis-Loop Magnetic Cores as Switching Elements

By J. F. Kaposi*, Dipl.Ing., A.M.I.E.E.

Magnetic cores with approximately rectangular hysteresis loops can be used as switching or storage elements because they possess two stable states of remanent magnetization. The loop which is traversed when such a core is switched from one state to the other depends partly on the magnetic properties of the core and partly on the switching waveform. In this article the theoretical basis of core switching using voltage and current pulses, experimental results of switching different types of cores, and some special applications are discussed.

(Voir page 313 pour le résumé en français: Zusammenfassung in deutscher Sprache Seite 318)

TOROIDAL magnetic cores with approximately rectangular hysteresis loops have been available for some years. Their application as bistable elements in switching and data-storage circuits has been studied fairly extensively, and this type of core can now be included among the devices at the disposal of the designer of digital systems. It does not appear to be generally appreciated that the performance of a core when constant voltage pulses are applied to its input winding differs significantly from that obtained when constant current switching pulses are used. For some purposes constant voltage switching is to be preferred; both methods of switching are discussed below.

Constant Current Switching

Switching by constant current pulses is implied in most of the published accounts of equipment employing rectangular hysteresis-loop cores, and this mode of operation is usually assumed in theoretical examinations of the switching process. Constant current switching has consequently been widely discussed[1,2,3,4,5], and only a brief summary of the main findings is given here.

With constant current operation, the switching waveform is a rectangular current pulse (Fig. 1) applied to an input coil wound on the core. The field H is directly proportional to the current, to the number of turns on the input winding, and inversely to the mean diameter of the core. The flux ϕ and hence the flux density B take up the values defined by the hysteresis loop. During switching the flux is reversed; this can be observed by examining the voltage pulse appearing across an output winding.

It can be demonstrated both theoretically and experimentally that a finite time, called the switching time, is needed to reverse the flux in the core. The relationship between the switching time τ, the driving pulse amplitude and the properties of the core material is given by

$$\tau(H_m - H_o) = S_w$$

where H_m is the maximum applied field.

H_o is the threshold field at which the average velocity of the domain boundaries in the core is zero.

S_w is the switching coefficient, which is a constant for a given core.

This equation holds if the switching current pulse has a very short rise-time in comparison with τ and it indicates that the switching time decreases as the driving field is increased. As H_o cannot be calculated very easily, the utility of the equation is limited. When H_m is very large, H_o can be neglected and the equation reduces to $\tau H_m = S_w$. In this form it can be used to deduce the switching time from the magnitude of the switching field.

* *Ericsson Telephones Ltd.*

The output voltage waveform is shown in Fig. 2 (full line). The amplitude is determined by the switching field, the core material and the cross-sectional area of the core. There is a minimum field below which a core will not switch fully. If a switching field of less than this minimum value is used, the output pulse shape becomes rather ill-defined. In Fig. 2 the dotted lines show the output pulses

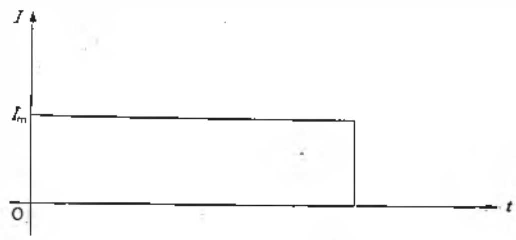

Fig. 1. Constant current waveform

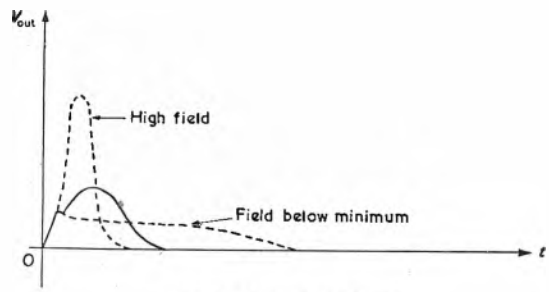

Fig. 2. Output waveforms

for lower and higher values of switching field than that which produces the full-line output pulse.

For the design of circuits in which square hysteresis-loop cores are employed, it is necessary to have more information than that which can be obtained from data-sheets and theoretical studies. The variation of the switching time with input ampere-turns is among the characteristics which need to be known, and by measurements on a range of cores it has been established that the general form of the curve is as shown in Fig. 3. For stable working it is clearly desirable to choose an operating point in a region where the switching time does not change rapidly with the driving current. The region between A and B on the curve is the most suitable from this point of view, but it is not always practicable to operate the core in this region. If the measured output voltage per turn is plotted against the switching time, the general form of the curve is again as shown in Fig. 3.

In Figs. 4 and 5 are shown the measured input ampere-turns against switching speed and output voltage per turn against input ampere turns for various commercially-available cores under constant current switching conditions. The switching time is considerably influenced by the rise-time of the driving pulse; these results were obtained with a driving pulse rise-time of less than 0·4μsec.

Constant Voltage Switching

When a core is switched by the application of a constant voltage pulse to its input winding, the resulting current waveform is of interest. The input impedance is determined mainly by the inductance of the winding, which in turn depends on the permeability of the core material. When switching begins, the permeability is low and the current rises rapidly; when the steep part of the hysteresis loop is reached, the core impedance becomes high and the rate of rise of current is much less. When the core is fully switched, there is little further flux change and the current

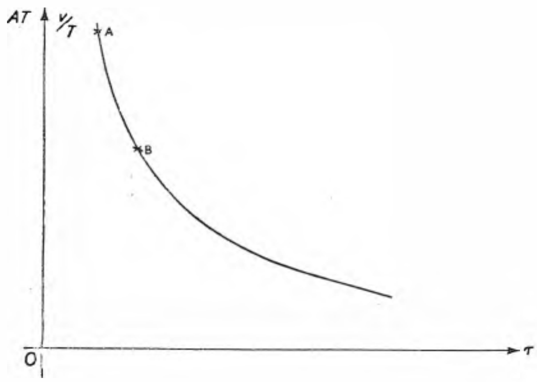

Fig. 3. General form of AT or V/T against τ curves

increases rapidly again, being limited mainly by the source impedance. If the coil is considered to be a pure inductance L, the current i is given by

$$V_o = d/dt\,(iL) \quad \ldots\ldots\ldots\ldots\ldots (1)$$

V_o being the applied voltage.

Since the permeability changes during the switching process, the inductance is not constant. However, if the idealized hysteresis loop of Fig. 6 is considered, the portion traversed during switching of the core from one state to the other is made up of four regions, OA, AB, BC, CD, and the inductance can be taken to be constant in each region separately. For the region OA, if N is the number of turns on the coil, A the cross-sectional area of the core, ϕ the total flux and B the flux density,

$$L = (N d\phi/di) = NA\,(dB/di)$$

and if l = the magnetic path length in the core,

$$i = (lH/N)$$

whence

$$\begin{aligned} L &= (N^2A/l)\,.\,dB/dH \\ &= (N^2A/l)\,.\,\mu_1 \\ &= L_o\mu_1 \text{ writing } (N^2A/l) = L_o \end{aligned} \quad \ldots\ldots\ldots\ldots (2)$$

The current is initially zero, so that from equations (1) and (2) the current while the portion OA of the hysteresis loop is being traversed is given by

$$i_{OA} = \frac{V_o}{L_o\mu_1}\,.\,t \quad \ldots\ldots\ldots\ldots\ldots (3)$$

This equation holds until the current reaches the value $I_A = (H_A l/N)$, which occurs at the time given by

$$t_1 = \frac{H_A l\, L_o\mu_1}{N V_o} \quad \ldots\ldots\ldots\ldots\ldots (4)$$

Similarly, while the portion AB of the hysteresis loop is traversed, the current in the input coil is given by

$$i_{AB} = \frac{V_o}{L_o\mu_2}(t - t_1) + I_A \quad \ldots\ldots\ldots\ldots (5)$$

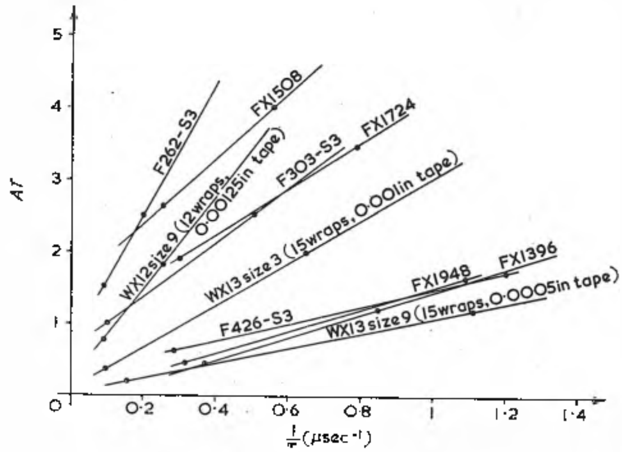

Fig. 4. AT against $1/\tau$ curves
(*Constant current switching*)

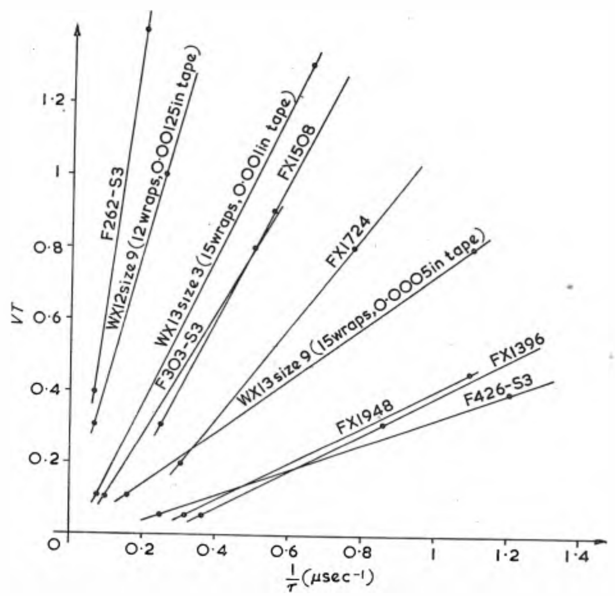

Fig. 5. V/T against $1/\tau$ curves
(*Constant current switching*)

The current at point B is $H_B l/N$ and this point is reached at the time t_2 given by

$$t_2 = \frac{\dfrac{H_B l}{N} - \dfrac{H_A l}{N}}{\dfrac{V_o}{L_o\mu_2}} + t_1 \quad \ldots\ldots\ldots\ldots (6)$$

While the portion BC of the loop is traversed, the current is given by

$$i_{BC} = \frac{V_o}{L_o\mu_1}(t - t_2) + I_B \quad \ldots\ldots\ldots\ldots (7)$$

If the duration of the switching pulse is sufficient, i_{BC} will increase until it is limited by the source impedance, and will remain at the maximum value so determined until the pulse ends. The current waveform represented by these equations is shown in Fig. 7. A somewhat better approximation to an actual hysteresis loop than that shown in Fig. 6 is obtained by considering the loop to consist of linear segments linked by parabolic arcs. However, the only effect this has on the calculated output current waveform is to round off the angles, and the arguments which follow are not substantially affected.

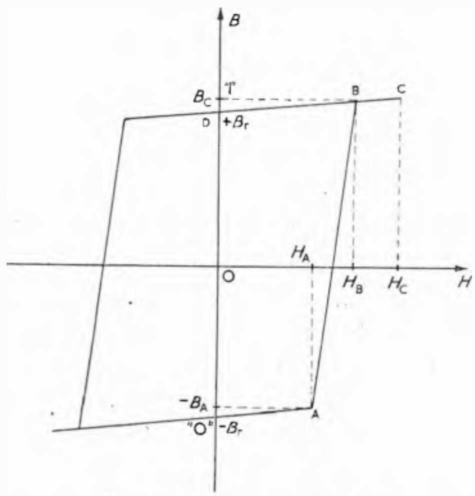

Fig. 6. Idealized hysteresis loop

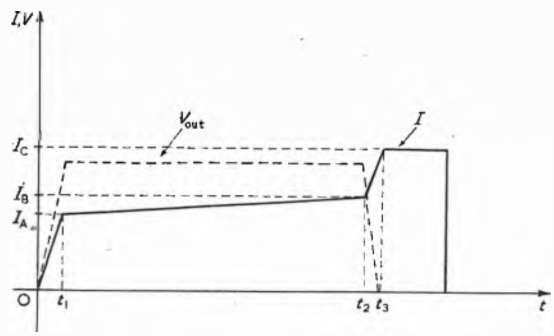

Fig. 7. Current and voltage waveforms
(Constant voltage switching)

The process of transferring the core from one remanent state to the other is completed when, at the end of the switching pulse, the final flux change from point C to point D on the hysteresis loop occurs. However, the intervals $0 - t_1$ and $t_2 - t_3$ in Fig. 7 can in practice always be neglected in comparison with the interval $t_1 - t_2$; in other words, the switching process can be considered to be completed when point B is reached, and the $t_1 - t_2$ interval can be taken as the switching time. Hence from equation (6),

$$\tau = \frac{I_B - I_A}{\dfrac{V_o}{(N^2A/l)\,\mu_2}} \quad \ldots\ldots\ldots\ldots\ldots\ (8)$$

and the switching time can be calculated from the core and coil characteristics and the driving pulse amplitude.

The concept of an equivalent resistor for a switching core, which is helpful in the design of circuits, was introduced by Sands[3]. The equivalent resistor is defined as the resistor which dissipates the same energy as does the core during the switching process, and Sands showed that for a core switched by a constant current pulse its value is

$$R_{cc} = \frac{N^2 \Delta B_m A}{H_m l} \quad \ldots\ldots\ldots\ldots\ldots\ (9)$$

ΔB_m being the total change in flux density.

Under constant voltage switching conditions, the value of the equivalent resistor can be arrived at as follows: In the previous paragraph it was shown that the interval $t_1 - t_2$ can be taken as the switching time. Correspondingly, i_{AB} can be considered as the input current to the core winding. Hence from equation (5),

$$i_{in} = \frac{V_o}{L_o \mu_2}\, t + I_A$$

and the input energy is given by

$$W_{in} = \int_0^{t_2} V_o i_{in}\, dt$$

$$= \int_0^{t_2} \left[\frac{V_o}{L_o \mu_2}\, t + I_A \right] dt$$

$$= \frac{V_o^2}{L_o \mu_2} \cdot (t_2^2/2) + V_o I_A t_2$$

If the core is replaced by an equivalent resistor R_{cv}, the energy dissipated during switching is $(V_o^2/R_{cv}) \cdot t_2$. Hence

$$V_o^2 t_2 / R_{cv} = \frac{V_o^2}{L_o \mu_2} \cdot (t_2^2/2) + V_o I_A t_2$$

whence

$$R_{cv} = \frac{V_o}{(V_o t_2 / 2L_o \mu_2) + I_A} = \frac{V_o}{(I_B + I_A)/2} \quad \ldots\ldots\ (10)$$

because

$$I_B = (V_o t_2 / L_o \mu_2) + I_A \text{ (equation (5))}$$

A special feature of constant voltage switching is that a core can be switched very slowly applying very low voltage pulses across the input winding.

It was shown previously that the input current rises until it reaches a maximum value determined by the source impedance.

Initially the rise is rapid, and the value I_A is reached in a very short time. Switching will be complete or partial depending on whether or not the current reaches a value greater than I_B. From equation (7) it can be seen that if the driving voltage pulse is long enough—that is, if it terminates later than t_2—the final increase of current may result in a value greater than I_B necessary for full switching being reached.

If the amplitude of the driving voltage pulse is very low, the time t_3 necessary for the current to reach the value I_B becomes longer, and if the width of the driving pulse is greater than t_3, the core will be switched in a very long time, which cannot be achieved with constant current switching.

Now it is easy to see that for a given driving pulse width there is a minimum amplitude, and for a given amplitude there is a minimum pulse width below which the core will not switch completely.

These considerations suggest two ways in which cores in an array might be selected by constant voltage switching. Firstly, selection by driving pulse duration might be possible since a core will switch completely if the driving pulse duration exceeds t_2 but will not switch if the pulse duration is less than t_1. Secondly, equation (6) solved for V_o shows that if the driving pulse amplitude V_o exceeds $\mu_2[(I_B - I_A)/t_2]L_o$ the core will be switched completely, while equation (4) again solved for V_o shows that switching cannot take place for amplitudes of V_o less than $I_A(L_o\mu_1/t^*)$. In that case the width of the driving pulse is longer than t_1 but shorter than t_2.

However, with the cores examined so far a practical application of these selection systems could not be found, the ratios t_2/t_1 and μ_2/μ_1 being too big.

In Fig. 8 practical waveforms of constant voltage switching can be seen.

In Figs 9 and 10 the results of experimental tests of constant voltage switching of cores are shown.

In Table 1 comparative results are shown regarding low field constant current and low voltage constant voltage results indicating that in the latter case the core has a useful and well defined output even at very low voltage per turn ratios.

Effect of Temperature

Tests have been carried out on the temperature dependence of the switching properties of various cores.

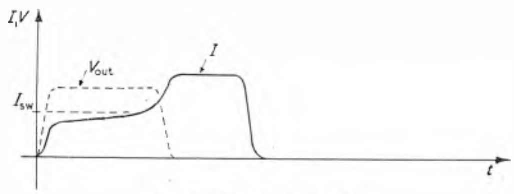

Fig. 8. Practical waveforms
(Constant voltage switching)

TABLE 1

$\frac{V}{T}$	OUTPUT WAVEFORM	
	CONSTANT VOLTAGE	CONSTANT CURRENT
1	5·3μsec	5·3μsec
0·7	8·4μsec	9μsec
0·5	9·5μsec	10μsec
0·3	12μsec	11μsec
0·2	16μsec	
0·1	24μsec	

It has been found that constant voltage switching gives the better results under variable temperature conditions.

As the temperature is raised with constant current switching the voltage per turn ratio increases and the switching time decreases. The rate of change is different for different types of cores and it can be quite significant at temperatures well below the Curie point.

With constant voltage switching, though the switching current decreases with temperature, the switching time—according to equation (8)—remains fairly constant, as $I_B - I_A$ remains constant with temperature. The voltage per turn ratio is not affected as far as the source can be considered a constant voltage source.

In Table 2 some results are collected of the switching of a core under variable temperature and driving conditions. It is evident that more stable operation can be obtained with constant voltage switching.

Some Aspects of Core Circuit Design

The usual practical circuit arrangement for a core is shown in Fig. 11. Suffix 1 indicates the input, suffix 2 indicates the output circuits. In the following it will be assumed that the core is reset when it is necessary and that this is carried out in the same way as the switching of the core.

When calculating the switching field one must take all the currents flowing in the different windings during switching into account.

The constant current switching will be discussed first.

In most practical cases the required output into a given load is known. Thus the output requirements are controlled by I_2, V_2, R and τ. For a given I_2 into a fixed R a par-

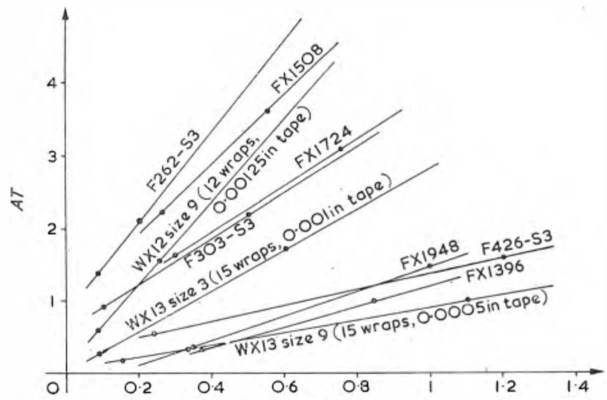

Fig. 9. AT against $1/\tau$ curves
(Constant voltage switching)

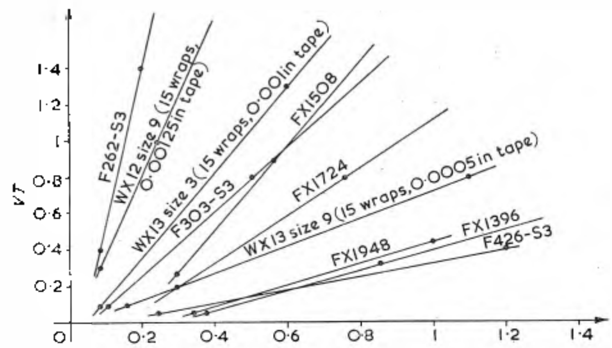

Fig. 10. V/T against $1/\tau$ curves
(Constant voltage switching)

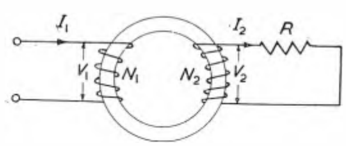

Fig. 11. Simplified core circuit

TABLE 2

°C	SWITCHING TIME (μSEC)		VOLTS/TURN (mV/TURN)	
	CONSTANT VOLTAGE	CONSTANT CURRENT	CONSTANT VOLTAGE	CONSTANT CURRENT
20	3·1	3·1	50	50
40	3·1	3·1	50	54
60	3·1	2·8	50	68
80	3·1	2·4	50	86
100	3·0	1·9	50	100

ticular V_2 is required at a particular τ.

$$V_2 = N_2 (d\phi/dt) = N_2 \frac{\text{Volts}}{\text{Turn}}$$

From that N_2 can be calculated.

$$N_2 = V_2 \div \frac{\text{Volts}}{\text{Turn}} \quad \ldots\ldots\ldots\ldots\ldots \ (11)$$

where the $\frac{\text{Volts}}{\text{turn}}$ ratio can be found from the diagrams in Fig. 4 and 5 after having chosen a particular core.

If there is no load, the magnetic field is given by equation (12) and can be obtained from manufacturers' data sheets.

$$H_{mo} = (N_1 I_{10}/l) \ \ldots\ldots\ldots\ldots\ldots\ldots \ (12)$$

where I_{10} = primary current without load.
l = mean magnetic path length.

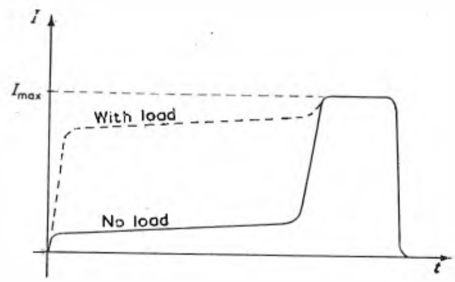

Fig. 12. Effect of load on the current waveform
(Constant voltage switching)

When a load I_2 is present, more current is required to switch the core in the same switching time.

$$H_1 = H_{mo} + H_2$$

$$\therefore \ I_1 = I_{10} + (N_2/N_1)\, I_2 \ \ldots\ldots\ldots\ldots \ (13)$$

It can be seen that, by increasing the number of the input turns, the necessary primary current becomes less. Obviously the voltage across the input winding increases. In each particular application a compromise between input turns and input current must be made and this is determined by the available driving source.

Equation (13) suggests that—if the necessary current is provided in the input—any load can be supplied.

In other words a core can be switched from the output of another core if the load is supplied and the driven core switches faster than the driving core. As the driving core is purely transferring the energy from the generator to the driven core, the energy required to switch the driving core is a loss. In order to reduce this it is advantageous to choose the driving core of small size.

In one particular case a WX13 size 9 core was switched by a Plessey F426 core, and it was found that 1·5 to 2At less were needed to perform this operation than with a larger driving core (Plessey F262).

When calculating the equivalent of a core-circuit one must be very careful in considering the value of the equivalent of cores. In the corresponding formula of equation (9) the value of B_m depends on the applied switching field and on the previous magnetic state of the core. If a large field of H_2 is applied to a core which had been recycling on a smaller loop, the total flux change is $B_{H2\text{-}H1} = \frac{1}{2}B_{H1} + \frac{1}{2}B_{H2}$, while if the core was left in point A, the total flux change is $B_{H2} > B_{H2\text{-}H1}$ (see Fig. 13).

The reasons for using constant voltage switching can be (1) a rectangular output pulse required, (2) minimizing the effect of changing temperature, and (3) the core may have to supply different loads from time to time. Assuming that the available driving current is always sufficient, variable loads can be put on the output of a constant voltage switched core, because the voltage per turn ratio, and hence the switching time, will not change. The core always takes the required current from the driving source, so that there is nothing to be adjusted. Calculations, however, should be made for defining the number of turns and the necessary maximum current.

Equations (11) for N_2 and (13) for I_{max} can be used in that

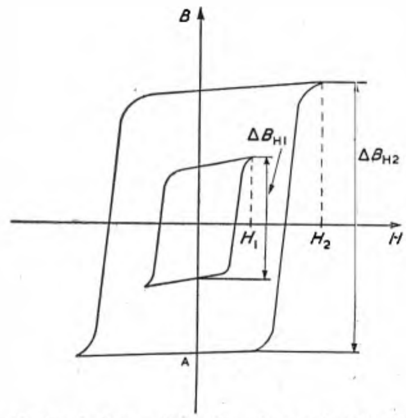

Fig. 13. Equivalent resistor depends on the flux change

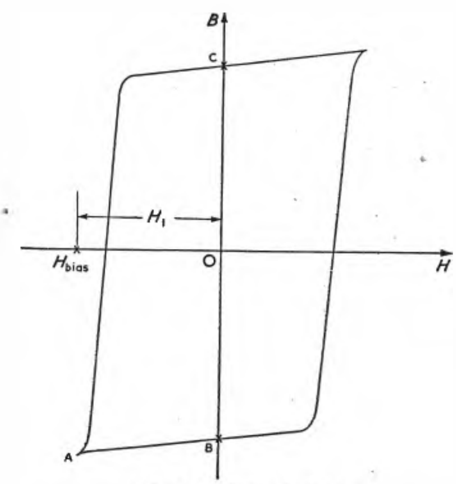

Fig. 14. Hysteresis loop
(Biased full-current matrix)

case as well, and the available maximum load current is given by equation (14).

$$I_{2(max)} = \frac{N_1(I_{max} - I_B)}{N_2} \ \ldots\ldots\ldots\ldots\ldots \ (14)$$

In Fig. 12 the input currents can be seen for different loads. The switching action is stable if the value of I_1 defined by equation (13) does not exceed the value of the final current. If it does, the actual switching field starts to decrease, hence the voltage per turn ratio decreases and the switching time increases.

Consider a biased-full-current-matrix. In this all the cores of a matrix are biased by a current pulse or by a direct current to some point A, Fig. 14. A constant current pulse is sent along the selected row and this, opposing the bias of the row, sets all the cores of this row to some point B, while the rest of the matrix cores are at point A. Shortly after the start of this current pulse a constant voltage (or constant current) pulse is applied to the selected column, switching the selected core. The switching back of the core is done by the bias current after the switching pulses have finished.

In the following the loading effect of the bias circuit is discussed.

A simplified circuit diagram of the selected core is shown in Fig. 15. When the core is switching, an induced voltage

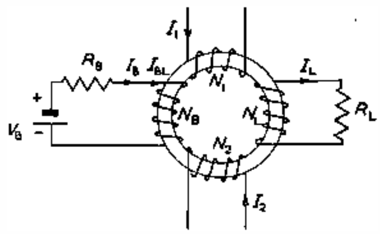

Fig. 15. Simplified diagram of a core in the biased full-current matrix

appears across each winding. Due to this voltage a current pulse will flow in the bias circuit, which tends to flow in the same direction as the bias current itself. In other words, the bias circuit represents an additional load.

According to Fig. 15 the switching field is given by

$$H_{SW} = (H_1 - H_{bias}) + [H_2 - (H_{load} + H_{B(load)})]$$

indicating that H_2 is available to supply the true load (H_{load}) and the unwanted load ($H_{B(load)}$).

Calculating $I_{B(load)}$,

$$I_{B(load)} = \frac{H_{bias} \cdot l \cdot V/T}{V_{bias}} \qquad (15)$$

where $H_{bias} \simeq H_{SW}$, l = mean magnetic path length, V/T = volts/turn — are fixed

Equation (15) suggests the method of decreasing $H_{B(load)}$ by increasing V_B and hence R_B for keeping H_B the same.

When designing a core circuit, the disturbances have to be considered. Here only a special note will be written in connexion with half-current coincidence matrices.

In this kind of system the disturbed output is affected by the tolerance on the 'read' and 'write' currents and the hysteresis loop on which the core is working.

Assume a particular application of small ferrite cores with a nominal full 'read' of 380mAt which, due to tolerances, can take any value between 340 and 420mAt. The half-write field can vary from 160 to 190mAt.

A series of tests has been made. The sequence of measurements and the results are presented in Table 3.

At first it can be seen that the variation of the half current has only a slight effect on the disturbed output. Increasing the full 'read' to 420mAt the disturbed output becomes smaller. Now decreasing the full 'read', the output becomes slightly less than it was before, working almost on the same hysteresis loop. After applying full 'write' of 340mAt and full 'read' of 340mAt and then half 'write' of 340mAt and full 'read' of 340mAt and then half 'write' of 160mAt and full 'read' of 340mAt, the disturbed output is the same as it was at the beginning of this series of tests.

TABLE 3

	WRITE (mA IN 1 TURN)	READ (mA IN 1 TURN)	OUTPUT (mV ACROSS 1 TURN)
1	340	340	20
2	160	340	2·0
3	190	340	2·1
4	190	420	1·6
5	190	340	1·3
6	340	340	20
7	190	340	2·1

It can be seen that the biggest disturbed output appears when working on the smallest hysteresis loop. The reason for that can be that the smallest loop is less square than the bigger ones.

Conclusions

There is no general formula to be followed when designing a core device. The choice of core material and size should be decided individually. However, a few points may be helpful.

Core Types and Materials

Fast and Slow Switching Cores

It is unnecessary to point out the importance of a fast computing or memory element. However, slow cores have several advantages for devices where the speed is not of primary importance. Some of these advantages are: (a) The driving source need not have such a high frequency response which results in lower equipment cost and (b) when switching the slow core at a faster rate, the operating point can be chosen on the more stable region of the characteristic.

Ferrite and Metal Tape Cores

Though available metal tape cores have a less square hysteresis loop than ferrites, their other useful properties make their application advantageous. (1) They need less switching field than ferrite cores of similar switching properties. (2) By controlling the number of wraps of the metal tape, practically any size and cross-sectional area can be obtained.

All the Plessey, Mullard and Scott cores are usable up to 50°C. The slow switching metal tape core showed the best performance at temperatures up to 100°C.

Choice of Driving Conditions

Constant voltage switching is recommended whenever there is a free choice. This method of operation is more stable, less temperature dependent and is able to supply a variable load without adjustment of the input.

Design

The equations give fairly accurate results and are very useful. However, it is always advisable to study all the available practical information on cores and experimental checks should always be carried out.

Acknowledgments

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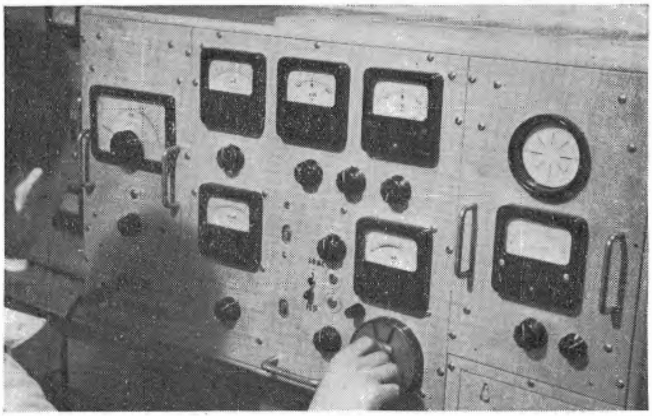
Accurate Measurement of Transistor Cut-off Frequency

By Yasuo Tarui*



An equipment is described for the quick and accurate measurement of current gain, α, of high frequency transistors in the frequency range between 1 to 20Mc/s. The absolute value and phase of α can be measured with an accuracy of 1·5 per cent. Quick determination of α cut-off frequency is possible.

(Voir page 313 pour le résumé en français: Zusammenfassung in deutscher Sprache Seite 318)

FOR the measurement of the current gain, α, the most important of the high frequency parameters, Coffey[1] has proposed a heterodyne method which uses a separate signal generator and a receiver. By his method the signal generator is set to a given frequency, the receiver is tuned to this frequency, and the measurement is done at this specific frequency. This is time consuming and at the same time the frequency drift in both the signal generator and the receiver local oscillator make it difficult to perform stable and accurate measurement.

In the method proposed in this article the frequency of the local oscillator is locked to the frequency of the signal generator by the difference of the intermediate frequency. The most important portion in the circuit is the termination of the transistor. The error caused by this termination was considered and the overall accuracy in the determination of f_α was estimated to be 5 per cent in a quick measurement and 2·5 per cent by repeated zero balancing around the specific frequency.

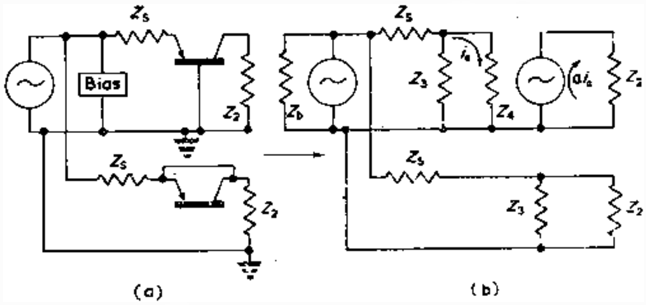


Fig. 1. The method adopted in this equipment

Frequency Range and Accuracy

A frequency range of 1 to 20Mc/s was decided upon and an accuracy of 5 per cent was set as a goal in the determination of f_α. It is necessary to know with what accuracy α must be measured, and for this purpose the following approximation of α was used:

$$\alpha = \frac{\alpha_0 \, e^{iKT}}{1 + i\,(f/f_\alpha)} \quad \ldots\ldots\ldots\ldots\ldots\ (1)$$

where K is a constant which is independent of the frequency, and α_0 is α at low frequency. Then,

$$|\alpha| = \frac{\alpha_0}{\sqrt{[1 + (f/f_\alpha)^2]}} \quad \ldots\ldots\ldots\ldots\ (2)$$

Putting

$$\frac{\sqrt{2}\,|\alpha|}{\alpha_0} = x \text{ and } f/f_\alpha = y$$

$$dy/dx = -\frac{x^2(2 - x^2)^{-\frac{1}{2}} + (2 - x^2)^{\frac{1}{2}}}{x}$$

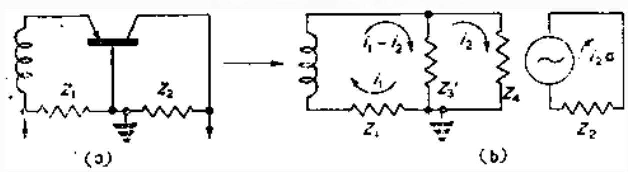


Fig. 2. The method adopted by W. N. Coffey

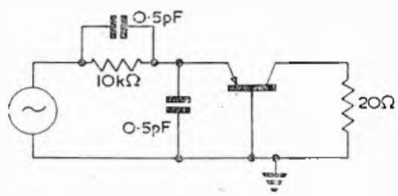


Fig. 3. Stray components assumed in error calculation

at α cut-off frequency $|\alpha| = (\alpha_0/\sqrt{2})$ i.e. $x = 1$

$dy/dx = -2.$

This means that if f_α is required to an accuracy of x per cent, then α must be measured to an accuracy of $x/2$ per cent. Therefore, to meet the stated requirements α must be measured to an accuracy of 2·5 per cent and the goal of the accuracy of phase measurement is 3°. From the consideration of accuracy in phase measurement and ease of amplification an i.f. of 10kc/s was adopted. In the measurement of α_0 10kc/s was used also. The error introduced by the fact that 10kc/s is not a very low frequency can be calculated from equation (1). It is shown that for the severest case of $f_\alpha = 1$Mc/s, the error is not greater than 10^{-4}.

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The above photograph shows the equipment in use.

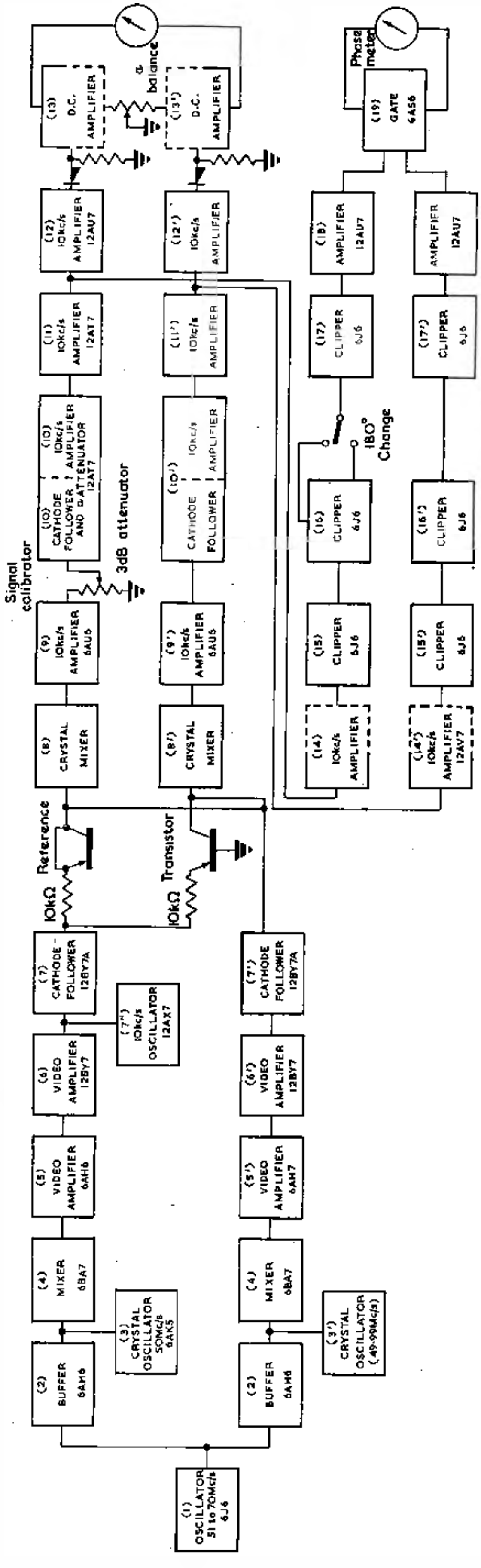


Fig. 4. Block diagram of the equipment

Termination

For the measurement of α a load resistor is inserted in the collector circuit. To make a rough approximation for the allowable value for this resistor the following equations are checked:

$$v_i = h_i i_i + h_r v_o \quad \ldots\ldots\ldots\ldots \quad (3)$$

$$i_o = h_f i_i + h_o v_o$$

$$v_o = R i_o \quad \ldots\ldots\ldots\ldots \quad (4)$$

where R is the load resistance.

Then

$$h_f = -\alpha = \frac{i_o - h_o v_o}{i_i} = i_o/i_i\,(1 - h_o R) \quad \ldots\ldots \quad (5)$$

The error introduced by this load resistor is in the ratio $(1 - h_o R)$. If it is assumed that the output impedance is of the order of ωC_c, then this term becomes:

$$(1 - h_{22}R) = (1 + \omega^2 C_c^2 R^2)^{\frac{1}{2}} \tan^{-1} \omega CR \quad \ldots\ldots \quad (6)$$

The error in $|\alpha|$ is approximately $\frac{1}{2}\omega^2 C_c^2 R^2$ and the error in phase is $\tan^{-1} \omega CR$. For an accuracy of 3° in phase at 20Mc/s with $C_c = 20$pF, the load resistance is calculated to be 20·8Ω. With a load resistance of 20Ω the error in $|\alpha|$ is calculated to be $1{\cdot}26 \times 10^{-3}$. This is a negligible error for the present purpose. It must be noticed, however, that the assumption of h_o is only an approximation. For a detailed check of the error it is necessary to consider the real value of h_o with the inclusion of $\gamma_{bb'}$.

For the termination on the input side two methods were compared; these are shown schematically in Fig. 1 and Fig. 2. Fig. 1 shows the method adopted in this equipment and Fig. 2 shows the one by Coffey. Figs. 1(a) and 2(a) are schematic diagrams, while Figs. 1(b) and 2(b) are equivalent circuits including stray elements. The method by Fig. 1 is to compare the output of transistor into Z_2 in the upper part to the output from the reference circuit in the lower part. The reference circuit is comprised of a circuit which has exactly the same construction as the transistor circuit except that the emitter and the collector are short-circuited at the socket. Measurements are made by comparing the output of the transistor to the voltage which is attenuated α' times the output of the reference circuit. The error introduced by measuring α of the transistor by indication of attenuation α' is estimated. The comparison described above implies:

$$\frac{\dfrac{Z_3}{Z_3 + Z_4}\,\alpha}{\left(Z_5 + \dfrac{Z_3 \,.\, Z_4}{Z_3 + Z_4}\right)} = \frac{\dfrac{Z_3}{Z_2 + Z_3}\,\alpha'}{\left(Z_5 + \dfrac{Z_2 \,.\, Z_3}{Z_2 + Z_3}\right)} \quad \ldots. \quad (7)$$

Then,

$$|1 - (\alpha/\alpha')| = \left|\frac{(Z_2 - Z_4)(Z_5 + Z_3)}{(Z_2 + Z_3)Z_5 + Z_2 \,.\, Z_3}\right|$$

Assuming $Z_2 \ll Z_3 \quad Z_2 \ll Z_5$

$$\simeq \left|(Z_2 - Z_4)\frac{(Z_5 + Z_3)}{Z_5 \,.\, Z_3}\right| = |Z_2 - Z_4|\left|\frac{Z_5 + Z_3}{Z_5 \,.\, Z_3}\right| \quad \ldots. \quad (8)$$

In the second method in Fig. 2 the emitter current and the collector current are compared directly. Z_3' is the stray impedance and Z_1 is the input impedance of the transistor. For the initial balance, the emitter and the collector are short-circuited at the socket, the magnitude and phase of the output of the two amplifiers which follow Z_1 and Z_2 are balanced. For the case of calculation the amplification and phase of the amplifiers are included in the value Z_1 and Z_2.

The condition for initial balance is:

$$Z_1 = \frac{Z_2 \cdot Z_3'}{Z_2 + Z_3} \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (9)$$

α of transistor is measured by balancing the two outputs

$$\left.\begin{aligned} \alpha' i_1 Z_1 &= \alpha i_2 Z_2 \\ i_2 \cdot Z_4 &= (i_1 - i_2) Z_3' \end{aligned}\right\} \quad \ldots\ldots\ldots\ldots\ldots \quad (10)$$

$$\begin{pmatrix} \alpha' Z_1 & -\alpha Z_2 \\ Z_3' & -(Z_3' + Z_4) \end{pmatrix} = 0 \quad \ldots\ldots\ldots\ldots \quad (11)$$

From equations (9) and (11) the ratio α/α' is known to be

$$\alpha/\alpha' = \frac{(Z_3' + Z_4)}{(Z_2 + Z_3')} \quad \ldots\ldots\ldots\ldots\ldots \quad (12)$$

Then

$$|1 - (\alpha/\alpha')| = \left|\frac{Z_2 - Z_4}{Z_3' + Z_2}\right|$$

Assuming $Z_2 \ll Z_3'$

$$\simeq |(Z_2 - Z_4)\, 1/Z_3'| = |(Z_2 - Z_4)|\ |1/Z_3'| \quad \ldots\ldots \quad (13)$$

Equations (8) and (13) show that the comparison of the accuracy of Fig. 1 and Fig. 2 are made by the two terms:

$$(1) \quad A = \left|\frac{Z_5 + Z_3}{Z_5 \cdot Z_3}\right| \quad \ldots\ldots\ldots\ldots\ldots \quad (14)$$

$$(2) \quad B = |1/Z_3'| \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (15)$$

Here it may be noticed that Z_3 and Z_3' are not of the same magnitude because Z_3 can be made small by directly connecting Z_5 to the transistor socket, whereas Z_3' includes stray elements up to the coil tapping. The accuracies can now be examined assuming specific values for each component. If it is assumed that Z_3 is a stray capacitance of 0·5pF and Z_5 is composed of a 10kΩ resistor and a 0·5pF capacitor in parallel, A in equation (14) is calculated to be:

$$A = 1{\cdot}6 \times 10^{-4}\ (\Omega^{-1})$$

Then if the inequality

$$|1/Z_3| < 1{\cdot}6 \times 10^{-4}$$

is satisfied, the method in Fig. 2 is better. It means that the stray capacitance must be smaller than 1·28pF, and this is rather difficult by the method in Fig. 2. Moreover by the method in Fig. 2 the error introduced by the by-pass effect in the coil layer itself cannot be estimated.

Therefore the method in Fig. 1 was utilized. The maximum error introduced by this method is estimated, by assuming the transistor input impedance to be in the range between 0 to 50Ω and the stray components as in Fig. 3, to be:

$$|Z_2 - Z_4| \cdot \left|\frac{Z_5 + Z_3}{Z_5 \cdot Z_3}\right| = 70 \times 1{\cdot}6 \times 10^4 = 1{\cdot}12 \text{ per cent}$$

Circuit Description

A schematic diagram of the equipment is shown in Fig. 4. The master oscillator (1) generates signals in the range 51 to 70Mc/s. This master signal is mixed with signals from the oscillators (3) and (3′) which are controlled by crystals to 50Mc/s and 49·99Mc/s respectively. Therefore the accuracy of the intermediate frequency (10kc/s) is determined by the difference of the stabilities of two crystal controlled oscillators. The signal from the mixer (4) is amplified to 0·5V by the video amplifiers (5) and (6). Then this signal is supplied to the transistor and the reference circuit through the cathode-followers (7). To measure the low frequency α i.e., α_0, the 10kc/s oscillator 7″ is added. The h.t. supply is switched to (7″) or (1) to (6) and (1′) to (6′) by a simple switch for the measurement of α_0 and high frequency α respectively. Output voltages of the transistor and reference circuit are mixed with the local signal from (7′). To avoid the change of conversion gain with the level of local signal, a large enough local signal (about 1V) is supplied to operate in the saturation range. 10kc/s signals from both the transistor and reference circuit are amplified by (9) and (9′). The reference channel is attenuated by α and 3dB attenuators in the cathode-follower (10). After amplification by (11), (12), (11′), (12′) and rectification, the signals are compared by d.c. reference meters (13) and (13′). For the phase measurement, the outputs of (11) and (11′) are amplified and clipped and supplied to g_1 and g_3 of a gate valve type 6AS6, which conducts only when g_1 and g_3 are both positive.

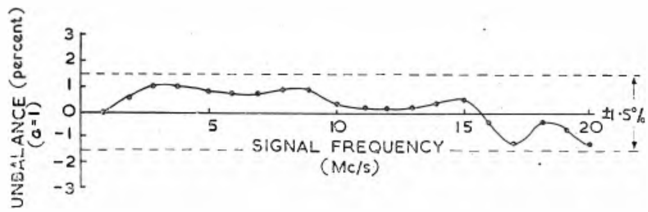


Fig. 5. Balance of 2 channel amplifiers

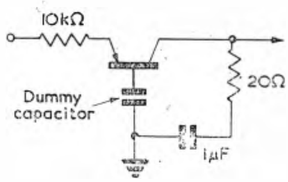


Fig. 6. Check with dummy

Measurements are made by the following procedures. To get an accurate value of α at a specific frequency:

(1) The transistor terminal is short-circuited at the emitter and the collector.

(2) The α attenuator is set at 1.

(3) The gains of the two channel amplifiers are equalized with the signal calibration attenuator by the output balancing meter.

(4) By inserting the transistor and balancing the output again with the α attenuator, α can be measured from the attenuator dial.

The maximum error in this case is estimated to be approximately 1·35 per cent.

(1) Check the balances at 10kc/s and 1Mc/s with a short-circuit at the emitter and the collector.

(2) α_0 is measured by the α attenuator at 10kc/s with the transistor in the socket.

(3) Switch to high frequency and attenuate the gain of the reference circuit by the 3dB attenuator.

(4) Then get a balance of the output meter by sweeping on the frequency dial. The frequency where the balance is again obtained is f_α.

In this case the error in the difference of frequency characteristics of the two channel amplifiers is included.

Checking the Test Set

Fig. 5 shows balance of the two channel amplifiers. This was taken by short-circuiting at both transistor and reference circuit and the signal calibration balanced at 1Mc/s. The frequency was then changed by the main dial and the deflexion in the output balance meter was measured.

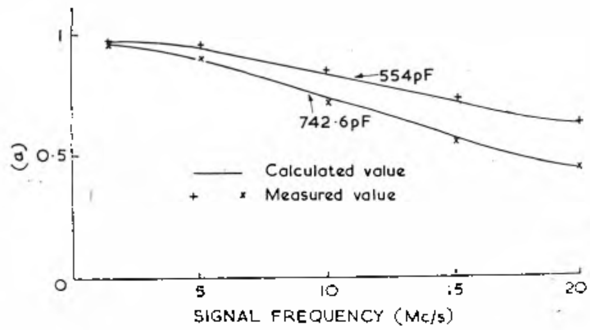


Fig. 7. Change of magnitude with dummies

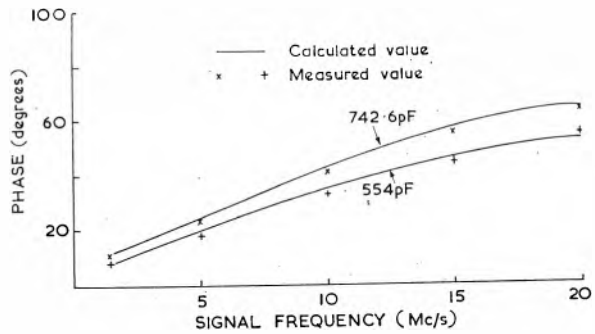


Fig. 8. Change of phase with dummies

An overall accuracy check was made by using an accurately measured short lead capacitor inserted as in Fig. 6, and comparing the calculated and the measured attenuation and phase shift. The results for two values of capacitors are shown in Fig. 7 and Fig. 8 for magnitude and phase respectively. From these figures it can be seen that in the range

$$f = 1 \sim 20\text{Mc/s}, \quad \alpha = 1 \sim 0{\cdot}5,$$

phase = 0 – 60°, accuracy in α is $\pm 1{\cdot}5$ per cent and in phase $\pm 2°$. This is not a complete check, although with the design considerations and the above checks it is believed that the required accuracy has been achieved.

Acknowledgment

This equipment was planned by the Transistor Surveying Committee in Japan, the Chairman of which is Dr. H. Wada, sponsored by Denpa Gijutsu Kyōkai, and constructed at Tokyo Denpa Kogyo Co.

The author wishes to express his sincere thanks to Dr. H. Wada and all members of the committee for their guidance and discussion and A. Obuchi, N. Sekita and Nabeta of Tokyo Denpa Kogyo Co. for the construction of this equipment.

REFERENCE

1. Coffey, W. N. Measuring R.F. Parameters of Junction Transistors. *Electronics, 29,* 152 (Feb. 1956).

Measurement of Vibration Amplitude

Equipment has been developed by the Mechanical Engineering Research Laboratory of the Department of Industrial Research for the direct measurement of peak-to-peak amplitude of mechanical vibrations. Its frequency range is from 0 to 4kc/s but it is particularly suitable for use at frequencies below 50c/s; it can conveniently measure vibration amplitudes down to 0·00002in.

The principle on which the instrument is based is the production of interference bands or moiré fringes from two gratings placed face to face with their lines not quite parallel. Relative motion between the gratings in a direction at right-angles to the lines produces motion of the fringes such that when the gratings move through one line pitch the fringes move through one fringe pitch.

If a pair of gratings are illuminated by collimated light, then the distribution of light intensity across the fringe pattern is approximately sinusoidal and four photocells can be arranged to receive light signals in quadrature. Applying the amplified output of the photocells to the plates of a cathode-ray tube produces a circular trace for continuous motion of the gratings, and an arc of a circle for vibratory motion. The angle subtended by the arc is directly proportional to the amplitude of vibration and when the arc just closes into a circle, the peak-to-peak amplitude is equal to the pitch of the gratings. Hence vibration amplitudes can be measured in terms of the known pitch of the gratings down to approximately 0·05 of the pitch and up to a value depending on the maximum number of complete circles which can be identified on the trace. Using 2 500 lines/inch gratings, this gives a discrimination of 0·00002in.

The light source consists of a low-wattage projection lamp and an aspherical collimating lens. The lamp is fitted in a sliding holder so that the transverse filament can be moved about the optical axis, thus altering the angle at which the light is incident on the gratings. The fixed grating is mounted in a holder which can be rotated about the optical axis by means of a lever and screw arrangement. This provides a sensitive means of adjusting the small angle by which the lines on the gratings are crossed. The holder can also be adjusted axially to set the gap between the gratings. The moving grating is mounted in a block cantilevered out from the main body on two parallel spring strips, the vibratory motion being imparted to the block through a ball-ended stylus.

The fringe field is focused on to the photocells by the combined four-element prism and lens. The light falling on each facet is deviated and focused on to a separate photocell, the geometry of the arrangement being governed by the physical size of the cells. In order to bring the signals into quadrature, the fringe pitch is made equal to the combined width of the four facets by adjusting the crossed angle between the lines on the gratings. This adjustment is effected in practice by observing the degree of circularity of the trace on the c.r.t. There are now four spectra focused on the plane of the photocell apertures which are approximately 0·060in square and act as slits to allow only light of the required wavelength to reach the cells. These are phototransistors and have peak spectral response in the near infra-red region, 1·5μ.

The useful frequency range of the equipment is limited by two factors. The first is the frequency response of the phototransistors, the cut-off frequency being quoted by the makers as 3kc/s. However, since the only effect of reducing the amplitude of the signals is to reduce the diameter of the circular trace, while the angle which a given arc subtends remains constant, the equipment can be used at frequencies above this value. The second factor is the effect of resonance in the spring-strip suspension of the moving grating. This trouble can be overcome in a number of ways such as adjusting the stiffness of the suspension to place its resonance outside the required frequency range. Alternatively, the suspension can be dispensed with altogether and the moving grating fixed directly to the vibration source or to a rigid extension fixed to the source. In this case the vibration must be restricted to motion in a straight line normal to the lines on the grating since components in other directions will give rise to errors.

Dynamic calibration is unnecessary since the accuracy of measurement depends only on the accuracy of the gratings and is independent of fluctuations in mains voltage.

The instrument can also be used to measure small displacements.

A Miniaturized Radio Telephone Terminal Equipment

By E. J. Allen*

A radio telephone terminal equipment is described which although only a fraction of the size of conventional equipment provides a comparable performance and provides for the simultaneous operation of up to four commercial grade speech channels.

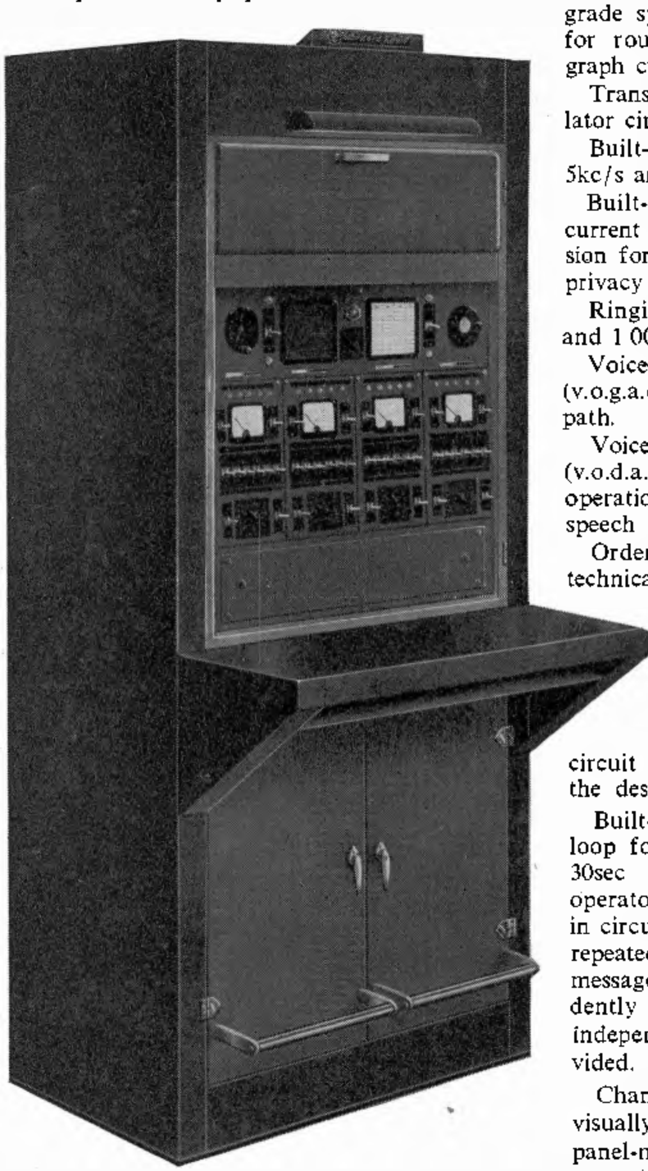
(Voir page 313 pour le résumé en français: Zusammenfassung in deutscher Sprache Seite 318)

PROSPECTIVE users of short-haul independent-sideband radio telephone links have in many cases found that the disproportionate size and cost of the associated radio telephone terminating and band-displacement equipment necessary to complete the connexion into the local telephone exchange system, with or without privacy facilities, has proved a serious stumbling block.

When the cost of a small independent-sideband transmitter and associated receiver capable of handling four commercial speech channels in a 12kc/s bandwidth is contemplated, it is something of a shock to find that the terminal and associated equipment for putting these communications to line will absorb a further sum more than double this initial amount, and that space for four 10ft 6in and three 9ft double-sided standard telephone type racks must be found.

The type TOP.20 radio telephone terminal has been introduced by Standard Telephones and Cables Limited to overcome these disadvantages. It occupies no more space than the familiar floor-standing private branch exchange, being 75in high, 33in wide and 19in deep, the maximum weight of a fully equipped terminal being approximately 700lb. The cost is about half that for the conventional equipment used previously.

While reference is made by the manufacturers to its suitability for the small link they emphasize that it is eminently suitable for larger and more ambitious systems, and was designed to replace entirely the older type of equipment in all types of i.s.b. radio telephone system.



The TOP.20 radio telephone terminal

The fully equipped TOP.20 incorporates the full facilities available with older equipments and some additional advantages, i.e.:—

Simultaneous operation of up to four commercial-grade speech channels with provision for routing of voice-frequency telegraph circuits.

Transistorized control and modulator circuits.

Built-in band displacement for both 5kc/s and 6kc/s systems.

Built-in invertor to work with all current inversion systems, with provision for use with current 'scrambler' privacy systems.

Ringing, send and receive, at 500c/s and 1 000c/s.

Voice operated gain adjustor device (v.o.g.a.d.) in each transmit and receive path.

Voice operated device anti-singing (v.o.d.a.s.) to prevent singing and false operation due to noise, with minimum speech clipping.

Order-wire facilities to enable the technical operator to communicate with the transmitting and receiving sites and also with the switchboard operator.

Fully automatic operation with no routine adjustment.

A high degree of long-term circuit stability is a major factor in the design.

Built-in record unit using a tape-loop for storage of a message up to 30sec in duration. The technical operator can record his speech while in circuit, and the information can be repeated as long as required. A message can be recorded independently for each speech channel, and independent erase facilities are provided.

Channel monitoring on two circuits visually by V.U. meter and aurally by panel-mounted loudspeaker or by operator's headset.

Extensive use of strip-units is made, each being quickly removed and replaced for servicing, which can be carried out externally with the assistance of a test set described later in this article. Space is saved by the use, throughout, of miniaturization techniques.

* *Standard Telephones and Cables Ltd.*

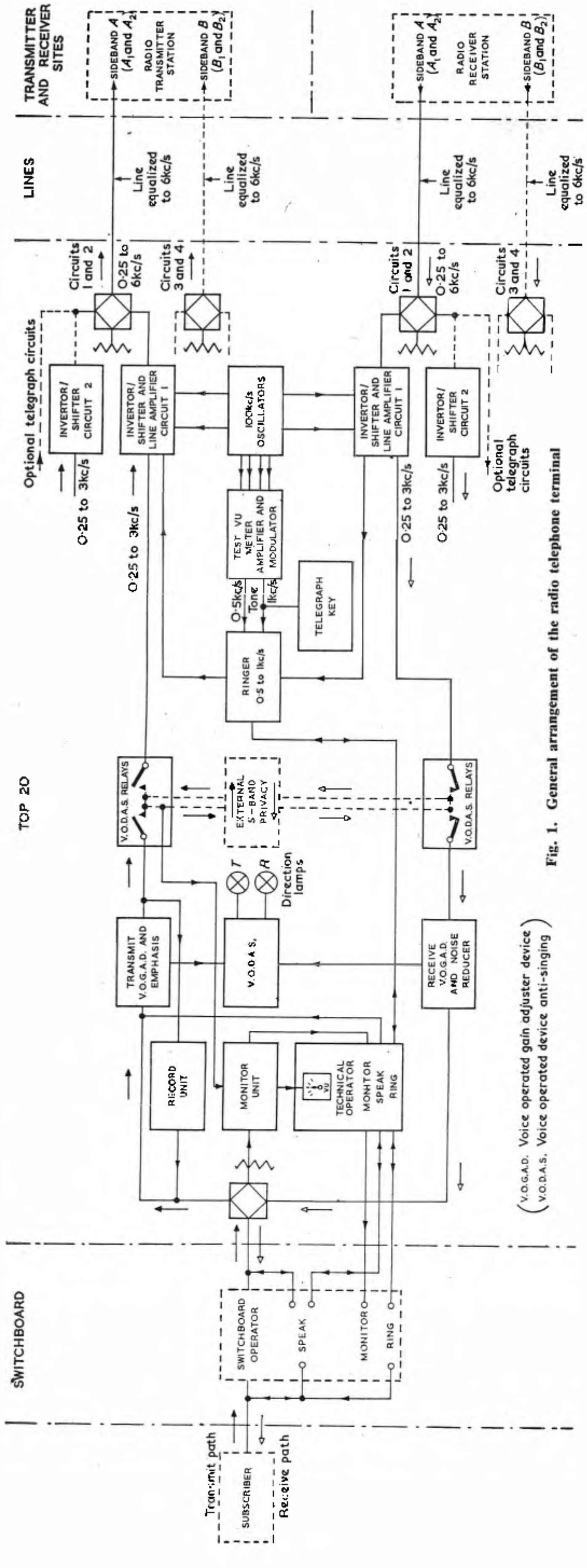


Fig. 1. General arrangement of the radio telephone terminal

The internal framework of the cabinet is of channel section carrying six cross-members. These in turn carry vertically mounted and easily detached strip units, each connected by plugs and sockets. A cable duct is provided at each side of the cabinet and entry may be at the base or top. Up to four control panels plus an intercommunicating panel and test panel are accommodated on a hinged housing on the front of the cabinet. This may be swung out to give access to strip units mounted behind. A detachable writing-shelf is fitted to the cabinet at a convenient position below the control panels. Below this shelf there are double doors giving access to fuse panels and strip units in the lower front section of the cabinet. At the rear are two full-length doors giving full access to all strip units mounted on the rear of the framework and the power supply equipment in the base of the cabinet. Both sets of doors are dust-sealed, those at the rear carrying air filters.

All cabinets are supplied fully wired, but need not be fully equipped initially. When expansion is necessary, additional units can quickly be fitted by the user, on site. The minimum equipment available as standard is for two speech channels. A single-channel version is not economical in this equipment; but for this specific purpose, where extreme economy is a factor and expansion is not foreseen, a single-channel desk or bench-mounting unit called the type TOP.21 is available.

One internationally-recognized valve type is used throughout, the Brimar Trustworthy type 6057, a direct equivalent to the U.S.A. ruggedized type of the same number. Both are derived from the prototype 12AX7 which may be used where the 6057 may not be readily available. While the circuits will work with normal efficiency there will be some sacrifice of reliability with the prototype valve. A total of 82 valves is used in the fully equipped TOP.20.

Valve heaters and a reduced h.t. supply are left on during breaks in normal operating schedules to ensure stable operating conditions and an enhanced valve and component life. The total power consumption for the full equipment is less than 500W.

Silicon power rectifiers are used, and the regulation is such that no significant change in output voltages occurs from full to minimum load. Input mains regulation, for maintenance of performance, however, should not exceed ±5 per cent in voltage. For variations outside these limits, an external automatic voltage regulator will be required.

Typical Performance

Bandwidth

The equipment is designed to handle voice frequencies between 250c/s and 3 000c/s and to eliminate frequencies not containing components of intelligence. The normal bandwidth on each speech circuit is 250c/s to 3 000c/s, but use of the narrow-band filter incorporated will limit the upper figure to 2 450c/s.

Voice Operated Device Anti-Singing

The signal-to-noise ratio from the radio path or from the subscriber should be at least 25dB for a good speech circuit, but satisfactory v.o.d.a.s. operation can be obtained with a signal-to-noise ratio as low as 10dB.

Speech Levels

Input from two-wire line (acceptable range)	−37 to +10 V.U.
Input to transmit v.o.g.a.d.	−40 to +7 V.U.
Output from transmit v.o.g.a.d. (+2 V.U.)	0 V.U.

(External privacy may be inserted at this point)*

Input to invertor-shifter	0 V.U.
Output from invertor-shifter (adjustable)	−2 to +13 V.U.
Output to radio transmitter line (adjustable)	−5 to +10 V.U.
Input from radio receiver line (acceptable range)	−26 to +5 V.U.
Input to invertor-shifter	−29 to +2 V.U.
Output from invertor-shifter	0 V.U.
(External privacy may be inserted at this point)*	
Input to receive v.o.g.a.d.	0 V.U.
Input to receive v.o.g.a.d. (acceptable range during fading)	−17 to +10 V.U.
Output from receive v.o.g.a.d. (±2 V.U.)	−2 V.U.
Output to two-wire line	−5 V.U.

EMPHASIS

Speech emphasis, giving as much as 18dB increase in level at 3 000c/s as compared with that at 500c/s is employed in the transmit v.o.g.a.d. in order to enhance speech energy at the higher frequencies and hence provide increased reliability of v.o.d.a.s. operation.

POWER REQUIREMENTS

200 to 260V, 40 to 60c/s single-phase. Voltage variation not exceeding ±5 per cent. Power consumption less than 500W for the fully equipped cabinet.

* *When an external scrambler system is used, input and output levels will be 0 V.U. at the insertion points mentioned above.*

Test Equipment

With the conventional type of terminal and allied equipment, operators will be only too aware of the annoyance of having to wait until a particular routine is finished before a bay can be taken out of service for testing a faulty or repaired unit.

The type of construction embodied in the TOP.20 equipment was expressly adopted to render the utmost facility in rapid servicing, and the test equipment makes possible the fullest use of this technique.

While routine testing facilities are incorporated in the TOP.20, use of this test equipment enables any unit to be worked on and tested under an exact simulation of working conditions independently of the TOP.20, which can remain in service while testing is being carried out.

Power supplies, the necessary signals and V.U. meter are provided for the unit under test, and provision is made for relay, ringer and v.o.d.a.s. switching tests with synthetic speech for v.o.g.a.d. testing. Only one such test equipment would be necessary on a station, irrespective of the number of TOP.20's installed.

Acknowledgment

The author wishes to thank Standard Telephones and Cables Ltd. for permission to publish this article.

An Electrophysiological Amplifier for Students' Use

By P. E. K. Donaldson*, M.A.

A compact, single-sided amplifier is described for action-potential recording from excised nerve or muscle. The apparatus is entirely mains-driven, and it is felt that in this respect, and in view of its simplicity, it may be of interest to other biological laboratories.

(Voir page 313 pour le résumé en français: Zusammenfassung in deutscher Sprache Seite 318)

IN leading off action-potentials from excised excitable tissue, the greatest difficulty is encountered (but often the most significant results obtained) when recording from small nerves, or even single fibres. For here the signal is smallest—tens of microvolts—and the preparation resistance highest—hundreds of kilohms. The signal-to-Johnson noise ratio is therefore at its worst. The situation is further aggravated because, under conditions of high preparation resistance, electrostatic interference is at its most troublesome. To exclude, as far as possible, hum and mains-borne transients from the eventual record, high-gain biological amplifiers have, at least in this laboratory, generally been battery-driven.

It is useful if action-potentials can be displayed on an oscilloscope, so that the student may examine their form, and simultaneously played through a loudspeaker, so that he may be aware of their presence when the eyes are directed elsewhere. With valves of second world war and earlier vintages, acoustical feedback was practically inevitable unless the loudspeaker and the first stage valve or valves were widely separated.

As a result of these considerations, apparatus for recording from nerves has comprised a number of separate items, namely:

(1) Pre-amplifier.

(2) Post-amplifier, with loudspeaker.

(3) H.T. batteries.

(4) Accumulators.

While it is certainly true that battery-operated gear is simpler in principle than mains-driven, it is doubtful if it is as convenient in practice. Accumulators demand a proper maintenance routine, and dry batteries need careful checking for incipient noise-production. Further, apparatus comprising a number of units suffers wear and tear at the

* *The Physiological Laboratory, Cambridge.*

connectors and flexible leads in being repeatedly set up at the beginning of, and taken apart at the end of, practical classes.

There seemed, therefore, to be a good case for investigating the feasibility of a small, self-contained, mains-driven amplifier, if this could be achieved without great additional complication and expense. The equipment described here, of which five specimens have been constructed, is the first attempt. Its size can be judged from Fig. 1. That it does not 'howl back' is attributable to the remarkable freedom from microphony of modern braced-structure valves. Absence of hum and noise may be judged from Fig. 2(a), taken with the input short-circuited. The two small 'pips' are 10μV calibration marks. That there is very little 50c/s pick-up with a high impedance signal source may be verified from Fig. 2(b), taken at the same gain; the dummy preparation was a 1MΩ, 0·5W, carbon composition resistor. Fig. 2(c) shows a 50c/s timing wave, taken at the same sweep speed.

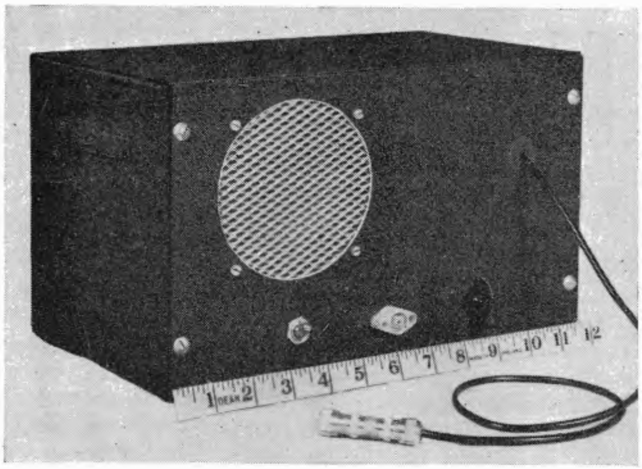


Fig. 1. A small, self-contained, mains-driven amplifier

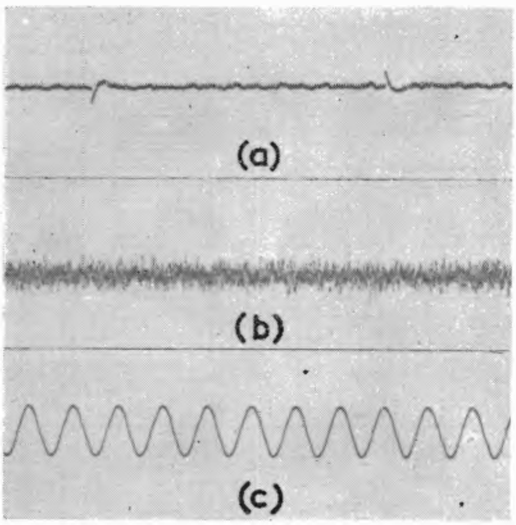


Fig. 2. Output waveforms of the amplifier

The circuit of the amplifier is shown in Fig. 3, and is quite straightforward. The heaters of the first two stages are supplied with direct current. Half-wave rectification proved adequate both here and in the h.t. supply. The latter includes no choke; hum transferred to the 5in loudspeaker, as a result of returning the top of the speaker transformer primary direct to the rectifier input capacitor, is quite imperceptible. The first two stages are enclosed in a die-cast box measuring 4½in by 3½in by 2¼in. This is sandwiched between two foam plastic sponges; the resilient mounting so formed is helpful in combating acoustical feedback, and renders the equipment insensitive to general vibration. Box and sponges can be adjusted about a vertical axis. In this manner, hum induced into the grid circuit of the first stage, by stray magnetic field from the mains transformer, can be made to pass through a sharp null. The input terminals should be short-circuited when carrying out this process. The mains transformer used is an unshrouded type having fully interleaved laminations.

It is essential that a separate wire be used to carry the negative heater supply to the first two stages; any attempt to combine the heater negative and h.t. negative supplies to these stages into one lead, or worse, to eliminate the lead altogether and use the chassis instead, will cause serious hum. The heater circuit as a whole should be earthed at the valveholder of the first stage, and nowhere else. This implies that it is advisable, if the heater supply smoothing electrolytics are of the metal-can type, to interpose some insulation between the can and the mounting clip.

No heat sink is necessary for the GJ3M rectifier.

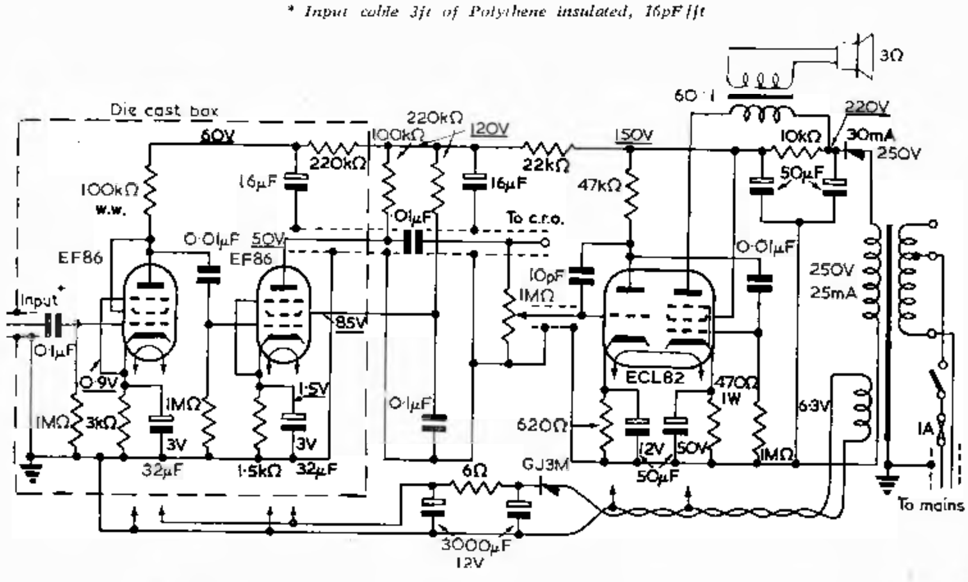


Fig. 3. Amplifier circuit
* *Input cable 3ft of Polythene insulated, 16pF/ft*

Conclusion

A need has been suggested for a compact, mains-driven biological amplifier. A design has been produced which has proved stable and repeatable. In the interests of simplicity and cheapness, a number of rules of 'good practice' have been transgressed; the performance is nevertheless satisfactory.

THE ELECTRONIC ENGINEERING ASSOCIATION

THE Electronic Engineering Association, whose headquarters is at 11, Green Street, London W.1, was formed in 1944 under its previous name of the Radio Communication and Electronic Engineering Association by a number of prominent manufacturers of electronic equipment for aeronautical use who had found that they had several common problems which might be resolved collectively. Since that time the Association has expanded considerably and now as a consequence covers aviation, data processing, marine and mobile radio uses of electronic equipment. In keeping with the times it is likely that a further section will be set up shortly to deal with electronics in most industrial applications.

The membership of the Association today embraces virtually all the leading manufacturers of electronic equipment in the United Kingdom. Provision has, however, been recently made whereby the smaller concerns who have only a limited interest in the Association's field of activities may join the Association in that section in which they are most interested.

Very briefly the objects of the Association are to promote the interests of its members in connexion with the manufacturing and marketing of those items of electronic equipment in which the Association interests itself and to promote research in connexion therewith. It also endeavours to maintain a high standard of quality, design and workmanship in the industry and gives advice on and otherwise deals with manufacturing problems and promotes the standardization of electronic equipment where this is desirable.

The Association also exists to assist manufacturers and others having commercial relationships with the industry and to give them advice and assistance as required. Further objects are to institute, promote and support measures improving the standard of employment in the industry, and to further the education and training of engineers and technicians for the industry.

Principal Functions of the Association

The Association's activities may be listed under two main headings:

(1) Commercial

(2) Technical.

Its functions in the commercial field are strictly limited by the wish of its members, that is to say, it does not exist as a price-fixing trade association and its activities do not come within the mischief of the Restrictive Trade Practices Act. Its value in this sphere lies rather in that it serves as a focal point whereby Government Departments such as the Ministry of Supply or Board of Trade may reach the capital goods side of the electronics industry when either of them may be aware of export business which is offering. Similarly it allows Government Departments to obtain from this part of industry a collective view on any particular commercial problem with which they may be faced from time to time. The Association is of course also able to keep its members in touch with the movements of foreign delegations and with overseas exhibitions, with a view to exploiting the export possibilities of the industry. The Association does interest itself in the contract conditions of Government and other large bodies, but it has no contract conditions of its own and has no power to force its members to accept its views on any of these matters. Rather it confines itself to the giving of advice and assistance in this sphere when required. So much for the functions of the Association in the commercial field. While these services are by no means insignificant, they do not in any way stand comparison with the services which are provided on the technical side. Here the Association is

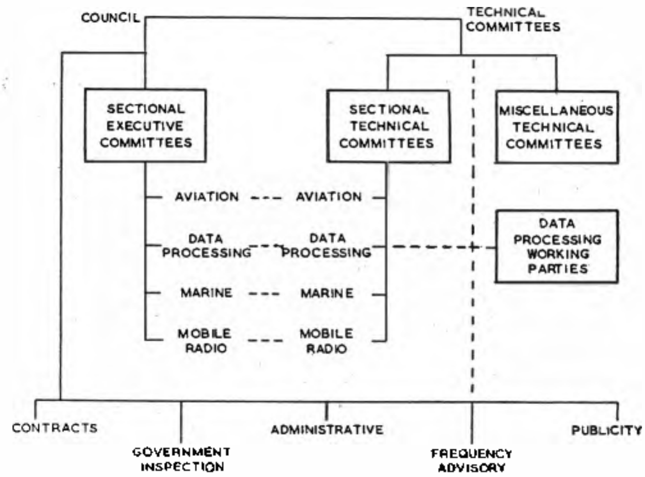


Fig. 1. E.E.A. main committee structure

undoubtedly of immense value to its members, partly through the promotion of standardization through the good offices of the British Standards Institution and partly by providing a forum for discussion on the manifold technical problems which arise within its member companies.

Perhaps at this stage it would be best to give some details of the actual structure of the body, through which its method of operation will be more clearly understood.

Structure of the Association

The main committee structure can be seen from Fig. 1. The controlling body is the Council, which is the only unit of the Association which has executive authority. All the sections and committees which may from time to time take some action do so on authority expressly delegated by the Council. For instance, the Aviation or Marine Executive Committees may initiate action in their respective fields as occasions demand, but they only do this on specific authority delegated by the Council.

Similarly, on the technical side, the Council operates through its Technical Committee. This in turn exercises control over the sectional technical committees and the miscellaneous technical committees.

The sectional technical committees have a dual responsibility. They report to their respective executive committees on matters requiring executive action, but they also report to the Technical Committee in order that technical co-ordination in all spheres of the Association's work may be achieved. Subsequently, executive committees report to the Council, which may take executive action as required.

In addition to controlling the sectional executive committees there are a number of direct Council committees as shown in the diagram. Some of these will be dealt with later, but perhaps a word or two concerning that dealing with Government Inspection, and also the Frequency Advisory Committee might not be out of place here.

The Government Inspection Working Party has been of great value to the Association whose members have to submit their equipment supplied to Government Departments or the Services to stringent Government Inspection. The Inspection service is controlled by the Ministry of Supply, and the Association is happy to record that its Inspection Working Party has been able to establish a most satisfactory forum for co-ordination with the Ministry on these matters.

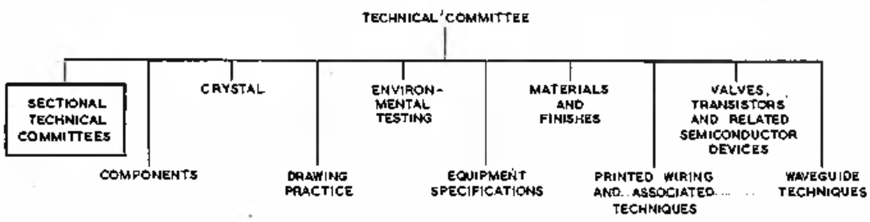


Fig. 2. E.E.A. technical committee structure

The Frequency Advisory Committee is another case in which it has been found necessary to adopt the dual reporting system. The matters it considers are very largely of political concern to the industry, and consequently for this purpose it reports directly to the Council. On the other hand there are frequently matters of a technical nature which need co-ordinating with the Association sections and here use is made of its direct link with the Technical Committee. The Postmaster-General also has a Frequency Advisory Committee to which he has appointed members whose names were put up by the Association's body, thereby enabling the Association to keep in close touch with developments in this field.

The Council's administrative committees are, of course, primarily concerned with the day-to-day functioning of the Association.

Its Work in the Commercial Field

It has been mentioned earlier that the Association has links with the Board of Trade and with the Ministry of Supply. Its connexions with Government Departments do not end there however. It has the closest of ties with the Postmaster-General's Department on matters relating to radio communication, and of course by the very nature of its members' business it keeps in close touch with the Service Departments.

It should be stressed that its liaison with these departments is far from being only on the level of manufacturer and customer. There are questions of type-approvals for equipment, technical advice to the Postmaster-General, assistance with standardization, representation at international conferences and a myriad of similar subjects where the views of industry must be sought and provided.

Prior reference has been made to the Association's interest in contractual matters. It limits itself here to giving advice to its members but its Contracts Committee, composed of experts in this type of work, is always ready to assist members with any problem on contracts which they may have on hand. In this sphere of action, the Committee maintains very close liaison with the Federation of British Industries and it has recently set up means whereby it can exchange ideas with other bodies such as the British Radio Valve Manufacturers' Association and the Radio and Electronic Component Manufacturers' Federation.

The Publicity Committee is maintained in order to publicize the capital goods side of the British electronics industry rather than to advertise the Association itself. It has before it certain projects for putting into effect world-wide publicity for the industry on a collective basis and it is expected that more will be heard of this later. This Committee has other routine functions. It advises members on the desirability of participating in exhibitions both at home and overseas. The Association is one of those which are sponsoring the Instruments, Electronics and Automation Exhibition. The Committee gives guidance here. Its help was most valuable for the staging of the Electronic Computer Exhibition which was held last year. This was the first of its kind to be staged in the United Kingdom and was judged to have had a considerable impact. For the last two years the Committee has, through the courtesy of the Society of British Aircraft Constructors, organized a press conference during the Farnborough Air Show, covering electronics as applied to aviation. It also arranges an annual press conference at the time of the Association's Annual General Meeting.

Little need be said about the sectional executive committees. Their names indicate clearly their respective spheres of action. The work in each section varies from time to time depending on the existing need for collective action, and it may stem equally from the technical committee or the executive committee.

Its Work in the Technical Field

A guide to the subjects covered on the technical side may be obtained from Fig. 2 which illustrates the technical committee structure. With regard to the sectional technical committees it should not be thought that these only work on subjects generated within themselves. Each technical committee is represented on one or more outside bodies, Governmental or otherwise, and it is through these that the views of the capital goods side of the electronics industry are disseminated. For instance the Aviation Section is represented on the Ministry of Transport's Civil Aviation Radio Advisory Committee. It has recently acquired representation on the newly formed Air Transport Electronics Council. Here it will have effective liaison with the civil aircraft operators and with the aircraft manufacturers.

The Marine Section is similarly represented on two committees of the Ministry of Transport and Civil Aviation. These are the Radio Aids to Marine Navigation Applications Committee and the Ships' Wireless Working Party. Subjects dealt with vary from the Convention of Safety of Life at Sea to the scale of a marine radar p.p.i. presentation.

The Mobile Radio Section works closely with the Postmaster-General's Mobile Radio Committee. Here such things are dealt with as channel spacing to be adopted for mobile radio communication.

In the sections, the Data Processing Section has produced by far the largest technical contribution during the past two years, and Fig. 3 which shows the Data Processing

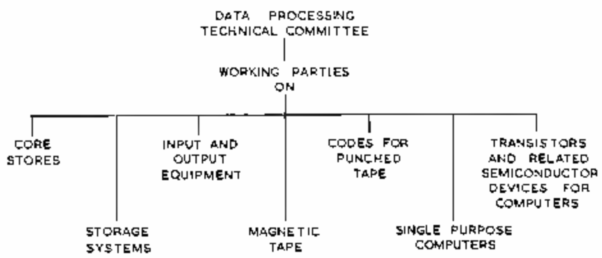


Fig. 3. E.E.A. data processing technical committee structure

Technical Committee structure shows the direction in which it has been proceeding. It is natural that this Section should at the moment have a considerable amount of technical work confronting it since it is working in a comparatively new sphere and there are many novel problems facing it, particularly in connexion with standardization. Good progress in this direction is being achieved. Since the Association as such does not issue any standards of its own to its members, those prepared by the Data Processing Section are passed to the British Standards Institution, where additional co-ordination with users takes place.

All the technical committees and working parties make a considerable contribution to the work undertaken by the B.S.I. and the Association is represented on upwards of 125 B.S.I. committees. The technical effort thus involved is undoubtedly a heavy drain on the technical resources of member companies, but there is no doubt that the work performed is both necessary and fruitful.

In the same way there is technical representation on a number of committees under the aegis of the Institution of Electrical Engineers. The Association also appoints technical representatives to sit on various Post Office committees, particularly where these are formulating specifications for radio communication equipment. In this case any of the appropriate sections may be involved.

Close touch is also maintained with various Government bodies which are interested in the electronics field, and there is representation on the Joint Electronics Standardization Committee and on the Radio Components Standardization Committee, both of the Ministry of Supply.

There is, of course, a necessity to have liaison with a number of other associations with direct interest in the capital goods electronics field. Consequently there are joint committees and means for consultation with bodies such as B.R.V.M.A. and R.E.C.M.F.

Electronics and in particular the application of computer techniques are spreading into many other industries. A means for technical consultation with these industries has been found in the Electronic Forum for Industry where contact is maintained with such bodies as the Machine Tool Trades Association, the Office Appliance and Business Equipment Trades Association, the Scientific Instrument Manufacturers' Association, the Society of British Aircraft Constructors and the Telecommunication Engineering and Manufacturing Association.

The Data Processing Section of E.E.A. was the main instigator in the setting up of this Electronic Forum for Industry, which is working only in the technical field with the other associations mentioned above. It is expected that the Forum may lead to the rapid extension of the use of computers throughout industry. It is as yet in its infancy.

Technical Education and Training

Everyone working in this field must be aware of the great importance which must be attached to technical education and training for the electronic industry. Since the recently announced re-organization of representation within the industry, the Electronic Engineering Association has agreed to assume responsibility for the major part of this work. Discussions are currently taking place with a view to seeing how these matters can best be handled. It is possible that a joint organization may be set up embracing the electrical, electronics, and telecommunication industries, whose educational problems have many common factors and whose technical training is very often closely linked. It is the intention that a new organization to deal with these subjects efficiently will be established with the least delay.

An Experimental Radio Station on Ascension Island

The Department of Scientific and Industrial Research, with the co-operation of Cable and Wireless Limited, will in June set up an experimental radio station at Ascension Island, from which two of their staff will conduct test transmissions to their Radio Research Station at Slough for a year. The object of the tests will be to investigate radio wave propagation over long distances.

The D.S.I.R. Officer in charge of the project at Slough is A. F. Wilkins, who was associated with Sir Robert Watson Watt in the development of radar. The two officers going to Ascension are L. A. Brackstone and R. Aria. On Cable and Wireless Limited's side, arrangements have been made on behalf of the Engineer-in-Chief by R. J. Hitchcock, head of the section dealing with radio propagation, frequencies and antennae. He is a member of the Ionospheric Communications Committee of the Radio Research Board.

A receiver and transmitter have been developed at Slough, which although separated by great distances, may yet remain accurately in tune with one another while the radio frequency is automatically changed over a very wide band. On any radio circuit conditions on one frequency can be very different from conditions on another, and it is this difference which tells propagation engineers so much.

This particular equipment is capable of sending and receiving signals on nearly 2 000 different frequencies between 5·5Mc/s and 50Mc/s in the space of about 15 minutes. Only the transmitter will go to Ascension. The receiver will remain at Slough, where the signals will be observed and analysed.

Antenna design is largely dictated by propagation considerations, and an interesting series of tests is planned to see the effects of shooting into the ionosphere at different angles. Cable and Wireless have always wanted to know the best angle to the horizon, at which they should transmit radio energy on long radio routes, and all the indications at present are that this angle should be very low. In this experiment three dipoles at different heights above ground are to be erected on a 150ft mast. By placing this mast as near as possible to the water's edge and transmitting from each dipole in turn, it is possible to alter the shape of the transmitted energy to be in the vertical plane and see what effect this has on the received signal at Slough. It is believed that this is the first time a controlled experiment of this nature has been made. In addition, the Cable and Wireless rhombic antennas at Ascension Island will be used to test the effects of good and poor sites as represented by having sea in front of one antenna and rock in front of the other.

A NEW TYPE OF RING COUNTER

By P. J. Westoby*

There are many occasions where a multi-stage ring counter is called for and the number of stages in the more usual schemes has been limited to 12, or so, for reasons mentioned. The article describes a system whereby any number of stages may be employed without the disadvantages met in the previous schemes. The method of working is explained and simple design considerations are indicated.

(Voir page 313 pour le résumé en français; Zusammenfassung in deutscher Sprache Seite 318)

RECENTLY many schemes have been evolved for transmitting intelligence, originating from many sources, through one channel. These schemes fall mainly into two types, time sharing and frequency sharing systems. Within the compass of the former fall such things as the passing of many private telephone conversations over a single pair of wires; or in telemetry where several channels of information regarding test vehicle behaviour may be transmitted to the ground over a single radio channel. To these two may be added the fairly common multi-beam oscilloscope that relies on an electronic switch for apparently obtaining many separate traces.

All the above mentioned schemes require some form of mechanical or electronic switching, or sampling, of the many information sources. If done mechanically the upper limit of the rate of sampling is not very high because of the mechanical unreliability of fast moving parts. An exception is, of course, in the case of a mechanical switch that has only to run for a limited period, as in a telemetered rocket test vehicle. In this case it can be made to work at a relatively high speed. However, as the majority of electronic equipment must be reliable over long periods the need for some form of electronic sampling is felt.

This may take the form of a neon counter device or a series of valves in the form of a ring. The advantage of the neon counter is its compactness and simplicity, while its disadvantages are two-fold. Firstly, the upper frequency limit is set by the de-ionization of the gas, and secondly the number of stages is limited to around 12 unless specially developed tubes are used. Two or more such tubes could be used in conjunction with electronic coincidence gates, to obtain a greater effective number of stages than one tube alone; but the upper frequency limit still holds, and some simplicity is lost.

The general scheme using a series of valves usually consists of the valves arranged in the form of a 'ring'. The circuit is so arranged that only one valve within the 'ring' can be in a state of conduction different from the remainder. Each of the ring valves is connected to a gate, and at the transmitter each gate controls the individual information channels to a common line, i.e. the transmitter channel. At the receiver, every gate has a common input and a separate output for feeding the information into the separate channels again. Some form of synchronization and drive is also employed to ensure the gates open and close at the correct time. This scheme is shown diagrammatically in Fig. 1.

Valves in the form of a 'ring' may be used for simple dividing but as the number of stages is increased, the use of 'ring' type dividing circuits for simple division becomes increasingly inefficient compared with the binary, or scale-of-two scheme, to produce a given division. This is best illustrated by considering two practical cases as shown below:

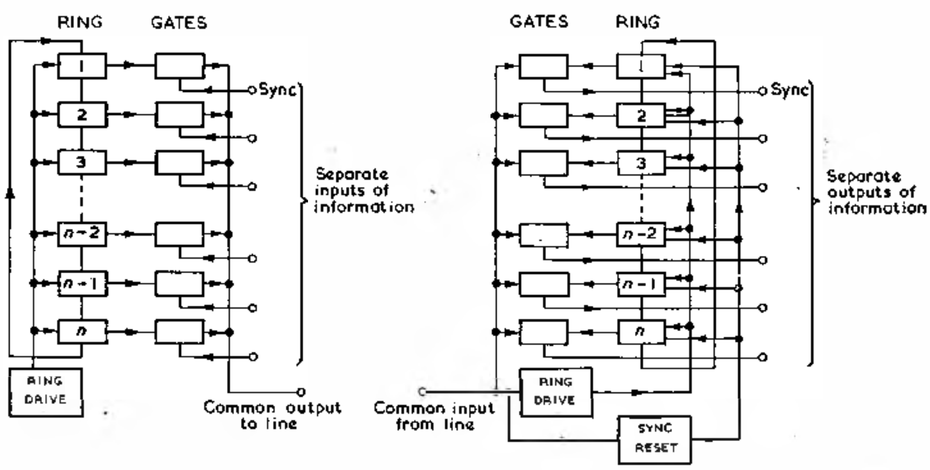


Fig. 1. General arrangement of ring counters and gates

Division Required	*Scale-of-Two Stages*	*Ring Stages*
5	3	5
500	9	500

In consequence, the main use for 'rings' of large numbers is found in the field of electronic switching as already described.

Up to now there have been two types of rings commonly used, the first type, designed by Lewis[1], and the second described by Sharples[2]. The main limitation of the Lewis scheme is that of a practical nature.

The circuit is similar to an Eccles-Jordan circuit in that one conducting valve is able to maintain the other valves in an 'off' condition. This is done by connecting every anode to every other valves grid via $(n-1)$ 'averaging resistors', where n is the number of stages in the system. These resistors load the anodes and so limit the upper frequency response of the system. To reduce this effect, high mutual conductance valves and low anode loads can be used, but even so an upper limit of about 5 or 6 stages is the maximum attainable in practice before the signal at the anode is insufficient in amplitude to turn the next stage over. In addition, the practical consideration of wiring such a complex scheme becomes

* *Rank Cintel Ltd.*

formidable especially when lead lengths are kept to a minimum. A circuit indicating the scheme is shown in Fig. 2.

An improvement on the above scheme is afforded by the second type of ring circuit mentioned above. This is the scheme that has been used in the American computer ENIAC. It consists of n stages of an Eccles-Jordan type circuit with all the corresponding cathodes connected together to two common resistors. The ratio of these two resistors is adjusted to be R and $R/(n-1)$. This scheme is shown in Fig. 3.

The resistor $R/(n-1)$ has $(n-1)$ valve currents passing through it, while the larger resistor has only one valve accuracy is of paramount importance. This problem is aggravated by the fact that slight variations in valve constants produce the incorrect potential to enable only one stage to be 'dissimilar', in consequence of which more than one stage can be in the 'dissimilar' state. When this occurs there is no simple way of assuring that the correct state of conductions has been attained.

It was because of this difficulty that when it was required to make a ring of 24 the above scheme was useless.

The new scheme consists of 24 entirely separate Eccles-Jordan circuits. These are connected in cascade with the anode of one stage feeding into the grid of the next, and so on to complete the ring. 23 of the 24 stages are in

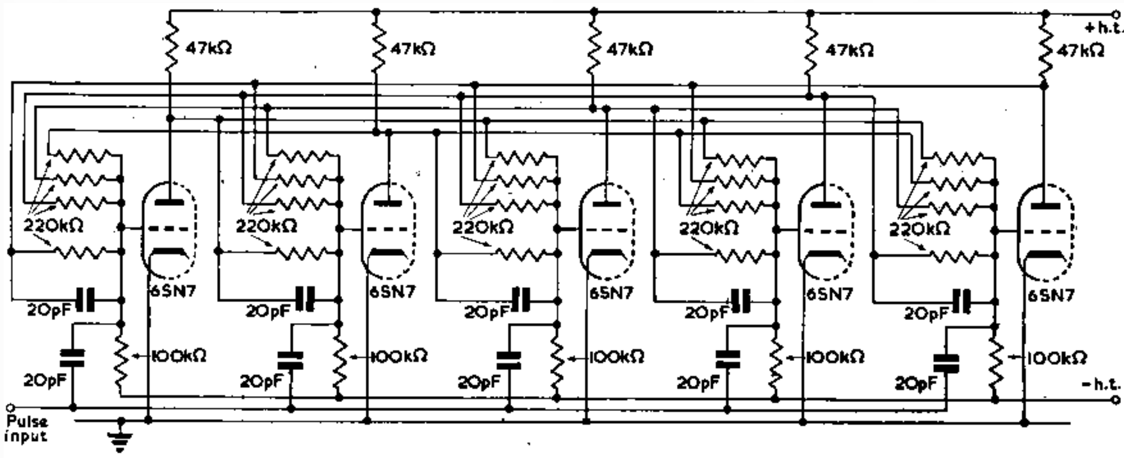


Fig. 2. Lewis five-stage ring counter

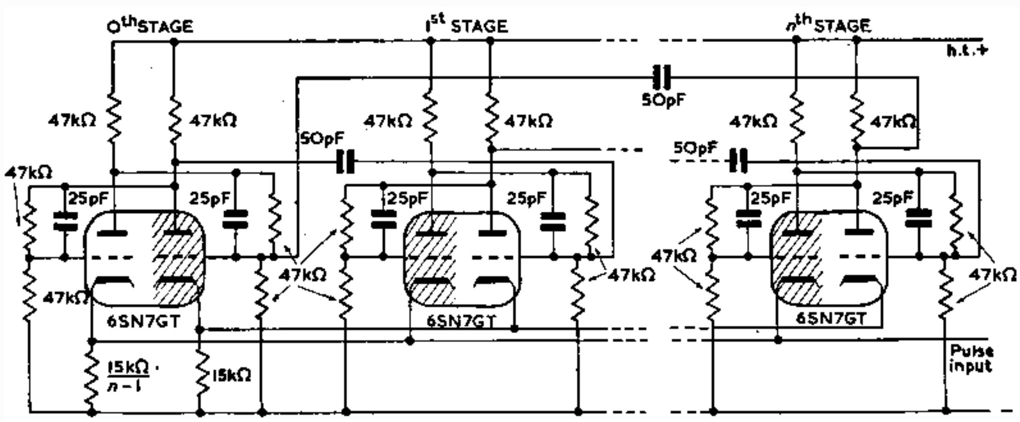


Fig. 3. The ring circuit used in ENIAC

current passing through it, all the other valves connected to it being cut off; in this way all cathode potentials are equal. If another of the non-conducting valves becomes conducting its cathode potential would rise, and the originally conducting one would then become cut off. Consequently only one stage can be dissimilar from the remainder.

A negative drive pulse is applied to the smaller cathode resistor which turns on the half of the dissimilar stage that was off. This by trigger action turns off the other half and in so doing, feeds a positive signal to the off half of the next stage. Thus the dissimilarity has been moved along one stage.

The scheme is quite satisfactory up to about 12 to 15 stages, but for larger number of stages, the value of $R/(n-1)$ becomes so small compared with R that its

a similar state of conduction while the 24th stage is in the opposite state. Fig. 4 indicates this by shading the halves of each stage that are in a state of conduction. Now the drive pulses, required to 'move the ring around', are of a negative polarity and are applied through diodes to corresponding grids on each stage. The function of these thermionic diodes will be discussed later. As the drive pulse is negative, it can only affect the conduction of an 'on' valve, and the only grid with drive pulses applied to it in this state is the dissimilar stage V_{1a}. Consequently, this stage is turned over. In doing so a positive going step voltage appears at V_{1a} anode, while a corresponding negative going waveform appears at V_{1b} anode. This negative going step is differentiated to produce a negative pulse that is passed along to the next stage via a crystal diode. The pulse is applied to the conducting half of V_2 i.e. V_{2b}, and so this

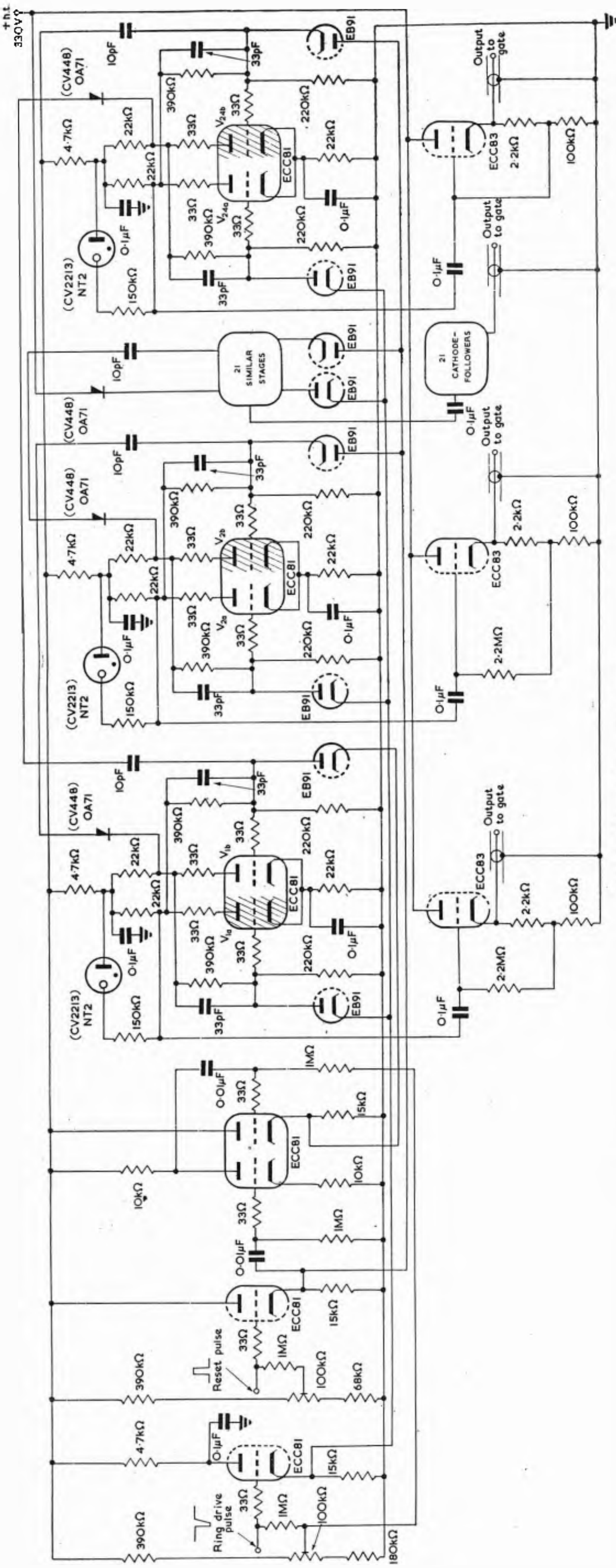


Fig. 4. Channel separator ring

stage is turned over. When this occurs a positive step appears at V_{2b} anode, but due to the series crystal diode, nothing is passed on to the next stage and hence the action stops. At first sight the crystal diode need not be used, as the succeeding stage can only respond to a negative pulse, but in practice it was found more reliable to include it. V_2 stage is now in the state previously held by V_1 and so, on receipt of the next negative drive pulse V_2 stage is the only stage that can respond. In this way, each stage in turn becomes dissimilar from the others on receipt of a drive pulse and as the 'last' is connected to the 'first', the ring is made to recycle every 24 input drive pulses.

The drive pulses are, as already mentioned, fed through thermionic diodes. These diodes are biased to such a potential that they only conduct when a drive pulse is applied. The potential at the grid of a conducting valve in the Eccles-Jordan circuit used is 105V, and so the diode cathode potential is arranged to be about 110V, in this way the source impedance of the drive pulses does not affect the operation of the trigger stages. In the equipment designed, the biasing potential was derived from a cathode-follower through which the drive pulses were also fed. This scheme allows quite a large stray capacitance to appear between the diode cathode and earth before the pulses are reduced sufficiently in amplitude to prevent them from turning the Eccles-Jordans over; at the same time it provides an easy way of adjusting the diode biasing potential. It may be noted in passing that when the stage is in the other state of conduction the grid to which the drive pulse is applied is at a lower potential of 70V and so the diode biasing is still satisfactory.

When the equipment is first switched on, each Eccles-Jordan stage may settle down into either of its two stable states, i.e. several of the stages may be 'dissimilar'. Consequently, a resetting pulse must be applied at the commencement of the operation. In practice, a periodic resetting pulse is applied each time the ring recycles This resetting pulse is applied to the opposite grid from which the drive pulse is applied, and, as it must bring this valve of each stage into a state of conduction, it must be of a positive polarity. In the case of the dissimilar stage, however, the opposite holds true and the 'reset' side of the trigger stage must not conduct. To achieve this, therefore, the resetting pulse is phase reversed and applied as a negative pulse to the grid. When the resetting pulse is positive, as on 23 stages, the diode, employed for reasons given above, must have its anode held at a lower potential than the grid to which is is connected. The grid potential of a non-conducting valve in one of the Eccles-Jordan triggers is 70V and so the biasing is set to be about 65V. In the case of the one dissimilar stage, the diode is connected in the reverse sense and its cathode, therefore, is connected to a similar biasing potential as the drive pulses.

As can be seen from the foregoing explanation, the amplitude of the reset and drive pulses is not very critical so long as they exceed the minimum voltage, governed by the extent to which the diodes are biased back, and the ability to turn the Eccles-Jordan trigger over. On the other hand, however, in the case of the positive reset pulse, it must not exceed a maximum governed by the potential at which grid current commences to flow; otherwise erratic action will occur. The width of the pulses is important. The resetting pulse must be wider than the 'feed along' pulse, from stage to stage, as its effect must be felt after all the stages that were in the incorrect state have all turned over. For a similar reason the drive pulse must be narrower than the 'feed along' pulse so that its effect ceases to be felt before the 'feed along' pulse stops.

Either a positive or a negative square wave is available at each stage dependent upon which anode is used as the output. To enable low impedance loads to be connected to the ring, cathode-followers are used for providing the outputs; these are of the high input impedance type, and so there is little loading effect on the ring itself.

Several 'rings of 24' with their associated gates have been constructed and have worked reliably for long periods.

Acknowledgment

In conclusion, the author wishes to thank the management of Rank Cintel for permission to publish this article and Mr. P. S. Smith, together with other members of the Instrument Development Department staff for their help.

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International Convention and Exhibition on Transistors

The Institution of Electrical Engineers have organized an International Convention on Transistors and this will be held in the Earls Court Exhibition Hall, London, from 21 to 27 May; associated with it, and in the same building there will be an international scientific exhibition where the leading laboratories and manufacturers of the world will be displaying their latest developments in the transistor field.

The Convention will be opened by The Rt. Hon. the Viscount Hailsham, Q.C., Lord President of the Council, following which the opening lectures will be delivered by Professor J. Bardeen, Dr. W. H. Brattain and Dr. W. B. Shockley. At the subsequent meetings there will be parallel sessions which will allow as wide a discussion as possible of the many parts of the subjects.

The Convention will be divided into 23 sessions, 12 of them General and 11 Specialist. Apart from the opening Session, which will be concerned with world trends in transistors and associated semiconductor devices, the Sessions will cover: Materials; Basic Theory; Characteristics and Their Measurements, and Transistors as a Circuit Element; Technology and Design and Manufacture; and Applications of Transistors.

The provisional programme is as follows:

Thursday, 21 May

10.30 a.m.	Session 1—Opening Session
12.30 for 1 p.m.	Convention Lunch
2.30 p.m.	Session 2—Materials
6.30 p.m.	Cocktail Party at the Institution

Friday, 22 May

10.30 a.m.	Session 3—Basic Theory Specialist Session 4—Materials
2.30 p.m.	Session 5—Characteristics and their Measurement and Transistors as a Circuit Element Specialist Session 6—Basic Theory
5.30 p.m.	Session 7—Characteristics and their Measurement and Transistors as a Circuit Element

Saturday, 23 May

10.30 a.m.	Specialist Session 8—Applications—Line Communication Systems Specialist Session 9—Applications—Use of Transistors as Computing Elements

Monday, 25 May

10.30 a.m.	Session 10—Characteristics and their Measurement and Transistors as a Circuit Element Specialist Session 11—Applications—Industrial Applications and Instrumentation.
12.30 for 1 p.m.	Convention Lunch
2.30 p.m.	Session 12—Technology and Design and Manufacture Specialist Session 13—Applications—Linear Amplification and Oscillators

Tuesday, 26 May

10.30 a.m.	Session 14—Technology and Design and and Manufacture Specialist Session 15—Industrial Applications—High-Speed Switching and Timing Circuits
2.30 p.m.	Session 16—Applications—Communications Specialist Session 17—Technology (Design)
5.30 p.m.	Session 18—Applications—Switching Specialist Session—19—Technology (Manufacture)

Wednesday, 27 May

10.30 a.m.	Session 20—Applications—Industrial Applications Specialist Session 21—Transistors as a Circuit Element—Circuit Theory
2.30 p.m.	Session 22—Applications—Industrial Applications Specialist Session 23—Characteristics and Measurements
7 for 7.30 p.m.	Convention Dinner at the Connaught Rooms

The exhibition will be essentially scientific in character and will cover all aspects of transistors and semiconductor devices and their applications. It will be broadly divided into three sections to link up with the general divisions of subject matter of the convention. The first section will cover Semiconductors, and will be open to firms engaged in the manufacture of transistors and other semi-conductor devices, materials and manufacturing techniques. The second section will be devoted to Semiconductor Applications and Associated Components. It will be open to firms engaged in the manufacture of equipment involving transistors and semiconductor devices and also to those supplying specialist components for use in transistorized equipment. The third section will be confined to Research Exhibits from industrial and commercial research laboratories as well as from universities.

LETTERS TO THE EDITOR

(We do not hold ourselves responsible for the opinions of our correspondents)

Stable Transistor Oscillator Circuits

DEAR SIR,—I was very interested to read Mr. McLeod's letter in the December issue describing the use of a series LC circuit in a transistor multivibrator to obtain a stable sinusoidal output. I agree that a circuit of this type can give excellent results.

There is a variation of his circuit which involves transferring the series LC components from the base to the emitter of the second transistor, which is now operated as a tuned class-A amplifier. Transistor X_1 remains a squaring amplifier. Since the emitter input impedance can be made about 50Ω, higher effective Q factors are obtainable. For instance, an LA7 Ferroxcube core can be wound for an inductance, of 1·36H and a Q of 40 at 1kc/s. At this frequency its series resonant impedance is therefore 200Ω. The additional 50Ω emitter input impedance only degrades the working Q to 32. Using the same inductance at 2·2kc/s, a working Q of 55 is possible.

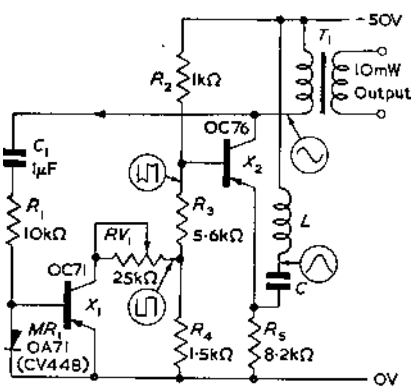


Fig. 1. Oscillator designed for telephone work

Fig. 1 shows a circuit which was developed for telephone work to operate from a 50V supply. Up to 10mW output is obtainable. Alternatively, the output transformer may be replaced by a 1kΩ resistor, the output being taken from the collector of X_2 via a coupling capacitor, or from a secondary winding on the tuning inductance L, provided the lowering of working Q can be tolerated. The output amplitude is set by RV_1.

Tests carried out on this circuit at 2·2kc/s with L = 0·2H and C = 0·27μF gave results which were almost identical to those of Mr. McLeod. The frequency drift of −0·01 per cent/°C was found to be entirely due to an increase of the tuning component values with temperature. A 10 per cent reduction in supply voltage increased the frequency by 0·5c/s or 0·02 per cent. Changing X_2 from the best to the worst available transistor changed the frequency by 5c/s or 0·2 per cent. I have no doubt that the use of a higher value of tuning inductance to give a greater working Q would have improved this figure.

The circuit oscillates readily when the supply voltage lies between 10 and 70V. Between 30 and 70V the output voltage is almost proportional to supply voltage. It is, of course, possible to redesign the circuit to operate at lower supply voltages.

Although untried, it appears possible with this circuit also to replace the series LC circuit with a series mode crystal, and obtain a sinusoidal output from the transformer in the collector of X_2.

The increased complexity of these circuits might be justified by the ability to use a smaller value of tuning inductance for the same frequency and stability, and the isolation of the output from the frequency determining circuit may also prove advantageous in some applications.

The writer wishes to thank the Director of the Central Electricity Research Laboratories for permission to publish this letter.

Yours faithfully,

J. N. PREWETT,

Central Electricity Research Laboratories, Leatherhead.

The Correspondent Replies:

DEAR SIR,—With reference to Mr. Prewett's circuit for a series LC oscillator, I would agree that this has many attractive and useful features, such as the choice of sine or square wave outputs from the transistor collectors.

With regard to the circuit Q, however, I feel that the total resistance between base and emitter of X_2 should be taken into account, as this largely determines the oscillatory current input to the class-A amplifier.

The frequency controlling portion of the oscillator circuit equation can be expressed as

$$\omega = \left[\frac{1}{LC} - \frac{R^2}{4L^2}\right]^{\frac{1}{2}} \quad \ldots\ldots\ldots\ldots (1)$$

where $2\pi/\omega$ is the oscillation time per cycle,

$$\text{i.e. } \omega = \omega_0 \left[1 - \frac{1}{4Q^2}\right]^{\frac{1}{2}} \ldots\ldots\ldots\ldots (2)$$

Expanding binomially and neglecting second order terms we have

$$\omega = \omega_0 \left[1 - \frac{1}{8Q^2}\right] \quad \ldots\ldots\ldots (3)$$

Equation (3) suggests that from the point of view of frequency stability, not much is to be gained by circuit Q's greater than about 20, and that reflected reactance is a more likely source of trouble. In this respect, transistors operated in an on-off condition would appear to be a better proposition than class-A operation.

For series mode crystal operation a fair amount of series resistance can be tolerated. The graph (Fig. A), taken from the manufacturer's literature[1] on

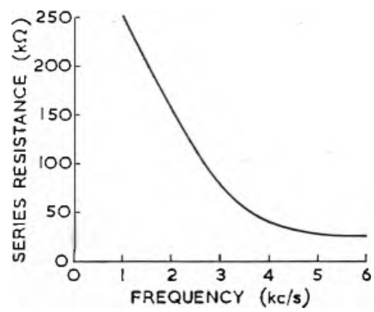


Fig. A. Manufacturer's graph of series resistance against frequency

XY flexural crystals, shows the maximum series resistance as a function of crystal frequency. It is important that the crystal should work into its correct load to prevent spurious mode oscillation.

With reference to my original note, the circuit described is the subject of British Patent Application No. 33495:58.

In Fig. 1, the "current in tuned circuit" waveform should be phase shifted 180°.

Yours faithfully,

D. D. MCLEOD

A.W.R.E. Aldermaston.

REFERENCE

1. FAIRWEATHER, D., RICHARDS, R. C. Quartz Crystals as Oscillators and Resonators (Marconi Wireless Telegraph Co. Ltd).

The Physical Realization of an Electronic Digital Computer

DEAR SIR,—In the course of studying Dr. A. D. Booth's article in the October, 1958, issue of ELECTRONIC ENGINEERING on input and output arrangements for computers, his comments on the low speed of solenoids for controlling parallel-input teleprinters were noted with interest.

The solenoid form of electromagnet has the disadvantage that its armature has a relatively large mass associated with a small pole area, so that the acceleration is low. Further, the armature stroke is relatively large, and combined with the low acceleration results in a long operating time.

Where it is desired to secure high speed operation of an electromagnet, it is appropriate to use a large pole-face area, a thin armature and a short stroke, giving high force, low inertia, high acceleration and low operating time.

Where the selector member has to be moved a long distance at high speed, it is not usually advantageous to arrange the armature pole-face to move the same distance at the same speed, because too much of the mechanical energy available is dissipated in moving the armature itself. An overall speed gain is usually secured if some form of leverage is used to couple a short-stroke armature to a long-stroke load, despite the added inertia of the lever system.

It is not correct to suggest that electromagnets of high power are necessarily slow. In a teleprinter the electrical signals received influence the selector mechanism by means of an electromagnet whose transit time is of the order of from 1 to 5msec; the work done is, however, small. If the work to be done is 100 times as great, it is clear that 100 such electromagnets operating simultaneously on the common load will do this work in 5msec, and these 100 electromagnets can be combined to form a single large one. It follows that for a large amount of work to be done in a short time not only must a large electromagnet be used, but it is essential to use a large pole-face area and a short armature movement.

The moving-iron type of reciprocating motor is perhaps not the best to use for this kind of purpose. As magnetic saturation sets a limit to the force attainable for a given structure, full advantage cannot be taken of duty cycles having short 'on' and long 'off' periods. The moving-coil type of motor is to be preferred for this reason. So much development has been carried out on loudspeaker driving systems that it should be possible now to produce a teleprinter selector motor of high speed at an economic price and of tolerable size.

Yours faithfully,

F. P. Mason,
Creed & Company Limited,
Croydon.

The Author Replies:

Dear Sir,—By and large I agree with Mr. Mason's remarks. His point regarding the mating of a thin, large area, small travel armature with some form of lever mechanism to obtain large travel is, of course, well known to every designer of mechanisms. It is probably more familiar to electronic engineers as the problem of impedance matching, since there exists a unique lever ratio which will produce the most rapid effect from a given armature. The determination of the optimum armature and leverage is, however, a matter of some difficulty and we have usually evaluated it by experimental methods which are themselves complicated by the fact that below a certain thickness the armature suffers magnetic saturation at its centre as it approaches the pole-faces. An example of the way in which leverage can be of help occurs in the rotary solenoid which we use to drive our tape reader, the lever in this case being replaced by an inclined plane. The system has some of the characteristics which Mr. Mason mentions as desirable, specifically a large pole-face area, a very small travel and a rotary mode of operation. The ball bearing system upon which the solenoid operates is useful in this case because it eliminates friction.

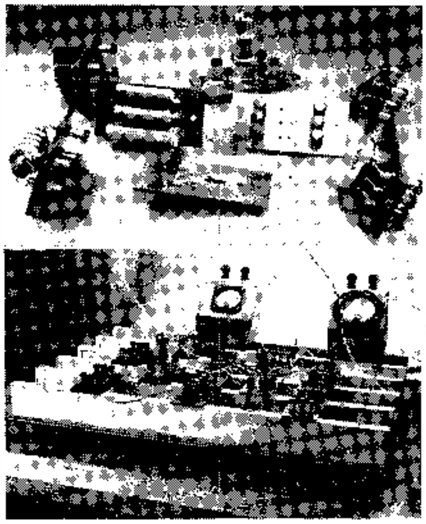
With regard to his suggestion that 100 small electromagnets operating simultaneously can produce large amounts of power, this is indeed quite true and, in principle, one should be able to design a large electromagnet to do the same thing. The proof of the pudding is, however, in the eating and I must confess that I have not personally seen an electromagnet of great size and really high speed.

Yours faithfully,

A. D. Booth,
Department of Numerical Automation,
Birkbeck College,
London, W.C.1.

A New Teaching Aid

Dear Sir,—With reference to your description in the February, 1959, issue of Electronic Engineering, of the American Erec-tronic Educational System, it is thought that there might be interest in the solution of the same problem which has been developed at this Establishment.



Sub-assemblies and a complete circuit

As can be seen in the photographs, components, valveholders, etc., are permanently wired to spring-loaded terminals mounted on panels. The panels are stored in drawers adapted for the purpose, and component panels are colour coded for ready identification.

To construct a circuit, panels containing the required parts are laid out on a baseboard. The baseboard is wired so that the placing of a valveholder panel in position connects the valve heater to one or other of two pairs of heater leads. This enables separate heater supplies to be used in circuits where cathodes are operated at significantly different potentials with respect to earth. Circuits may quickly be constructed using any suitable wire to link the spring-loaded terminals. Changes in circuits are also made easily.

This system has been in use for about ten years, although the use of spring-loaded terminals is a more recent innovation.

Yours faithfully,

R. H. James,
Ministry of Transport
and Civil Aviation,
Bletchley.

A Microwave Reflectometer Display System

Dear Sir,—Should Messrs. J. C. Dix and M. Sherry wish to increase the stability of output power in the X-band reflectometer they describe in Electronic Engineering, January, 1959, this can readily be achieved by applying the feedback to the anode, rather than the grid, of the b.w.o.

When feedback is applied, it is essential that the collector be fed from a low impedance source. The appreciable changes in beam current that occur will distort the output frequency characteristic. An electronic modulator, which fulfils this requirement, and provides for a linear frequency sweep, is the subject of an article to be published.

Messrs. Dix and Sherry rightly stress the importance of high directivity. It should be pointed out that when measuring the v.s.w.r. of a component over the whole band, even with the excellent couplers he describes, the inaccuracy increases very rapidly at the edges of the band. Moreover the ripple which appears makes reading extremely difficult. For instance, if the square law detector is followed by a linear amplifier, given a directivity of 30dB and a v.s.w.r. of 1·5 the trace will swing between approximately ± 30 per cent of its mean value.

It can be shown that the ripple frequency, f_r, is given by

$$f_r = \frac{6 \cdot 66 l}{\sqrt{[1 - (f_c/f)^2]}} \quad \ldots\ldots\ldots\ldots\ldots \text{(1)}$$

cycles per kMc/s of scan

where l = length of waveguide between coupler and test components in metres.

f_c = cut-off frequency of waveguide.

f = output frequency.

Thus by using sufficient line length l, the ripple frequency can be raised to a high value such that the incorporation of a low-pass filter before the display removes the ripple, without losing important information from the trace.

For example, using a line-length of 2 metres of WG 16 (f_c = 6·55kMc/s), and a 10msec scan (i.e. 400kMc/s/s) the frequency is 6·5kc/s at 11·5kMc/s and 11kc/s at 7·5kMc/s, and this can be removed by a low-pass filter cutting off at 6·5kc/s. The introduction of this filter will of course remove any detail which occurs in a time that is less than 1·8 per cent of the total scan. The resolution can be further improved by increasing the line-length, but apart from being cumbersome, errors will eventually be introduced by the line-loss.

There is also an error introduced if the output from the square-law detector is averaged without square-rooting beforehand. This error is approximately $(D/\rho)^2$ where D is the directivity and ρ the reflection coefficient. With 30dB directivity and a v.s.w.r. of 1·5 this is about 2·5 per cent.

The derivation of equation (1) is as follows:

Peaks occur when $2l/\lambda_g$ is an integer (n)

$$(n/2l)^2 = (1/\lambda_g)^2 = (1/\lambda)^2 - (1/\lambda_c)^2$$

$$= (f/c)^2 - (f_c/c)^2$$

$$\therefore\ n = 2lf/c\ \sqrt{[1 - (f_c/f)^2]}$$

$$\text{and } dn/df = 2l/c\ \frac{1}{\sqrt{[1 - (f_c/f)^2]}}$$

cycles per cycle of scan.

where λ_g = guide wavelength

λ_c = cut-off wavelength

f_c = cut-off frequency

$c = 3 \times 10^{10}$

Substituting for c and expressing l in metres and f_r in cycles per kMc/s gives equation (1).

Yours faithfully,

G. M. Clarke,
R. D. Rookes,
Ferranti Ltd.

The Authors Reply:

Dear Sir,—We thank Messrs. Clarke and Rookes for bringing to our attention the various points regarding reflectometer design detailed in their letter.

In reply we would point out that:—

(1) We used grid rather than anode feedback to avoid having to provide a low impedance driving stage. Although the power requirements at X-band are not high they may prove more troublesome when using the lower frequency b.w.o.'s where the anode takes considerably more current.

(2) We are aware of the necessity of feeding the line (collector) from a low impedance source. This was one of the reasons for deriving the scan voltage from the mains supply. A small amount of f.m. does take place as a result of feedback but this does not appreciably affect the performance of the instrument when used on the wide band displays.

(3) The method of overcoming the ripple due to imperfect directivity is ingenious and would be useful in some cases. From the point of view of a general purpose instrument it has, however, the following disadvantages:—

(a) The filter necessary is dependent on the width of scan.

(b) Loss of detail results and this could quite easily obscure the detection of an unwanted sharp resonance.

Your faithfully,

J. C. Dix and M. Sherry.
Research Laboratories,
The General Electric Company Limited,
Wembley.

Multivibrator Circuit Using PNP and NPN Junction Transistors

Dear Sir,—In some of the past issues of Electronic Engineering[1,2,3,4,5] the problem of obtaining sharp waveforms at the collectors of a transistor multivibrator circuit was fully dealt with, including some different solutions concerning the reduction of the integrating effect of the cross-coupling capacitors.

In further consideration of this problem the circuit is greatly improved by the use of combined pnp and npn junction transistors arranged as shown in Fig. 1. The

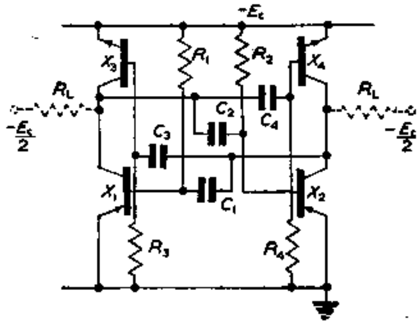


Fig. 1. Multivibrator circuit using pnp and npn junction transistors

main advantages of this arrangement are low output impedance and small power consumption. It consists of the combination of two pairs of complementary transistors (X_1X_2, pnp type, X_3X_4, npn type). Simultaneously either both X_1 and X_4 or X_2 and X_3 are conducting. The bases and collectors are cross-coupled through four timing capacitors. The base resistors R_1 and R_2 are connected to the negative and R_3 and R_4 to the positive pole of the battery. The collector resistors R_L may exist or not. The configuration leads to a fast rise and decay time as the discharge of the coupling capacitors now takes place through the low impedance paths of 'on'-going transistors when regenerative action occurs but not through

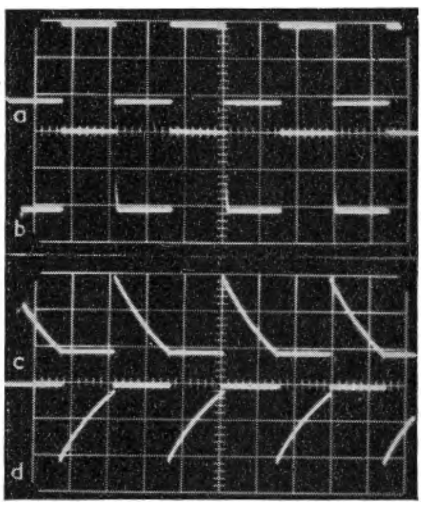
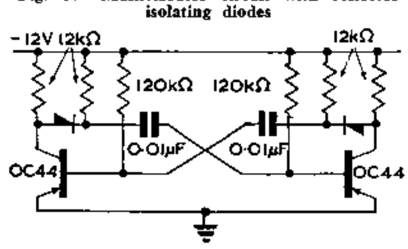


Fig. 2. Waveforms of the circuit (Fig. 1)

(a) Collectors X_1X_2 unloaded; (b) Collectors X_1X_2 capacitively loaded with C = 30pF; (c) Base X_2; (d) Base X_3; X_1X_2 — OC44; X_3X_4 — 2 N168A; $R_{1,2,3,4}$ = 120kΩ; $C_{1,2,3,4}$ = 10pF; R_2 = infinite; E_c = 12V; I_{total} = 0·8mA; f = 2kc/s

Fig. 3. Multivibrator circuit with collector isolating diodes



collector resistors R_L as was the case of the basic multivibrator circuit.

Very low static and dynamic output impedance of the collector terminals enables a small number of low input impedance circuits to be driven with no appreciable change in the waveform characteristics. This feature of the circuit is illustrated in waveforms (a) and (b) of Fig. 2. In comparison the multivibrator circuit with collector isolating diodes[3],

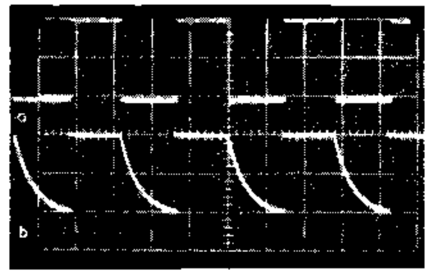


Fig. 4. Collector waveforms of the circuit shown in Fig. 3

(a) Collectors unloaded; (b) Collectors capacitively loaded with C = 30pF; frequency = 2kc/s

shown in Fig. 3, gives a rectangular output under no load conditions (Fig. 4(a) but with capacitive loading the effect is shown in Fig. 4(b).

If it is desired to take the output from one pair of collectors in circuit Fig. 1, any one transistor may be replaced with a resistor, thus reducing the number of transistors used from four to three, without any change in the output waveform characteristics.

The circuit in Fig. 1 can be made monostable by replacing one a.c. coupling between collectors and bases with d.c. coupling.

Yours faithfully,

D. T. Jovanovic,
Department d'Electronique,
Centre d'Etudes Nucléaires de Saclay,
France.

REFERENCES

1. Jacket, A. E. Multivibrator Circuits using Junction Transistors. *Electronic Engng.* 28, 184 (1956).
2. Armstrong, H. L. A Transistor Multivibrator Circuit. *Electronic Engng.* 29, 251 (1957).
3. Rozner, F. Transistor Multivibrator Circuit. *Electronic Engng.* 29, 455 (1957).
4. Jacket, A. E. A Method for Sharpening the Output Waveform of Junction Transistor Multivibrator Circuits. *Electronic Engng.* 30, 371 (1958).
5. Leece, N. L. Multivibrator Circuits, *Electronic Engng.* 30, 615 (1958).

Correction:

In the 'Letter to the Editor' 'Response of a Capacitance-Resistance Divider' by R. H. Evans on page 232 of the April issue, the following error occurred:

In the line

$$V_s(t) = k\,(1 - \alpha e^{-t\nu})$$

the last part should read $e^{-t/\nu}$. In equation (2), the form $e^{-(t\times)/\nu}$ should be $e^{-(t-x)/\nu}$.

The second line of 'The Author Replies' should read divider response, not divided response.

BOOK REVIEWS

Electronic Apparatus for Biological Research

By P. E. K. Donaldson and contributors. 718 pp. 80 figs. Royal 8vo. Academic Press Inc., New York. Butterworths Scientific Publications, London. 1958. 120s.

IN his preface, Mr. Donaldson states that he has emphasized techniques, and largely left the spotting of applications to the reader. Since there are comparatively few electronic techniques, it is more efficient to explain these systematically rather than survey the whole extent of biological research and describe the appropriate apparatus under each heading. In adopting this plan, the author has divided the book into four parts: (I) Theory (282 pp.); (II) Practice (50 pp.); (III) Transducers, Electrodes & Indicators (252 pp.); and (IV) Complete Apparatus (126 pp.).

The first section provides a concise introduction to electronics, requiring only an understanding of algebra and simple differential equations. Much of little interest to biological workers has been omitted by concentrating mainly on frequencies from the audio range down to d.c. Starting with ideal generators and Ohm's Law, networks, a.c. theory and filters are clearly explained. After developing the elementary theories of valve and rectifier operation, numbers of basic circuits are studied with a healthy emphasis on design considerations and working limits. In succeeding chapters 'functional units' are discussed, including the stages required for balanced d.c. amplifiers, tuned amplifiers and waveform generators. The importance of feedback is stressed throughout. The text is supplemented by clearly drawn graphs, many of which are usable for design purposes as they are plotted in non-dimensional or generalized units. This application may have been the reason for grouping them together at the end of Part I, but this arrangement is very inconvenient when the book is being read systematically, especially as graph numbers are quoted without reference to the page. On pages 81 and 82 some graph numbers are given incorrectly.

In the practical section, common electronic components are described and their uses and limitations discussed. Battery economics are treated fully since much electrophysiological apparatus is not easily mains operated. The semi-empirical design and construction of chokes and transformers using Ferroxcube is also covered in some detail. This part closes with suggestions for chassis design, layout and wiring. The plates illustrating components are well produced and informative.

Part III, except for the author's chapter on Visual Indicators, is written by specialist contributors. Some unevenness of treatment is inevitable, but this is offset by the advantages of authoritative descriptions. There are chapters covering the control and measurement of temperature, humidity and light; the assay of radioactivity; electromechanical transducers; relays; and various biological electrodes. There is insufficient space to consider these subjects in detail, but Dr. K. E. Machin's treatment of transducers using electrical analogues is particularly successful, since he is able to refer to the graphs in Part I. The correct matching of mechanical and electrical impedances is often overlooked in planning experiments, and it is hoped that this chapter will prevent such mistakes. All the contributors to this section provide useful reference lists; that given by Dr. D. W. Kennard on microelectrode technique being especially comprehensive. Some photographs in this part of the book are reproduced rather badly.

In the fourth part, the author deals with complete apparatus, and gives a number of practical circuits with component values. Subjects covered include stabilized power supplies, timers, counters, reduction of interference, fault diagnosis and design procedure. A chapter on stimulators includes a useful discussion on stimulus artifact and its relation to the preparation-input equivalent circuit. As was expected, more space is allotted to biological amplifiers than to other circuits, but their importance could justify more than the 22 pages given. The five-page chapter on recording methods is also disappointingly short. Tape systems with responses from d.c. towards 10kc/s are becoming essential for experiments requiring resolution of random transients during prolonged recording. In this application photographic methods become uneconomical. This development is treated very briefly despite the author's work on the problem. The final chapter is referred to as a 'stop press' item. It concerns transistors, which are but slowly penetrating into this field, retarded by their past and present shortcomings. In the course of 33 pages one is led from simple description to complete circuits, most of this space being devoted to equivalent circuits and the chaotic multitude of transistor parameters.

The author has achieved his aim in producing a readable text book and work of reference in one volume. The biologist new to electronics will certainly find this the ideal combination, but others familiar with basic electronic theory may object to paying a high price for surplus information, however well written. A division into two volumes with extra space devoted to usable circuits might have been the better solution. However, as Professor Sir Bryan Matthews says in his foreword, ". . . this book will be valuable both to the beginner entering this field, and also to the more experienced, for its comprehensiveness makes it a reference work of electronics in biology".

A. S. VELATE

Transistor Technology, Volume III

Edited by F. J. Biondi. 416 pp. 60 figs. Demy 8vo. D. Van Nostrand Co. Ltd. 1958. Price 94s.

THIS book is the third of a series of three volumes which deal with semiconductor technology and devices. The volume is in four parts covering the preparation of p-n junctions, fabrication technology, measurement of the parameters of materials and devices and transistor reliability.

Part I—Preparation of Junction Structures—has three chapters which deal with the two fundamental processes—freezing and diffusion—which are used to control the impurity distribution in silicon and germanium. The use of crystal growing, alloying and diffusion techniques to form p-n junctions is discussed in detail.

Part II—Fabrication Technology—occupying about half the volume, is concerned with the chemical processing of materials and devices, techniques for making ohmic contacts, alloying and diffusion techniques for fabricating diodes and transistors in both germanium and silicon, diffused base germanium transistors and diffused emitter and base silicon transistors.

Part III—Measurement and Characterization—deals with methods for the determination of volume life times and surface recombination velocity in materials and with the measurement of the electrical parameters of transistors.

Part IV—Transistor Reliability—discusses the effects of temperature, shock and ageing effects on the life of transistors and gives an analysis of failures of different types of transistors in field equipment.

Appendix I gives the list and definition of semiconductor terms which was published in the Proceedings of the I.R.E. in 1954, while Appendix II gives the I.R.E. Standard Letter Symbols published in 1956.

Much of the material included in this volume is new. The detailed information given on device fabrication techniques, particularly that on diffusion, will make the book of inestimable value for many years to every engineer or physicist concerned with semiconductor devices.

E. G. JAMES

The Theory of the Potential

By W. D. MacMillan, 465 pp. 112 figs. Demy 8vo. Dover Publications Inc., New York. 1959. Price $2.25.

BOTH physicists and mathematicians will find this book a useful presentation of the theory of the potential. Only reasonable mathematical proficiency is assumed and all mathematical theorems are fully developed as they become necessary. The mathematical material is treated in extensive detail, covering gravitational mechanics as well as electrostatics and magnetostatics. Problems are included at the end of the chapters.

Free Radicals as Studied by Electron Spin Resonance

By D. J. E. Ingram. 274 pp. 66 figs. Demy 8vo. Butterworths Scientific Publications. 1958. Price 50s.

A FREE radical is defined by the author as 'a molecule, or part of a molecule, in which the normal chemical binding has been modified so that an unpaired electron is left associated with the system'. The presence of free radicals is shown by the occurrence of a high chemical reactivity and by a magnetic moment due to the uncompensated spin motion of the odd electron. The first of these properties is obviously of great interest to chemists, while the second provides the possibility of detecting relatively small numbers of free radicals. The most powerful method for this is to place a sample of the material in a microwave resonant cavity and to apply a static magnetic field. The absorption of energy in the cavity shows a marked increase when the magnetic field has an appropriate value. The shape of the resonance curve, displaying absorption against magnetic field can provide very detailed information about the internal structure of the sample. This book provides a comprehensive account of the techniques used in this work and of the ways in which the results may be interpreted.

The description of the microwave equipment is directed towards the reader who has no previous knowledge of the subject and provides a complete description of all the components and electronic circuits needed. This should prove extremely useful to workers entering this field. A chapter on the theory of the shapes of the resonance curves assumes a good knowledge of quantum theory. The remainder of the book discusses in detail results which have been obtained by the resonance method for stable-free radicals, and for free radicals produced by irradiation, polymerization and pyrolysis. The use of the methods in biological and medical work is summarized in the final chapter.

It is obvious from the contents that this book is of most use for the physicist, biologist or medical worker who is anticipating the use of the electron resonance method or who is concerned with the results which have been already obtained. Such readers will find that it serves their purpose admirably. It can also be recommended for more casual reading to the inquisitive microwave engineer, who is interested in knowing to what uses his waveguides and resonators can be put.

J. Brown

Advances in Electronics and Electron Physics, Volume X

Edited by L. Marton. 320 pp. 80 figs. Demy 8vo. Academic Press, Inc., New York, London. 1958. $10.00.

TEN years have now passed since this authoritative series began to appear. The cumulative subject index which forms a useful appendix to the present volume indicates that few aspects of electronics have escaped discussion; but there seems no reason to doubt that at current rates of progress the series will find plenty of new material, and of appreciative readers, for its next decade.

Both electron and microwave 'optics' are represented, the first in a basic theoretical analysis, by W. G. Dow, of magnetic focusing as applied to non-uniform beams for which Brillouin flow cannot be assumed Here the notions of ordinary optics are of little assistance. By contrast, J. Brown's treatment of microwave optics is concerned to indicate the wealth of optical theory which can be transferred ready-made to the microwave field. The advantages are mutual, for many optical-type experiments—e.g. on the system of the eye—are more conveniently carried out on scaled-up microwave models.

Because of their purity, modern semiconductor crystals offer unusually attractive material for the study of the solid state. E. Billig and P. J. Holmes present a valuable summary of work on the mechanical and other effects of residual imperfections in such materials, destined doubtless to have implications for electronics as well as for fundamental physics.

The Logical Organization of Computers is not specifically an electronic problem, but electronic engineers will find both stimulus and food for thought in W. J. Lawless's relatively brief review. E. G. Rowe's discussion of valve reliability will certainly be of interest to computer designers and many others.

The volume ends with a welcome survey of post-war developments in c.r. oscillography, by one well-qualified to evaluate them, Dr. Jack E. Day. Both circuit and tube design are covered, with some practical data on special components, and the section on new techniques developed for u.h.f. oscillography is of special interest.

Perhaps little is lost by the fact that the survey covers mainly American publications, though (if it is not indelicate to say so) the present reviewer was surprised to find an American reference of 1949 as the earliest given on p. 253 for the war-time suggestion of a distributed-plate deflexion system (published here in Proc. Phys. Soc., L. C., LXI, p. 242, 1948).

A 7-page subject index and an author index much enhances the value of this stimulating work.

D. M. MacKay

The Elements of Non-Euclidean Geometry

By D. M. Y. Sommerville. 267 pp. 129 figs. Demy 8vo. Dover Publications Inc., New York. 1959. Price $1.50.

ALTHOUGH this is essentially an elementary book, the author deals with such important and difficult subjects as the relation between parataxy and parallelism, the pseudosphere, the absolute measure, and geodesic representation as well as others, in a manner which can be easily understood. For the benefit of the student, there are 126 problems at chapter endings.

Guide to the Literature of Mathematics and Physics including Related Works on Engineering Science

By N. G. Parke III. 436 pp. Demy 8vo. 2nd Edition. Dover Publications Inc. 1959. Price $2.49.

THIS book includes an up-to-date listing of agencies and individuals who are engaged in Russian translation programmes. The guide lists more than 5 500 key works under 120 subject headings such as Projective Geometry, Geometric Optics, and Cosmic Rays. Whenever a foreign language title appears, any translations are also listed. If there is no English translation of a Russian work, for example, the guide will mention if there is one in German.

Kempe's Engineers' Year Book 1958

3 000 pp. Crown 8vo. (Two volumes in case). 64th Edition. Morgan Bros. (Publishers) Ltd. Price 82s. 6d.

EACH year, every chapter of this year book receives the close attention of an eminent authority, thus ensuring that the contents are kept up to date. In this latest edition four chapters have been rewritten and valuable additions have been made to no less than 24 other chapters while all have been revised where necessary.

The seventy-nine chapters cover practically every branch of engineering. A comprehensive index, with more than 17 000 entries, makes reference quick and easy, while most chapters include a bibliography giving a list of standard works on the subject.

ELECTRONIC EQUIPMENT

A description, compiled from information supplied by the manufacturers, of new components, accessories and test instruments.

(Voir page 309 pour la traduction français: Deutsche Übersetzung Seite 314)

PEAK VOLTMETER

(Illustrated below)

Wayne Kerr Laboratories Ltd, Chessington, Surrey

The Wayne Kerr Laboratories in collaboration with the United Kingdom Atomic Energy Authorities have developed a 50kV peak voltmeter for the measurement of single or repetitive pulses of 1μsec duration or greater. The incoming pulse is applied to a voltage dividing capacitor network with a ratio of 50:1 and the voltage across the larger capacitor is measured by means of a 0 to 1kV electrostatic meter. A level restoring diode shunts the meter to allow repetitive pulses to be measured. The indicated pulse slowly discharges through an insulation resistance of greater than

$10^{15}\Omega$, the measuring circuit being hermetically sealed to maintain this figure. Discharge can be obtained by pressing a button of the front panel which operates a relay within the sealed compartment.

Two ranges are readily available, 10kV to 50kV and 5kV to 25kV. The accuracy of measurement is better than ± 5 per cent using calibration curves supplied showing initial calibration to irradiated sphere gap to B.S. 358.

EE 9751 *for further details*

CABLE TESTER

(Illustrated above right)

Tyser & Co. Ltd, Merrow, Guildford, Surrey

A new instrument, able to give high-speed detection of faults in cables, has been developed and put into production by Tyser and Co. Ltd, successor to P.A.M. Ltd in the Southern Areas Electric Corporation.

It is known as the Perram cable tester, indicator and recorder, and its function is to detect faults in the cover insulation of cables during manufacture. It offers a non-destructive method of assessing

the quality of the insulation of cables, which can be graded accordingly.

It comprises a non-lethal high voltage d.c. generator, with a continuously adjustable output between 4 to 20 kV, d.c., a high speed electromagnetic counter, an alarm indicator employing a lamp and bell, and an automatic overload protection circuit.

The high voltage output is applied to a small ring electrode used for examining the cable insulation; when a fault or break in the insulation is detected the resulting voltage discharge is registered by the counter.

When a fault is located, the alarm system operates for a minimum period of 4sec, giving visible and audible warning. A close succession of faults will always be recorded on the counter, even if the alarm system will not follow the speed of repetition.

For a sustained fault, an automatic overload protection circuit comes into operation switching off the high voltage generator, and keeping the alarm indicators working until the operator resets the instrument to normal.

The circuit permits a very high-speed of operation, and a fault recognition time as short as 1msec will be recorded by the instrument, this allows a cable speed of 3 000ft/min, even when using a very short ring electrode.

EE 9752 *for further details*

TRANSIENT STORAGE OSCILLOSCOPE

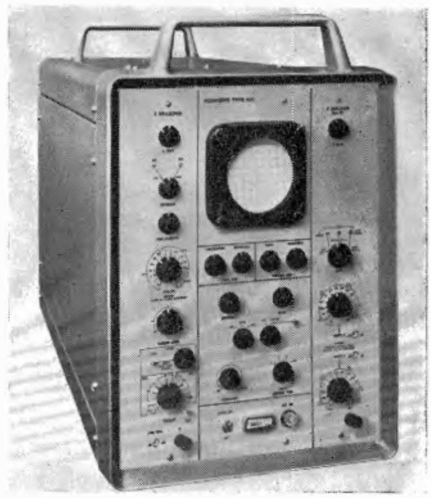
(Illustrated above right)

A. E. Cawkell, Electronics Ltd, 6-8 Victory Arcade, Southall, Middlesex

Known as the 'Remscope' this unit greatly extends the range of measurements and observations possible without recourse to photographic recording.

Previous storage oscilloscopes made by Cawkell's have used the American 'Memotron' cathode-ray tube and their performance has been limited by the relatively slow response of that tube. Now, with the advent of the English Electric type E702 storage and display tube, this limitation no longer applies and the 'Remscope' has been designed to make the fullest use of the tube.

The E702 tube, contains a storage system which allows a transient to be stored for up to a week and displayed for a limited period during that week. In the 'Remscope' this display time (which may be taken piece-meal) exceeds 15 minutes



and may, on all but the fastest signals, be extended to nearly 2 hours. Some slight reduction of brightness occurs on the longest display times but an adequate display is still obtained. In order that the maximum use may be made of the stored image, the display may be left off until required (storage only occurring) or may be switched on automatically when a signal is received.

The persistence of the display can be varied from a few seconds up to 15min. This facility is particularly useful when comparing one event from a series with its predecessor. By adjusting the persistence the first signal can be made to fade as the third appears and so on. Varying the setting will allow other sequencies to be produced.

Erasure is rapid and may be obtained by depressing an internal press-button or making an external circuit. An external erase timer, allowing the signal to be removed after a pre-set interval, will be available.

Initially a single-channel wide-band Y amplifier with differential d.c. inputs will

THE TRANSISTOR EXHIBITION

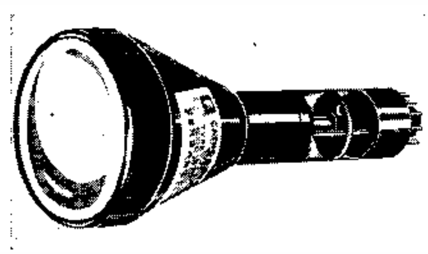
'Electronic Engineering' will occupy Stand No. 73 at the Transistor Exhibition at Earls Court from 21 to 27 May. Full details of the exhibits are not yet available and cannot, therefore, be included in this issue. A review of the exhibition will, however, be given in a subsequent issue.

be fitted. To enable it to be replaced by different units in the future this amplifier is of a plug-in type.

As the oscilloscope is intended primarily for the study of transient signals great care has been taken to ensure reliable and easy triggering. A.C. or d.c. signals may be used to start the time-base and starting is automatic once the input potential exceeds a pre-set level.

Apart from those enumerated above all the standard facilities expected of a high grade oscilloscope are incorporated. The time-base covers a wide range of velocities from 3cm/μsec to 0·1cm/sec and a five times expansion is possible by means of the X amplifier. The Y amplifier has a rise-time of 0·15μsec but is also provided with a high-gain narrow-band operating condition. Both amplifiers and the triggering circuits are direct-coupled throughout. Crystal-controlled time markers are provided for calibrating the time-base and they are clamped accurately to provide a voltage standard. The voltage measuring accuracy is ± 1 per cent. Stabilized h.t. and heater supplies ensure the minimum of drift.

EE 9753 *for further details*



HIGH SPEED FLASH TUBES

(Illustrated above)

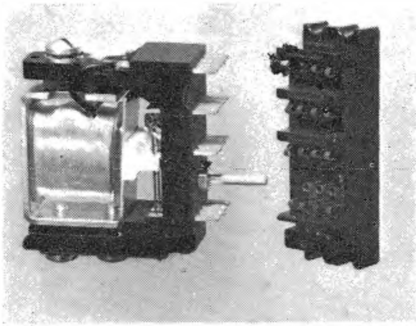
Ferranti Ltd, Hollinwood, Manchester, Lancashire

A new range of high-speed cathode-ray flash tubes which produce a flash of the same order of intensity as a gas-filled electronic flash tube, such as is used in photography, has been introduced by the Electronics Department of Ferranti Ltd.

These tubes known as the CL60 and CL70 series light up all over simultaneously in contrast to the normal behaviour of the cathode-ray tube, and have the additional advantage that flashes can be repeated much more rapidly. Flashes of up to one million per second are obtainable. It is also possible to produce long flashes, and flashes which contain an internal fine structure consisting of a multiplicity of smaller flashes.

There are three different types of tube currently available. The CL60 series are triodes, 3½in in diameter, the CL70 series are miniaturized, 1in in diameter, and the CL1 series are diodes, 6in in diameter. Tubes can be made with phosphors of all colours including ultra-violet and infra-red. It is also possible to produce 'white' light free from the spectral line structure associated with gas-discharge lamps and extending over specified spectral ranges.

EE 9754 *for further details*



WIRE CONTACT RELAYS

(Illustrated above)

Woden Transformer Co. Ltd, Moxley Road, Bilston, Staffordshire

The Woden-T.S.P. wire contact relay has been introduced to meet the needs for a high speed multi-pole relay in computer and general control engineering fields

In basic form it is a plug-in relay designed for use in individual or multi-holders which can be soldered directly or used with a quick release plug.

The basic unit is a 4-pole changeover relay with twin wire contacts which are readily interchangeable. The unit and all fixed contacts are in a single piece moulding of high dielectric strength material. All fixed contacts are of silver alloy.

The design of the relay is such that high speed operation is obtainable with operating and release times down to 3msec. This type of relay may be used for general applications. but finds particular use in computer and video switching circuits, the low capacitance between contacts being particularly helpful.

The unit has an expected contact life of 3 000 000 operations and the range of contact life can be considerably extended by the use of quenching circuits or use at lower ratings.

The standard 4-pole unit is dimensioned approximately as follows:—1¾in × ¾in × 1½in; 6 and 12-pole units are also available.

The relays are suitable for use in maximum ambient temperatures of 50°C.

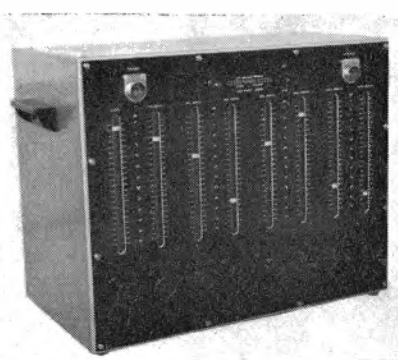
EE 9755 *for further details*

MULTI-OCTAVE A.F. FILTER

(Illustrated below)

Peekel, Laboratorium Voor Electronics N.V., Alblasstraat 1, Rotterdam 8, Holland

This filter enables an arbitrary frequency characteristic to be obtained in

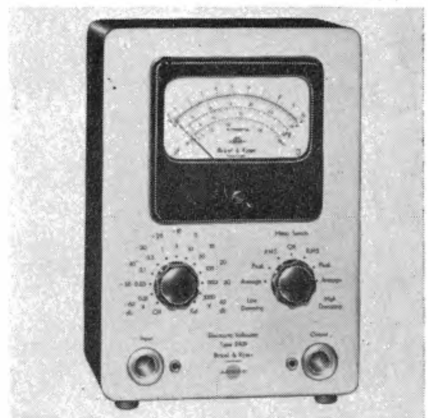


the audio frequency range. The frequency range is 30c/s to 16kc/s and it is divided into narrow bands by means of eight filters, the action of which can be regulated individually by means of attenuators calibrated in decibels. The attenuators are designed in such a way that the position of each one may be read on a vertical scale and they are arranged so as to form a system of co-ordinates with the frequency as abscissa and the attenuation as ordinates; the position of the eight attenuators thus showing at a glance the frequency characteristic obtained.

The average gradient of each filter is 30dB/octave and each filter can be adjusted between 0 and -55dB in steps of $-2{\cdot}5$dB. When all the filters are adjusted to the same level a linear frequency characteristic is obtained between 30c/s and 16kc/s; when the attenuators are at the 0dB position the unit acts as an amplifier with unity gain.

The input and output impedance of the multi-octave filter is 200Ω and the normal voltage which may be applied to the input is 1V.

EE 9756 *for further details*



VALVE-VOLTMETER

(Illustrated above)

Briel and Kjær, Nærum, Denmark

The new electronic voltmeter type 2409, developed by Bruel and Kjaer, can be used for all kinds of voltage measurements in the frequency range 2c/s to 200kc/s and within the voltage range from about 1mV to 1kV. It has two voltage scales with full-scale deflexion for 10V and 31·5V respectively. For measurements of amplification or attenuation ratios, a decibel scale is provided. When the decibel scale is used in connexion with the calibrated attenuator, any voltage measured may be read directly in decibels relative to 1V.

Furthermore, a dBm scale is supplied for measurements on a.f.-transmission lines, telephone lines, etc. The reference value is 0·775V corresponding to 1mW in 600Ω.

A feature of the instrument is its capability of measuring the peak, the arithmetic average and the true r.m.s. value of the input signal.

In order to enable easy, and accurate meter readings for both high and low frequency signals, two different meter damping characteristics can be selected. One of these is in accordance with the standard V.U.-measurements. The second damping characteristic is used to obtain a steady deflexion of the meter pointer when measuring low frequency signals.

By means of the output jack the instrument can be used as a calibrated amplifier, featuring a low output impedance and a maximum amplification of 60dB variable in steps of 10dB.

EE 9757 *for further details*

GENERAL PURPOSE OSCILLOSCOPE

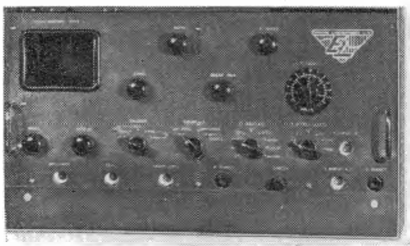
(Illustrated below)

Erskine Laboratories Ltd, Scalby, Scarborough, Yorkshire

The Erskine type 5 oscilloscope is a general purpose instrument for the observation of waveforms and their time and amplitude measurements.

It is light and compact yet adequately ventilated and component accessibility is extremely good.

The cathode-ray tube (four anode type) is positioned horizontally and is viewed through a 45° mirror. This arrangement has advantages. It allows a viewing hood to be confined within the overall cover dimensions. It allows a better functional lay-out of internal circuits, the front



panel does not become cramped with controls and the depth of the instrument is no longer dictated by the c.r.t.

All the necessary input and output connexions are arranged along the bottom edge of the front panel, and positioned adjacent to their associated controls. A full length compartment for connector stowage is provided immediately below the panel. This also serves as a convection air inlet for ventilation, the necessary outlet being over the top of the panel, which is sloped, the top edge finishing behind and below the top edge of the dust cover.

The oscilloscope has a stabilized h.t. supply which allows a direct reading against calibrated graticules regardless of supply fluctuations. A delay circuit allows any part of a repetitive waveform to be viewed up to 20msec from the synchronizing edge. The time-base has nine preset speeds between 3μsec and 30msec. The X-plate sensitivity is 30V/cm and the Y-plate sensitivity is 45V/cm; the Y amplifier has a bandwidth of from d.c. to 4Mc/s and a gain of 100.

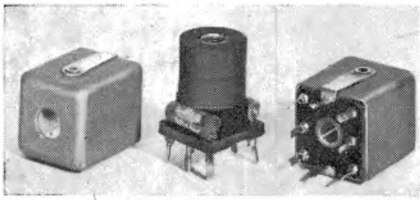
EE 9758 *for further details*

MINIATURE POT CORE ASSEMBLIES

(Illustrated above right)

Neosid Ltd, Stonehills House, Welwyn Garden City, Hertfordshire

This assembly which is enclosed by a ½in square can has been developed for



transistor circuits, in particular where printed circuits are employed. The tag arrangement on the base complies with the B.S.I. recommendations for printed circuits. The can earthing clips perform the dual function of earthing and retention of the whole assembly. A rubber compression pad maintains pressure against the mating surfaces of the pot cores and inductance adjustment is obtained by a 4mm screw core. The components are all selected to operate under tropical conditions without deterioration of performance.

The construction permits multi winding techniques to be employed and sub-assembly to be carried out separately. The bending over of tags prior to dip soldering is facilitated by careful attention to detail and hot tin dipped tags are used to establish good soldering techniques.

The cores can be supplied in several grades of ferrite according to requirements, dictated by inductance required and frequency of operation. *Q*'s of 200 are readily obtainable at the lower frequencies and at 20Mc/s *Q*'s of 100 can be obtained. Work is being carried out with the object of improving the highest frequency *Q*'s. The pot cores have been used up to 40Mc/s.

The low frequency assemblies can be wound to give 11mH inductance up to 4Mc/s with correspondingly less as the material permeability is reduced for higher frequency performance.

The temperature coefficient of permeability can be made for the assembly less than 100 parts in 10^6/°C.

EE 9759 *for further details*

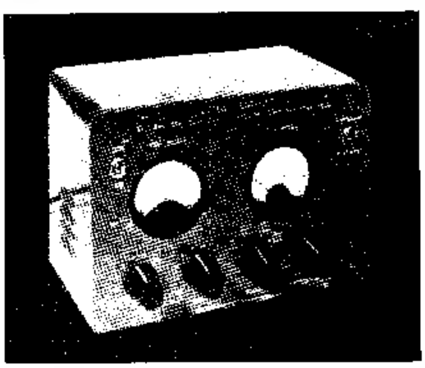
DIODE TESTER

(Illustrated below)

R. E. Thompson & Co. (Instruments) Ltd, Hersham Trading Estate, Walton-on-Thames, Surrey

This instrument covers current ranges from 50μA to 5A and voltages from 3V to 1·2kV. Current and voltage can be read simultaneously on separate meters.

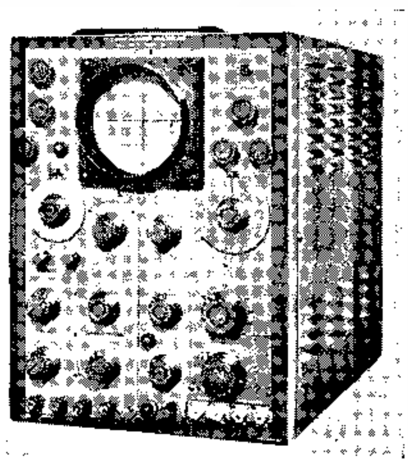
All semiconductor diodes, including



Zener types, can be fully tested as the ranges are operative for both forward and reverse characteristics. A forward-reverse switch reverses the polarity of the output.

A four position range switch controls both the current and voltage to the output terminals and also adjusts the appropriate meter scales, which reduces the chances of damaging either diode or meters. Subsidiary switches allow the voltage scale to be expanded by 4 times, and the current scale by up to 1 000 times (in three steps: 10, 100 and 1 000). Small lights indicate which range on each meter is in use for any setting of the switches. The voltage or current at any position of the range switch is fully adjustable by means of coarse and fine controls. These fully cover every range in ten coarse steps with an infinitely variable fine control.

EE 9760 *for further details*



TELEVISION MEASURING OSCILLOSCOPE

(Illustrated above)

Marconi Instruments Ltd, St. Albans, Hertfordshire.

THIS special-purpose oscilloscope, type TF 1277, enables precise measurements to be made on the waveforms used for testing television broadcast equipment, transmitters and links. It is suitable for 405-525- and 625-line systems, colour or monochrome, or for use as a studio waveform monitor or high-grade general-purpose oscilloscope.

The instrument is particularly suitable for the quantitative assessment of the quality of a television picture in terms of its *K* rating. The accuracy with which it displays the test waveforms used for determining picture quality allows measurement of *K* factors down to values well below that normally accepted as a reasonable design objective.

Either free-running or triggered sweep can be selected, and the start of the display can be delayed by a variable amount. The time-base can also be switched to allow the display of pulse and bar waveforms, or to provide a line-strobe facility with a marker to modulate a picture monitor so that the line required for analysis can be readily selected from the raster.

The d.c. coupled Y amplifier has a substantially flat response from d.c. to 10Mc/s with a maximum sensitivity of 100mV/cm; by using the a.c. coupled pre-amplifier, a sensitivity of 5mV/cm is obtained in a pass-band of 3c/s to 10Mc/s.

Two coaxial Y-input sockets are provided; these can be selected independently or connected for differential measurements. On unbalanced inputs, any hum present between the earth systems of the signal source and the oscilloscope is cancelled out, allowing residual sub-carrier or noise to be measured down to −50dB on a 1V signal.

The response of the X amplifier extends from d.c. to 1Mc/s. The expansion can be varied from ×1 to ×50, with a sweep width of 10cm at the minimum setting.

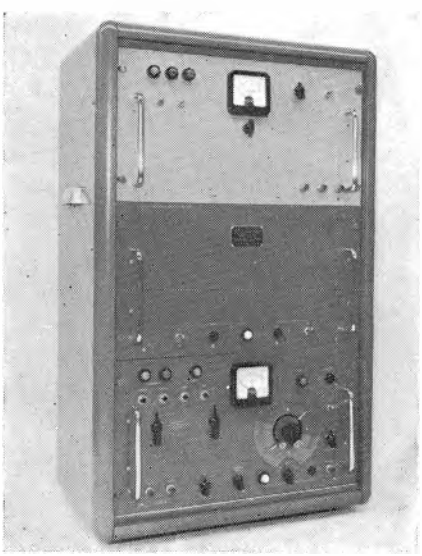
EE 9761 *for further details*

SWEEP OSCILLATOR AND VIBRATION CONTROLLER

(Illustrated below)

Sir W. G. Armstrong Whitworth Aircraft Ltd, Baginton, Coventry

This unit is designed to drive vibrator amplifiers. It has a frequency range from 10c/s to 32kc/s and accurate control of peak acceleration up to ±40*g* is provided. The automatic frequency sweep facility is provided with 7 sweep speeds for each frequency range. The sweep can be arranged to reverse or stop at any desired point, and its limits of travel are adjustable anywhere in the range being used. Manual operation of push-buttons can override automatic control sweep at any time. Panel lights indicate the various modes of operation. A square wave output is provided to drive a stroboflood at oscillator frequency or at oscillator frequency plus a small increment of 0·5c/s, 1c/s or 2c/s. The vibration control section of the apparatus is designed to control the peak acceleration of a vibrator table by automatic adjustment of the vibrator amplifier input during mechanical or electrical resonances. Specifications other than 'constant *g*' can be followed by the insertion of an appropriate function amplifier between the pick off and comparator in the control loop.



The controller is capable of controlling in the range of 10c/s to 32kc/s. The upper frequency limit, however, is determined by the vibrator equipment. It will control up to ±40*g* or as determined at low frequencies by the excursion of the table.

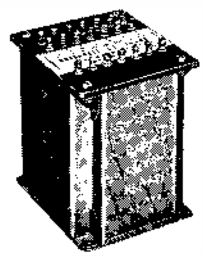
EE 9762 *for further details*

UNIVERSAL L.T. TRANSFORMER

(Illustrated below)

Radiospares Ltd, 4-8 Maple Street, London, W.1

These universal l.t. transformers are designed as a source of 'odd' voltages in laboratories and workshops. They are of a fully shrouded potted construction and all connexions are to turret lugs on a high grade laminated panel. There are



seven secondary windings and five tappings on the primary. By suitably connecting the secondary windings in series and parallel combinations all voltages from 1V to 40V in steps of 1V may be obtained. The total loading on the transformer must not exceed 90W.

EE 9763 *for further details*

AUTOMATIC NUCLEAR COUNTING SYSTEM

(Illustrated above right)

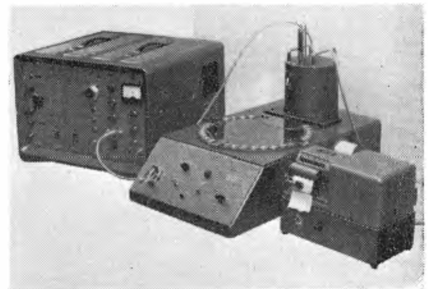
Società Elettronica Lombarda, Via Teodosio 70, Milan

The system is composed of a fast scaler, an automatic sample changer of simple and sturdy design and a time printing machine.

The scaler, of modular construction, comprises an e.h.t. supply unit (300 to 2 000V, stability ±0·05 per cent, resettability ± 0·1 per cent); a non-overloading input amplifier unit (gain — 100, risetime better than 0·2μsec); a fast discriminator, or a single channel analyser unit 5 to 100V threshold, 1, 2, 4 or 8V channel width); three electronic decades with 1μsec resolving time, and lastly the automatic counting unit (auto-time from 10 to 9 999sec in 1sec steps; auto-count from 1 000 to 10^6).

The automatic sample changer, of patented design, is mechanically very simple and sturdy; it takes 30 samples on 1in planchets or 12 samples on 2in planchets.

It is possible to use three different detectors by replacing the detector head, namely an end window G.M. tube; an ultra thin window (0·9mg/cm²) flow G.M. or proportional tube; a scintillation coun-



ter with alfa, beta, or gamma phosphors.

The G.M. counting gas used with G.M. flow tubes is an inexpensive mixture of argon-buthane.

The modular construction not only makes maintenance very easy, but also permits successive extensions or variations of the system while avoiding any obsolescence.

EE 9764 *for further details*

PROPORTIONAL TEMPERATURE CONTROLLER

(Illustrated below)

Ether Ltd, Caxton Way, Stevenage, Hertfordshire

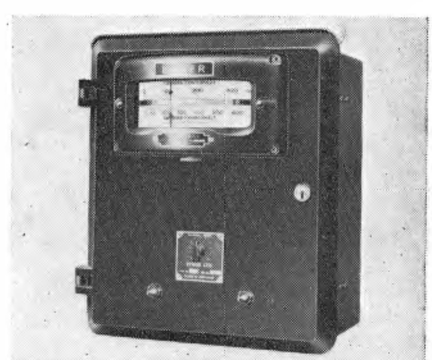
Ether Limited have now extended their range of Transitrol temperature controllers, to a total of six, with the introduction of a continuously-acting proportional controller with reset feature, known as the type 995.

The type 995 Transitrol, which has a 7in indicating scale, operates in conjunction with a proportioning-motorized valve and affords proportional control, to very close accuracy, on gas or oil-firing applications. It can also maintain steam at a constant pressure under varying loads.

Temperature control is attained by the use of a conventional galvanometer as the measuring system. An indicator-pointer operates a simple photo-electric system which, in turn, controls the valve-positioning motor. The latter is powered by a 24V, 50c/s, shaded-pole, reversing motor comprising two stators and two rotors, coaxially mounted on a common shaft which tranmits power to the final drive shaft through a silent and accurately-machined gear train.

The entire instrument, which is suitable for wall or panel-mounting, is designed as a plug-in unit and is easily withdrawn from its casing.

EE 9765 *for further details*



Short News Items

The Central Electricity Generating Board are to make application to the Minister of Power and to the Local Planning Authority for consent to develop a site of about 175 acres at Oldbury-on-Severn for a nuclear power station. The site lies on the south eastern bank of the Severn Estuary fourteen miles north of Bristol and four-and-a-half miles down-stream of Berkeley where one of the Board's first two nuclear power stations is under construction. Advantage will be taken of a rock shelf off-shore to create an artificial tidal reservoir from which the large quantities of water required for cooling the condensers of the steam turbines can be drawn during low tide periods. This unusual arrangement will reduce the costs that would normally be incurred in obtaining cooling water from an estuary with as wide a tidal range as the Severn. Foundation conditions on shore are good for supporting heavy reactors. Present indications are that a station on this site might have a total output of about 1 000MW and be connected to the 275kV supergrid.

Standard Telephones and Cables Limited have received a contract from the Royal Board of Swedish Telecommunications for a microwave network in Europe worth approximately three-quarters of a million pounds sterling. This provides for a propagation survey of the proposed routes to determine suitable sites for the equipment stations, and for the supply and installation of microwave equipment for operation in the 4 000Mc/s band. Stations at Sundsvall and Boden will be linked to form an extensive network for the transmission of telephone and television signals. The system will join the large communications network of the south of Sweden to that in the north by connecting the Stockholm-Sundsvall coaxial system with Boden, whence a coaxial system extends to Kiruna. This coaxial and microwave equipment will be in service over nearly the whole of the route from Malmo in the south to Kiruna inside the Arctic Circle in the north, a distance of over 1 600km (1 000 miles).

The British Computer Society Limited will hold its first Conference in Cambridge from 22 to 25 June, 1959, which has been designed to appeal to all users of digital computers, whether their interests lie in the business or the scientific field. Members and others who wish to attend can obtain application forms from the Conference Secretary, University Mathematical Laboratory, Corn Exchange Street, Cambridge.

The Second International Conference on Medical Electronics will be held at the new UNESCO Building in Paris from Wednesday, 24 June, at 2 p.m. to Saturday, 27 June, at 6 p.m. It is hoped particularly to bring together doctors and research workers with unsolved problems and engineers with new techniques and ideas for possible solutions. All requests for accommodation should be sent to S.O.C.F.I., I ter Rue chanez, Paris 16ᵉ. Further information can be obtained from the Secretariat, 131 Boulevard Malesherbes, Paris XVIIᵉ.

Brush Crystal Co. Ltd of Southampton announce that commercial production of new piezoelectric materials has now started. The materials are polycrystalline ceramics based on a lead zirconate-titanate solid solution and they possess a very high conversion efficiency combined with a useable temperature range of up to 250°C. Applications for the range of piezoelectric ceramics include pick-up elements, accelerometer and pressure cartridges, flaw detection probes and high power ultrasonic transducers. The temperature characteristics of one type are of an order that makes the production of ceramic filter elements a practical possibility.

The University of Birmingham, Electrical Engineering Department, is arranging two informal residential conferences in September 1959. The first, on Dielectric Devices, will be held from 14-17 September and the second, on Modern Network Theory, from 21-24 September. Preliminary programmes of papers and discussions have already been arranged and those interested in attending are invited to write to the Secretary of the Electrical Engineering Department, The University, Birmingham 15, for full details.

The Department of Scientific & Industrial Research will award longer-term grants for special researches in future. These grants, which have normally been limited to a maximum of five years, may now be awarded for an initial period of up to seven years with the possibility of continuing support for a further ten years. The main conditions are that an initial award will be given for periods of three to seven years with the possibility of two renewals, each with a maximum duration of five years. A third renewal would not normally be given. Details of the scheme are contained in the 1959 edition of 'Notes on D.S.I.R. Grants', published by Her Majesty's Stationery Office in January.

The Institution of Electronics will hold their Fourteenth Annual Electronics Exhibition and Convention at the Manchester College of Science and Technology during the period 9-15 July, 1959. This Exhibition will consist of an extensive display of electronic devices of interest to members of all branches of science and industry, and will consist of a 'Manufacturer's Section' and a 'Scientific and Industrial Research Section'. A Convention will be incorporated consisting of a series of lectures and film shows on electronic topics. Further particulars may be obtained from the Honorary Exhibition Organizer, Mr. W. Birtwistle, 78 Shaw Road, Rochdale, Lancs.

MEETINGS THIS MONTH

BRITISH INSTITUTION OF RADIO ENGINEERS

All London meetings will be held at the London School of Hygiene and Tropical Medicine, Keppel Street, Gower Street, London, W.C.1, and will start at 6.30 p.m.

Date: 5 May.
Lecture: An Experimental Diode Parametric Amplifier and its Properties.
By: I. M. Ross, C. P. Lea-Wilson, A. J. Monk and A. F. H. Thomson.

Date: 13 May.
Lecture: Improving Communication Techniques—what have engineers to learn from information theory?
By: Professor D. Gabor.

South Midlands Section

Date: 1 May. Time: 7 p.m.
Held at: North Gloucestershire Technical College, Cheltenham.
Lecture: Transistor Amplifiers.
By: F. Butler.

INSTITUTION OF ELECTRICAL ENGINEERS

All London meetings, unless otherwise stated, will be held at the Institution, commencing at 5.30 p.m.

Radio and Telecommunication Section

Date: 13 May.
Lecture: The Application of Statistical Techniques to the Electronic Valve Industry.
By: E. G. Rowe.

Date: 14 May. Time: 6.30 p.m.
Lecture: On the Conceivable Future of Telecommunications.
By: Professor E. C. Cherry.

South Midland Centre

Date: 4 May. Time: 6 p.m.
Held at: James Watt Memorial Institute.
Lecture: Some Problems Associated with the Operation of Large Turbo-Alternators.
By: A. Abbott.

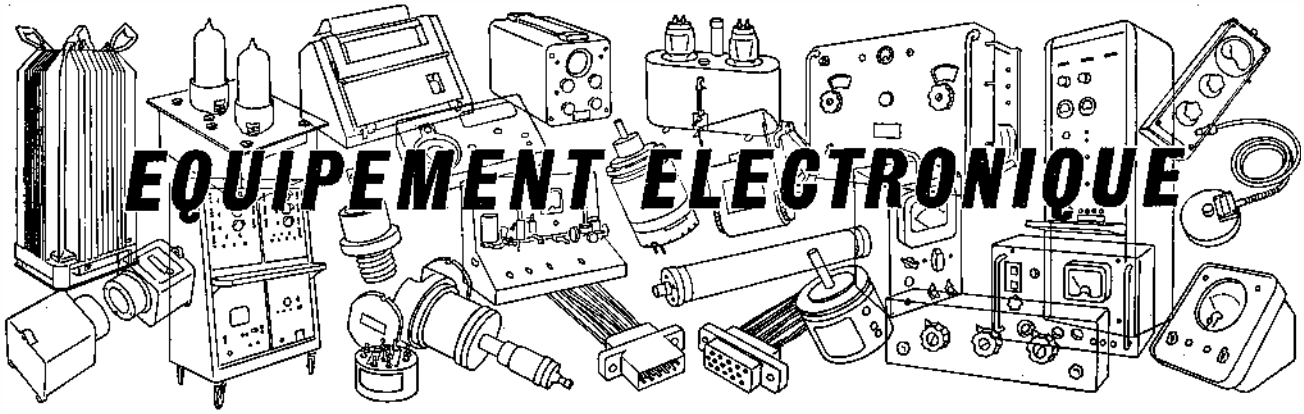
INSTITUTION OF ELECTRONICS

Date: 8 May. Time: 7 p.m.
Held at: Reynold Hall, Manchester College of Science and Technology.
Lecture: The Design of Transistor H.F. Amplifiers.
By: L. E. Jansson.

SOCIETY OF INSTRUMENT TECHNOLOGY

Date: 13 May. Time: 6 p.m.
Held at: Manson House, Portland Place, London, W.1.
Annual General Meeting of Data Processing Section followed by
Lecture: A Multipoint Digital Strain-Gauge Recorder.
By: J. R. Sturgeon.

(Pages 309 to 318 French and German section)



EQUIPEMENT ELECTRONIQUE

Une description basée sur des renseignements fourn is par les fabricants de nouveaux organes, accessoires et instruments d'essai *Traduction des pages 304 à 307*

VOLTMÈTRE DE CRÊTE

(*Illustration à la page 304*)

Wayne Kerr Laboratories Ltd, Chessington, Surrey

Les laboratoires Wayne Kerr, en collaboration avec le Commissariat britannique à l'énergie atomique (U.K.A.E.A.), viennent de mettre au point un voltmètre de crête de 50 kV pour la mesure d'impulsions simples ou répétées, d'une durée de 1 microseconde ou davantage.

L'impulsion reçue est appliquée à un réseau à condensateurs de division de tensions, d'un rapport de 50 : 1, et la tension parcourant le condensateur principal est mesurée au moyen d'un compteur électrostatique de 0 à 1 kV. Une diode à rétablissement de niveau shunte le compteur afin de permettre la mesure d'impulsions répétées. L'impulsion indiquée se décharge lentement par une résistance d'isolement supérieure à 10^{15} Ohms, le circuit de mesure étant hermétiquement fermé afin de maintenir ce chiffre d'isolement. La décharge est effectuée en pressant un bouton sur le panneau avant. Ce bouton actionne un relais à l'intérieur du compartiment scellé.

Deux gammes peuvent être obtenues facilement, soit 10 kV à 50 kV et 5 kV à 25 kV. Le degré de précision de la mesure est supérieur à $\pm 5\%$, en employant les courbes d'étalonnage fournies à cet effet, qui indiquent l'étalonnage initial jusqu'à l'espace sphérique irradié, selon la norme britannique 358.

EE 9751 *pour plus amples renseignements*

CONTRÔLEUR DE CÂBLES

(*Illustration à la page 304*)

Tyser & Co. Ltd, Merrow, Guildford, Surrey

Un nouvel instrument pour la détection rapide des défauts dans les câbles vient d'être mis au point et lancé sur le marché par la société Tyser & Co. Ltd (les successeurs de la société P.A.M. Ltd, au sein de la Southern Areas Electric Corporation).

Il s'agit du Contrôleur - indicateur - enregistreur PERRAM, dont la fonction est de déceler les défauts dans le guipage isolant des câbles en cours de fabrication. Il représente une méthode non-destructive d'évaluation de la qualité de l'isolement des câbles, qu'on peut ensuite classer en conséquence.

Il comprend un générateur de haute tension c.c. non-mortelle, dont la puissance de sortie à variation continue est de 4 à 20 kV c.c., ainsi qu'un compteur électromagnétique de grande vitesse, un indicateur d'alarme à lampe et sonnerie, et un circuit de protection de surcharge automatique.

La sortie de haute tension est appliquée à une électrode à petit anneau, employée pour examiner l'isolement du câble. Lorsqu'un défaut ou une rupture sont décelés dans l'isolement, la décharge de tension qui en résulte est enregistrée par le compteur.

Lorsqu'un défaut est localisé, le système d'avertissement est déclanché pour une durée minimum de 4 secondes, donnant une avertissement à la fois visible et audible. Une succession serrée de défauts sera toujours enregistrée par le compteur, même si le système d'avertissement ne suit pas la vitesse de répétition.

Dans le cas d'un défaut prolongé, le circuit de protection de surcharge automatique entre en action, débranchant le générateur de haute tension et gardant les indicateurs d'alarme en fonctionnement jusqu'à ce que l'opérateur remette l'instrument en état normal.

Le circuit permet une très grande vitesse de fonctionnement et l'instrument peut enregistrer une détection de défaut d'une durée aussi réduite que 1 m/sec. Ceci permet une vitesse d'avance de câble de 914 m/min., même en employant une électrode à anneaux très courte.

EE 9752 *pour plus amples renseignements*

> **LE SALON DU TRANSISTOR**
>
> **'Electronic Engineering' occupera le Stand 73 au Salon du Transistor, à Earls Court, Londres, du 21 au 27 mai 1959. Des détails complets sur les fabrications qui y seront exposées ne nous sont pas encore parvenus et ne peuvent donc être inclus dans le présent numéro. Un compte-rendu du Salon sera donné, cependant, dans un prochain numéro.**

OSCILLOSCOPE À MÉMOIRE DE SIGNAUX TRANSITOIRES

(*Illustration à la page 304*)

A. E. Cawkell, 6-8 Victory Arcade, Southall, Middlesex

Cet appareil, qui porte le nom de REMSCOPE, étend considérablement la série des mesures et observations réalisables sans recours à l'enregistrement photographique.

Les oscilloscopes à mémoire précédents, construits par la société Cawkell, faisaient usage du tube à rayons cathodiques MEMOTRON et leurs performances étaient limitées par le rendement relativement lent de ce type de tube. Avec l'avènement du tube à mémoire et image English Electric Type E702, cette restriction a disparu et l'oscilloscope REMSCOPE a donc été étudié pour faire l'usage le plus complet du nouveau tube.

Le tube E702 contient un système à mémoire qui permet l'emmagasinage de signaux transitoires pendant une semaine et leur représentation sous forme d'image pendant un court laps de temps durant cette semaine. Avec les REMSCOPE, la durée de l'image (qui peut être présentée fragmentairement) dépasse 15 minutes et peut, pour tous les signaux, sauf les plus rapides, être étendue à environ deux heures. La luminosité est quelque peu affectée par des temps de projection prolongés, quoiqu'on obtienne quand même une image satisfaisante. Afin de pouvoir utiliser au maximum l'image "emmagasinée", la représentation sur l'écran peut être différée jusqu'au moment voulu, ou elle peut être branchée automatiquement à la réception d'un signal.

La persistance de la représentation peut varier de quelques secondes jusqu'à 15 minutes. Cet avantage est particulièrement utile lorsqu'on compare un détail d'une série avec un détail précédent. En réglant la persistance, on peut faire disparaître le premier signal tandis qu'apparaît le troisième, et ainsi de suite. En variant le réglage on permet la production d'autres séquences.

L'image est rapidement supprimée en pressant un bouton intérieur ou en fermant un circuit externe. Une minuterie de suppression qui élimine le signal après un intervalle préréglé est prévue à cet effet.

A la livraison, l'appareil est muni d'un amplificateur Y à large bande avec entrées différentielles c.c. Afin de permettre son remplacement, par la suite, par des éléments différents, cet amplificateur est du type à prises.

Etant donné que l'oscilloscope REMSCOPE a été principalement conçu pour l'étude des signaux transitoires, un soin méticuleux a été apporté à rendre le déclanchement sûr et aisé. On peut employer des signaux en courant alternatif ou continu pour amorcer la base de temps. L'amorçage est automatique dès que le potentiel d'entrée dépasse un niveau préréglé.

En dehors des particularités décrites plus haut, l'appareil comprend tous les avantages usuels d'un oscilloscope de haute qualité. La base de temps comporte une gamme étendue de vitesses, de 3 cm/μsec à 0,1 cm/sec. et une amplification au quintuple peut être effectuée au moyen de l'amplificateur X. L'amplificateur Y a un temps de montée de 0,15 μsec, mais il est également muni d'un dispositif à grand gain et bande étroite. Les deux amplificateurs et les circuits de déclanchement sont entièrement à couplage direct. Des marqueurs de temps à contrôle piézoélectrique assurent l'étalonnage de la base de temps. Ils sont étroitement serrés afin de fournir un étalon de tension. La précision de mesure de la tension est $\pm 1\%$. Un glissement minimum est garanti par une haute tension stabilisée et des sources d'alimentation de chauffage.

EE 9753 *pour plus amples renseignements*

TUBES ÉCLAIR À RAYONS CATHODIQUES

(Illustration à la page 305)

Ferranti Ltd, Hollinwood, Manchester, Lancashire

Le Service d'Electronique de la Société Ferranti Ltd, vient de lancer une nouvelle gamme de tubes-éclair à rayons cathodiques qui produisent un éclat lumineux du même ordre d'intensité que celui des tubes éclair électroniques à gaz employés pour la photographie.

Ces tubes, appartenant aux séries CL60 et CL70, s'allument en entier, simultanément, par contraste avec le comportement normal des tubes à rayons cathodiques, et ils présentent l'avantage supplémentaire de pouvoir répéter les éclats beaucoup plus rapidement: on peut obtenir jusqu'à 1 million d'éclats lumineux par seconde. On peut également produire des éclats contenant une fine structure intérieure consistant en une multiplicité de plus petits éclats.

Trois types différents de tubes sont fournis actuellement. La série CL60, comprenant des triodes de 99 mm de diamètre; la série CL70, comprenant des tubes miniature de 25 mm de diamètre et la série CL1, qui comprend des diodes de 152 mm de diamètre. Ces tubes peuvent être revêtus de substances luminescentes de toutes les couleurs, y compris l'ultra-violet et l'infra-rouge. Il est également possible de produire une lumière "blanche", débarrassée de la structure à ligne spectrale que l'on trouve dans les lampes à décharge et qui s'étend sur des gammes spectrales spécifiées.

EE 9754 *pour plus amples renseignements*

RELAIS DE CONTACT À FILS

(Illustration à la page 305)

Woden Transformer Co. Ltd, Moxley Road, Bilston, Staffordshire

Le relais de contact à fils WODEN-T.S.P. a été réalisé en vue de répondre à la nécessité de pouvoir disposer d'un relais multipolaire rapide pour les calculateurs électroniques ainsi que pour les appareils de contrôle en général.

En principe, c'est un relais à prises pour supports multiples ou individuels qui peut être soudé directement ou employé avec une fiche à déclanchement rapide.

Son élément de base est un relais commutateur à quatre pôles avec contacts de fils jumelés, facilement interchangeables. Cet élément et tous les contacts fixes sont faits d'un moulage en une seule pièce, d'une matière de grande solidité diélectrique. Tous les contacts fixes sont en alliage d'argent.

La construction du relais permet un fonctionnement très rapide, les durées de déclanchement et de fonctionnement étant réduites à 3 microsecondes. Ce type de relais convient à diverses applications, mais il est particulièrement utile pour les calculateurs électroniques et les circuits de commutation vidéo, la faible capacité entre les contacts étant extrêmement utile.

Cet élément a une durée de contact prévue de 3 millions d'opérations et l'étendue de la durée de contact peut être considérablement accrue par l'usage de circuits éliminateurs d'étincelles ou par l'utilisation à des régimes plus faibles.

L'élément quadripolaire courant a environ les dimensions suivantes: 44,5 $\times$ 15,9 $\times$ 38 mm; il existe également des éléments à 6 et 12 pôles.

Ces relais sont étudiés pour être employés dans des températures ambiantes maxima de 50°C, mais ces températures peuvent être plus élevées pour certaines applications spéciales.

EE 9755 *pour plus amples renseignements*

FILTRE MULTI-OCTAVES BASSE FRÉQUENCE

(Illustration à la page 305)

Peekel, Laboratorium voor Electronics N.V., Alblasstraat 1, Rotterdam 8, Pays-Bas

Ce filtre permet d'obtenir une caractéristique de fréquence arbitraire dans les gammes des basses fréquences. Cette gamme s'étend de 30 Hz à 16 kHz et elle est divisée en bandes étroites par huit filtres, dont l'action peut être réglée individuellement au moyen d'atténuateurs étalonnés en décibels. Les atténuateurs sont étudiés de telle façon que la position de chacun d'eux peut être lue sur une échelle verticale et ils sont disposés de manière à former un système de coordonnées avec la fréquence en abscisses et l'atténuation en ordonnées, la position des huit atténuateurs indiquant ainsi, au coup d'oeil, la caractéristique de fréquence obtenue.

La pente moyenne de chaque filtre est de 30 dB/octave et chaque filtre peut être réglé entre 0 et -55 décibels, en paliers de $-2,5$ décibels. Lorsque tous les filtres sont ajustés au même niveau, on obtient une caractéristique de fréquence linéaire entre 30 Hz et 16 kHz. Lorsque les atténuateurs sont à la position 0 dB, le filtre agit comme amplificateur à gain 1.

L'impédance d'entrée et de sortie du filtre multi-octaves est de 200 ohms et la tension normale pouvant être appliquée à l'entrée est de 1 volt.

EE 9756 *pour plus amples renseignements*

VOLTMÈTRE ÉLECTRONIQUE

(Illustration à la page 305)

Brüel et Kjær, Nærum, Danemark

Le nouveau voltmètre électronique Type 2409 réalisé par la société Brüel & Kjaer peut être employé pour toutes sortes de mesures de tension dans la gamme de fréquences de 2 Hz à 200 kHz et dans la gamme de tensions d'environ 1 mV à 1 kV. Il est pourvu de deux échelles de tensions, à déviation totale, pour 10 volts et 31,5 volts respectivement. Une échelle de décibels est prévue pour mesurer les rapports d'amplification et d'atténuation. Lorsque l'échelle de décibels est employée avec l'atténuateur étalonné, toute mesure de tension peut être lue directement en décibels par rapport à 1 volt.

En outre, le voltmètre est muni d'une échelle de décibels (par rapport à 1 mW) pour les mesures de transmission basse fréquence, les lignes téléphoniques, etc. La valeur de référence est de 0,775 V, correspondant à 1 mW dans 600 ohms.

L'instrument se distingue aussi par le fait qu'il peut mesurer la crête, la moyenne arithmétique et la valeur efficace réelle du signal d'entrée.

Deux caractéristiques d'amortissement différentes peuvent être choisies afin d'obtenir sans difficulté des indications de voltmètre précises, tant pour les signaux à haute fréquence que pour ceux à basse fréquence. La première caractéristique se conforme à la norme pour les mesures de décibels. Quant à la seconde, elle est employée pour obtenir une déviation stable de l'aiguille du voltmètre lorsqu'on mesure les signaux de basse fréquence.

Enfin, en faisant usage d'un jack de sortie, cet instrument peut être utilisé comme amplificateur étalonné à basse impédance de sortie et amplification maximum de 60 décibels, variable en paliers de 10 décibels.

EE 9757 *pour plus amples renseignements*

OSCILLOSCOPE UNIVERSEL

(Illustration à la page 306)

Erskine Laboratories Ltd, Scalby, Scarborough, Yorkshire

L'Oscilloscope Erskine Type 5 est un instrument universel pour l'observation des formes d'ondes et la mesure de leur durée et de leur amplitude.

Il est léger et compact, tout en étant suffisamment aéré, et ses composantes sont d'accès extrêmement aisé.

Le tube cathodique (du type à quatre anodes) est placé horizontalement et il se voit à travers un miroir de 45°. Ce montage présente de multiples avantages. Il permet d'avoir un capot de vision ne dépassant pas les dimensions hors tout du couvercle. Il permet aussi une meilleure disposition des circuits internes. Le panneau avant n'est plus bourré de commandes et la profondeur de l'instrument ne dépend plus du tube cathodique.

Toutes les connexions nécessaires, d'entrée et de sortie, sont placées le long du bord inférieur du panneau avant et elles sont adjacentes aux commandes auxquelles elles se rapportent. Juste au-dessous du panneau, se trouve un tiroir de la longueur entière du panneau pour ranger les connecteurs. Il sert également d'entrée d'air de convection pour l'aération, la sortie nécessaire se trouvant au-dessus de la partie supérieure du panneau. Ce dernier est en pente, son bord supérieur se terminant à l'arrière et au-dessous du bord supérieur du cache-poussière.

L'Oscilloscope Type 5 est alimenté en haute tension stabilisée permettant la lecture directe sur micromètres étalonnés, quelles que soient les fluctuations de l'alimentation. Un circuit à retard permet de voir n'importe quelle partie d'une forme d'onde à répétition jusqu'à 20 m/sec. du bord de synchronisation. La base de temps comporte neuf vitesses préréglées entre 3 et 30 microsecondes. La sensibilité de la plaque X est de 30 volts/cm et celle de la plaque Y est de 45 volts/cm. L'amplificateur Y a une largeur de bande allant du courant continu à 4 MHz et une amplification de 100.

EE 9758 *pour plus amples renseignements*

ENSEMBLES MINIATURE DE NOYAUX EN POT

(Illustration à la page 306)

Neosid Ltd, Stonehills House, Welwyn Garden City, Hertfordshire

Cet ensemble, renfermé dans un boîtier carré de 12,7 mm de côté, a été étudié pour les circuits à transistors et, en particulier, les circuits imprimés. Le dispositif à cosses, placé à la base de l'appareil, est conforme aux avis de l'Institut britannique de normalisation, concernant les circuits imprimés. Les pinces de terre du boîtier ont la double fonction de mettre à la terre et de tenir tout l'ensemble. Un tampon compresseur en caoutchouc maintient la pression contre les surfaces d'accouplement des noyaux en pot, tandis que le réglage de l'impédance se fait par un noyau fileté de 4 mm. Les composantes ont toutes été prévues pour des conditions tropicales, sans que cela ne nuise au fonctionnement de l'ensemble.

La construction particulière de l'appareil permet de faire usage de la technique des enroulements multiples et d'effectuer séparément le sous-assemblage. La courbure des cosses, avant la soudure, par immersion est facilitée par le soin apporté aux détails, cependant que des cosses trempées dans un tain à chaud sont utilisées pour assurer une bonne soudure.

Les noyaux peuvent être fournis en ferrite de différentes qualités, selon les besoins, qui dépendent de l'impédance voulue et de la fréquence de fonctionnement. On peut facilement obtenir des facteurs de surtension de 200 aux basses fréquences, et des facteurs de 100 à 20 MHz. Les facteurs de plus hautes fréquences sont en cours d'amélioration. Les noyaux en pot ont été employés jusqu'à 40 MHz.

Les ensembles à basses fréquences peuvent être bobinés pour donner une impédance de 11 mH jusqu'à 4 MHz, ou baisser dans une proportion correspondante à mesure que la perméabilité du matériel est réduite pour un fonctionnement à une plus haute fréquence.

Le coefficient de perméabilité de température peut être réduit pour l'ensemble à moins de 100 parties dans 10^6/°C.

EE 9759 *pour plus amples renseignements*

CONTRÔLEUR DE DIODES

(Illustration à la page 306)

R. E. Thompson & Co. (Instruments) Ltd, Hersham Trading Estate, Walton-on-Thames, Surrey, Angleterre

Cet instrument comporte les gammes courantes de 50 microampères à 5 Ampères et les tensions de 3 volts à 1,2 kilovolts. Le courant et la tension peuvent être lus simultanément sur des compteurs individuels.

Toutes les diodes semi-conductrices, y compris les diodes du type Zéner, peuvent être pleinement contrôlées, car les gammes s'appliquent tant à la marche avant qu'inversée. Un commutateur "avant-inversée" inverse la polarité de la sortie.

Un commutateur de gammes à quatres positions contrôle tant le courant que la tension vers les bornes de sortie et règle également les échelles de compteur appropriées, ce qui réduit les risques d'endommagement des diodes ou des compteurs. Des commutateurs auxiliaires permettent d'étendre l'échelle de tension quatre fois plus et celle de courant jusqu'à 1 000 fois davantage (par 3 paliers: de 10, 100 et 1 000). Des petites lampes indiquent la gamme employée sur chaque compteur, pour n'importe quel réglage des commutateurs. La tension ou le courant, à n'importe quelle position du commutateur de gammes, peuvent être ajustés au moyen de commandes de réglage approximatif ou précis. Ces dernières sont prévues pour toutes les gammes, soit dix paliers de réglage approximatif et une commande précise, infiniment variable.

EE 9760 *pour plus amples renseignements*

OSCILLOSCOPE DE MESURES POUR LA TÉLÉVISION

(Illustration à la page 306)

Marconi Instruments Ltd, St. Albans, Hertfordshire, Angleterre

Cet oscilloscope spécial, type TF 1277, permet d'effectuer des mesures précises des formes d'ondes utilisées pour l'essai du matériel de télévision, des émetteurs et des coupleurs. Il convient aux systèmes à 405, 525 et 625 lignes, en couleur ou monochrome, ou à l'emploi comme contrôleur de formes d'ondes de studio, ou bien encore comme oscilloscope universel de haute qualité.

Il est tout particulièrement indiqué pour déterminer quantitativement la qualité d'une image de télévision en fonction de son indice K. La précision avec laquelle il reproduit les formes d'ondes d'essai utilisées pour évaluer la qualité de l'image, permet de mesurer les facteurs K jusqu'à des valeurs bien au dessous de celles généralement admises comme objectifs raisonnables d'images. On peut choisir le balayage libre ou déclenché et le début de la projection peut être retardé dans une mesure variable. On peut aussi brancher la base de temps pour permettre la projection de formes d'ondes d'impulsions ou en barres, ou pour fournir un système à traces linéaires avec marqueur pour moduler un contrôleur d'image, de façon à ce que la ligne voulue pour l'analyse puisse être prise facilement de la trame.

L'amplificateur Y à courant continu a un rendement sensiblement plat, du courant continu à 10 MHz, avec une sensibilité maximum de 100 mV/cm. En employant l'amplificateur à courant alternatif, on obtient une sensibilité de 5 mV/cm dans une bande passante de 3 Hz à 10 MHz.

L'appareil est pourvu de deux raccords coaxiaux d'entrée Y qui peuvent être choisis indépendamment ou branchés pour des mesures différentielles. Dans les entrées déséquilibrées, tout ronflement existant entre les prises de terre de la source de signaux et l'oscilloscope est éliminé, permettant de mesurer tout bruit ou onde porteuse résiduels jusqu'à −5 décibels sur un signal de 1 volt.

Le rendement de l'amplificateur X s'étend du courant continu à 1 MHz. L'étalement peut être varié de ×1 à ×50, avec une largeur de balayage de 10 cm au réglage minimum.

EE 9761 *pour plus amples renseignements*

OSCILLATEUR DE BALAYAGE ET CONTRÔLEUR DE VIBRATIONS

(Illustration à la page 307)

Sir W. G. Armstrong Whitworth Aircraft Ltd, Baginton, Coventry

Cet appareil a été conçu pour la commande des amplificateurs à vibreur. Sa gamme de fréquences s'étend de 10 Hz à 32 kHz et il permet un réglage précis de l'accélération de crête jusqu'à ±40 *g*. Le dispositif de balayage automatique est prévu pour sept vitesses de balayage, pour chaque gamme de fréquences. Le balayage peut être réglé de manière à s'arrêter ou aller en sens inverse, à l'instant voulu, et les limites de sa course peuvent être ajustées en n'importe quel point de la gamme employée. Le balayage à contrôle automatique peut être remplacé, à n'importe quel moment, par des boutons-poussoirs actionnés à la main. Les lumières sur le panneau indiquent les différents modes de fonctionnement. Une sortie d'ondes carrées fournit une projection stroboscopique à la fréquence d'oscillateur ou à la fréquence d'oscillateur plus une légère augmentation de 0,5 Hz, 1 Hz ou 2 Hz. La partie de l'appareil qui contrôle les vibrations est prévue pour la commande de l'accélération de crête d'une table à vibreur, au moyen du réglage automatique de l'entrée de l'amplificateur à vibreur lors de résonances mécaniques ou électriques. Des spécifications autres que la "constante g" peuvent être suivies par l'insertion d'un amplificateur de fonction adéquat entre le sélecteur et le comparateur de la boucle de contrôle.

Le Contrôleur de vibrations est étudié pour le contrôle dans la gamme de 10 Hz à 32 kHz. La limite de fréquences maximum est déterminée, cependant, par l'équipement à vibreur. Son contrôle s'exercera jusqu'à ±40 *g* ou, selon l'excursion de la table, aux basses fréquences.

EE 9762 *pour plus amples renseignements*

TRANSFORMATEUR UNIVERSEL B.T.

(*Illustration à la page 307*)

Radiospares Ltd, 4-8 Maple Street, London, W.1

Ce transformateur universel a été étudié pour servir de source de tensions "non courantes" dans les laboratoires et ateliers. Il est de construction compacte et entièrement blindé. Toutes les connexions sont sur cosses-tourelles d'un panneau laminé de haute qualité. Ils sont munis de sept enroulements secondaires et de cinq prises de tension primaire. Par le branchement en série des enroulements secondaires et en effectuant des combinaisons parallèles, on peut obtenir toutes les tensions de 1 volt à 40 volts, en paliers de 1 volt. La charge totale sur ces transformateurs ne doit pas dépasser 90 watts.

EE 9763 *pour plus amples renseignements*

SYSTÈME DE COMPTAGE NUCLÉAIRE AUTOMATIQUE

(*Illustration à la page 307*)

Società Elettronica Lombarda, 70 via Teodosio, Milan

Ce système se compose d'une échelle d'un changeur d'échantillons automatique, d'un modèle simple et robuste, et d'une machine à imprimer horaire.

L'échelle, de construction entièrement modulaire, comprend un bloc d'alimentation très haute tension (300 à 2 020 volts, stabilité ±0,05%, réenclanchement ±0,1%); un amplificateur d'entrée ne surchargeant pas (gain: 100, temps de montée supérieur à 0,2 microsecondes); un discriminateur rapide ou un analyseur à voie unique (seuil de 5 à 100 volts, largeur de voie: 1, 2, 4 ou 8 volts); trois décades électroniques à temps de résolution de 1 microseconde et, enfin, l'élément de comptage automatique (auto-temps: de 10 à 9 999 secondes, en paliers de 1 seconde; auto-comptage de 1 000 à 10^6).

Le changeur d'échantillons automatique, d'un modèle breveté, est d'une grande simplicité mécanique et fort robuste. Il peut recevoir 30 échantillons sur planchettes de 25 mm ou 12 échantillons sur planchettes de 50 mm.

Il est possible d'employer trois détecteurs différents en remplaçant simplement la tête de détecteur, soit: un tube de Geiger à vitre d'extrémité; ou un tube proportionnel de Geiger à vitre ultra-mince (0,9 mg/cm^2); ou encore un compteur de scintillations avec substances phosphorescentes alfa, beta ou gamma.

Le gaz de comptage Geiger employé dans les tubes de débit Geiger est un mélange peu coûteux d'argon et de butane.

La construction modulaire a, non seulement l'avantage de rendre l'entretien fort aisé, mais elle permet également des extensions ou variations successives du système, en lui évitant ainsi de tomber en désuétude.

EE 9764 *pour plus amples renseignements*

CONTRÔLEUR PROPORTIONNEL DE TEMPÉRATURE

(*Illustration à la page 307*)

Ether Ltd, Caxton Way, Stevenage, Hertfordshire

La société Ether Limited vient de porter son assortiment de contrôleurs de température TRANSITROL à un total de six modèles, en présentant son contrôleur proportionnel Type 995 à action continue et réenclancheur.

Le TRANSITROL 995 qui est doté d'une échelle indicatrice de 178 mm. fonctionne conjointement avec une vanne à moteur de dosage et assure le contrôle proportionnel, à un degré de précision très avancé, d'installations à gaz ou à chauffage aux huiles combustibles. Il peut également maintenir la vapeur à une pression constante sous des charges différentes.

Le contrôle de la température s'effectue par l'emploi d'un galvanomètre courant comme système de mesure. Un indicateur à aiguille actionne un simple système photoélectrique qui, à son tour, commande le moteur régulateur de la vanne. Ce dernier consiste en un moteur inversible à enroulement en court-circuit de 24 volts, 50 Hz, comprenant deux stators et deux rotors, montés coaxialement sur un arbre commun qui actionne l'arbre de commande final au moyen d'un train d'engrenages silencieux et usinés avec précision.

L'appareil complet, qui peut être monté sur tableau ou sur mur, est construit sous forme d'élément à prises et peut être facilement retiré de son boîtier.

EE 9765 *pour plus amples renseignements*

Résumés des Principaux Articles

L'application des circuits imprimés aux faisceaux hertziens à large bande. par R. Rowland

Résumé de l'article aux pages 256 à 261

L'auteur indique la possibilité d'appliquer la technique des circuits imprimés aux études d'équipement à fréquences intermédiaires et porteuses pour faisceaux hertziens à large bande. Il souligne les avantages de cette application en démontrant que, non seulement la production de ces éléments coûte moins cher, mais qu'on obtient également un meilleur rendement, une stabilité accrue et plus d'uniformité. De plus, l'entretien est simplifié et les besoins en appareils d'essais sont considérablement réduits. Il décrit, en conclusion, quelques éléments types de circuits imprimés.

Transistors à avalanche: Une évaluation de leurs propriétés et de leurs utilisations par R. C. V. Macario

Résumé de l'article aux pages 262 à 267

Les résultats d'un examen des propriétés de certains transistors expérimentaux à jonction alliée, dits "d'avalanche", ont permis d'évaluer leurs caractéristiques. Ces dernières ont un rapport bien déterminé avec les propriétés des jonctions semi-conductrices, ce qui est fort utile lorsqu'on fait l'étude d'un dispositif selon une spécification donnée. L'auteur donne des détails d'études et résume quelques applications.

Un électrocardiographe enregistreur à réponse rapide par A. W. Melville et J. B. Cornwall

Résumé de l'article aux pages 268 à 271

L'instrument décrit dans cet article a été étudié pour l'emploi en salle d'opération. Il assure une excellente discrimination anti-parasites et comporte un circuit de contrôle d'amplitude d'impulsions des plus efficaces, ainsi qu'un compteur à constantes de temps alternatives, fournissant la réponse rapide nécessaire à l'évaluation d'incitations motrices à effet immédiat.

Transistors et noyaux de circuits de comptage. par F. Rozner et P. Pengelly

Résumé de l'article aux pages 272 à 274

Le noyau magnétique à boucle carrée et le transistor sont aujourd'hui des éléments de circuits d'usage courant dans les calculateurs arithmétiques. Cet article indique le moyen de combiner ces deux éléments de manière à former des circuits de comptage sûrs et non critiques, pouvant recevoir des fréquences d'entrée allant jusqu'à 750 kHz. Il décrit également une modification permettant d'utiliser le circuit comme enregistreur de décalage.

L'accumulation de charge dans un transistor de jonction durant la mise hors circuit de la région active. par R. S. C. Cobbold

Résumé de l'article aux pages 275 à 277

En résolvant l'équation à diffusion unidimentionnelle d'un transistor de jonction, on obtient les équations des courants d'émission et d'arrêt existant dans des conditions de mise hors circuit de durée minimum. Ces équations s'accordent sensiblement, ainsi qu'il est démontré dans cet article, avec les résultats pratiques.

Noyaux magnétiques à boucles d'hystérésis rectangulaires employés comme éléments de commutation. par J. F. Kaposi

Résumé de l'article aux pages 278 à 283

Les noyaux magnétiques à boucles d'hystérésis quasi rectangulaires peuvent être utilisés comme éléments de commutation ou d'accumulation car ils possèdent deux états stables de magnétisation rémanente. La boucle parcourue lors de la commutation d'un noyau à un autre dépend, d'une part, des propriétés magnétiques du noyau et, d'autre part, de la forme d'onde de commutation. Cet article examine les bases théoriques de la commutation des noyaux par l'emploi d'impulsions de tension et de courant, ainsi que les résultats expérimentaux de la commutation de différents types de noyaux et quelques applications particulières.

Mesure précise de la fréquence de coupure de transistors par Yasuo Tarui

Résumé de l'article aux pages 284 à 287

Cet article décrit un appareil pour mesurer rapidement et avec précision le gain de courant ainsi que le α de transistors haute fréquence, dans une gamme de fréquences allant de 1 à 20 MHz. La valeur absolue et la phase de α peuvent être mesurées avec une précision de 1,5%. On peut aussi déterminer rapidement la fréquence de coupure du courant.

Un équipement terminal radiotéléphonique miniature. par E. J. Allen

Résumé de l'article aux pages 288 à 290

Cet article décrit un équipement terminal radiotéléphonique qui, bien que de dimensions considérablement plus réduites que les installations ordinaires, assure un rendement comparable et permet la transmission simultanée de quatre voies téléphoniques commerciales.

Un amplificateur électrophysiologique à l'usage des étudiants. par P. E. K. Donaldson

Résumé de l'article aux pages 290 à 291

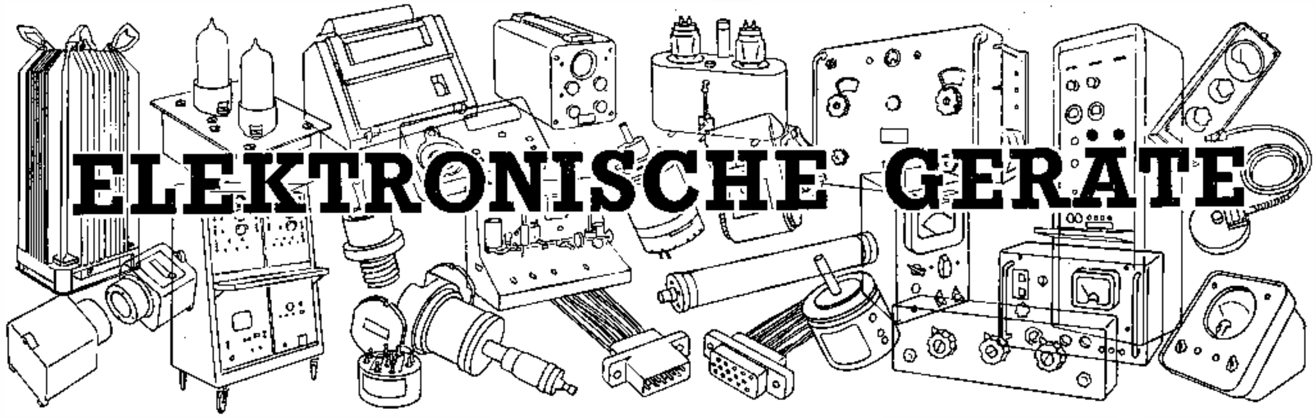
L'auteur décrit un amplificateur compact à un seul côté pour l'enregistrement du potentiel d'action de nerfs ou de muscles excisés. L'appareil fonctionne entièrement sur alimentation secteur, ce qui, ajouté à sa simplicité, en fait un accessoire fort utile à tout laboratoire biologique.

Un nouveau type de compteur d'anneaux. par P. J. Westoby

Résumé de l'article aux pages 295 à 298

Les compteurs d'anneaux à plusieurs étages sont d'un usage des plus fréquents. Le nombre d'étages dans les modèles courants est généralement limité à 12, pour les raisons que définit l'auteur. Il décrit, ensuite, un nouveau système permettant l'emploi de n'importe quel nombre d'étages, sans pour celà rencontrer les inconvénient des réalisations précédentes. Il explique également le principe de fonctionnement du système et ajoute quelques suggestions de dessin simple.

ELEKTRONISCHE GERÄTE



Beschreibungen neuer Bauelemente, Zubehörteile und Prüfgeräte auf Grund der von Herstellern gemachten Angaben. *(Ubersetzung der Seiten 304 bis 307)*

SPITZEN-SPANNUNGSMESSER

(Abbildung Seite 304)

Wayne Kerr Laboratories Ltd, Chessington, Surrey

In Zusammenarbeit mit der englischen Kernenergiebehörde haben die Wayne Kerr Laboratories einen 50 kV Spitzen-Spannungsmesser für einzelne oder wiederholte Impulse von mindestens 1 µsec Dauer entwickelt. Der Eingangsimpuls wird auf einen kapazitativen Spannungsteiler mit einem 50 : 1 Verhältnis gegeben und die Spannung des grossen Kondensators mit einem 0-1 kV elektrostatischen Instrument gemessen. Eine das Instrument überbrückende Diode ermöglicht die Messung aufeinanderfolgender Pulse durch Wiedereinstellung des Pegels. Der angezeigte Impuls wird langsam durch einen Isolierwiderstand von mindestens 10^{15} Ohm entladen. Um diesen Wert aufrecht zu erhalten, ist der ganze Messkreis luftdicht abgeschlossen. Betätigung eines Druckknopfes auf der Bedienungstafel führt zur Entladung durch ein Relais im luftdichten Abteil.

Die beiden Messbereiche 10 kV-50 kV und 5 kV-25 kV sind leicht einstellbar. Bessere Messgenauigkeiten als ±5% können unter Bezugnahme auf die Kurven der ursprünglichen Eichung mit einer überstrahlenden Kugelfunkstrecke nach der englischen Norm BS 358 erreicht werden.

EE 9751 *für weitere Einzelheiten*

KABELPRÜFER

(Abbildung Seite 304)

Tyser & Co. Ltd, Merrow, Guildford, Surrey

Tyser and Co. Ltd, Nachfolgefirma der P.A.M. Ltd in der Southern Areas Electric Corporation, hat ein neues Gerät für den Schnellnachweis von Kabelfehlern entwickelt und in das Fertigungsprogramm aufgenommen.

Es wird als der Perram Kabelprüfer, -anzeiger und -registrierapparat bezeichnet und wird zum Nachweis von Kabelmantelfehlern während der Fertigung eingesetzt. Es bietet eine zerstörungsfreie Methode für die Gütebeurteilung der Kabelisolation, die entsprechend eingestuft werden kann.

Das Gerät besteht aus einem Hochspannungs-Gleichstromgenerator, der nicht tödlich wirken kann, und dessen Ausgangsspannung stufenlos von 4 bis 20 kV regelbar ist, einem elektromagentischen Schnellzähler, einem Warnungsanzeiger mit Lampe und Klingel und einem Überlastungsschutz.

Die Hochspannung wird an eine kleine Lochelektrode gelegt, die für die Untersuchung der Kabelisolation benutzt wird. Jeder entdeckte Fehler oder Bruch in der Isolation ruft eine Spannungsentladung hervor, die durch den Zähler registriert wird.

Der gefundene Fehler betätigt das Warngerät für mindestens 4 Sekunden mit Sicht- und Höranzeige. Schnell hintereinanderfolgende Fehler werden durch den Zähler registriert, selbst wenn die Wernanlage der Fehlergeschwindigkeit nicht folgen kann.

Bei einem Dauerfehler wird der automatische Überlastungsschutz betätigt, der den Hochspannungsgenerator ausschaltet und die Warnanlage in Betrieb hält, bis Rückstellung durch den Maschinenwärter erfolgt.

Die Schaltung ermöglicht eine sehr hohe Arbeitsgeschwindigkeit, und ein nur für 1 msec auftretender Fehler wird registriert. Selbst mit einer sehr kurzen Lochelektrode können daher Kabelgeschwindigkeiten von 915 m/min erreicht werden.

EE 9752 *für weitere Einzelheiten*

DIE TRANSISTOR AUSSTELLUNG

Während der Transistor Ausstellung in Earls Court vom 21. bis 27. Mai werden Sie uns auf Stand Nr. 73 finden. Volle Einzelheiten der ausgestellten Fabrikate stehen noch nicht zur Verfügung und können deshalb nicht in diesem Heft beschrieben werden. Ein Bericht über die Ausstellung wird jedoch in einem der folgenden Hefte erscheinen.

SPEICHER-OSZILLOGRAF FÜR KURZLEBIGE VORGÄNGE

(Abbildung Seite 304)

A. E. Cawkell, 6-8 Victory Arcade, Southall, Middlesex

Das als „Remscope" bezeichnete Gerät erweitert wesentlich den Umfang der Messungen und Beobachtungen, die ohne fotografische Aufzeichnung möglich sind.

Cawkell haben früher die amerikanische „Memotron" Elektronenstrahlröhre für ihre Speicher-Oszillografen verwandt, deren Leistung durch die verhältnismässig langsam ansprechende Röhre begrenzt war. Jetzt steht jedoch die English Electric Speicher- und Sichtanzeigeröhre Type E 702 zur Verfügung. Dadurch fallen die Beschränkungen weg und, um von dieser Röhre den vollsten Gebrauch zu machen, wurde „Remscope" entwicklet.

Die Röhre E 702 enthält eine Speichervorrichtung, die die Aufbewahrung eines kurzlebigen Vorganges bis zu einer Woche ermöglicht. In dieser Woche kann der Vorgang für eine begrenzte Zeit auf dem Bildschirm dargestellt werden. Die Sichtanzeigezeit, die abschnittsweise Verwendung finden kann, übersteigt 15 Minuten und kann, ausser für die schnellsten Signale, bis auf nahezu zwei Stunden ausgedehnt werden. Die Helligkeit lässt bei den längsten Anzeigezeiten etwas nach, ist jedoch für die Sichtanzeige ausreichend. Um den besten Gebrauch von dem aufgespeicherten Bild zu machen, kann die Sichtanzeige aufgeschoben werden, bis sie benötigt wird, und in diesem Falle ist nur die Speichereinrichtung in Betrieb. Die Sichtanzeige kann durch ein Signal automatisch eingeschaltet werden.

Die Nachleuchtdauer ist von wenigen Sekunden bis zu 15 Minuten variable. Das ist besonders angenehm, wenn ein Vorgang in der Versuchsreihe mit seinem Vorgänger verglichen werden soll. Durch Einreglung der Nachleuchtdauer des ersten Signals kann man es erblassen lassen, wenn das dritte Signal erscheint usw. Eine Einstellungsänderung wird eine Reihenfolgeänderung nach sich ziehen.

Die Anzeige wird schnell durch einen eingebauten Druckschalter oder einen aussenliegenden Schaltkreis gelöscht. Ein anschliessbarer Löschsignalgeber mit einstellbarem Zeitintervall kann geliefert werden.

Zunächst wird ein Einkanal-Breitband-Y-Verstärker mit Gleichstrom-Differentialeingabe in einschiebbarer Form geliefert, um spätere Auswechslung gegen andere Verstärker möglich zu machen.

Da der Oszillograf in erster Linie für die Untersuchung kurzlebiger Vorgänge gedacht war, wurde zuverlässigem und einfachem Triggering besondere Aufmerksamkeit gewidmet. Die Betätigung kann durch Gleichstrom- oder Wechselstromsignale erfolgen und wird automatisch, sobald eine vorgewählte Spannung überschritten wird.

Ausser den angeführten sind auch alle zu einem hochwertigen Oszillgrafen gehörenden Einrichtungen vorgesehen. Der Ablenkteil hat einen grossen Geschwindigkeitsbereich, der von 3 cm/μsec bis zu 0,1 cm/sec geht. Durch den X-Verstärker wird eine fünffache Ausdehnung ermöglicht. Der Y-Verstärker hat eine 0,15 μsec Ansteigezeit, kann aber auch mit hoher Verstärkung in einem engen Bande arbeiten. Beide Verstärker und die Trigger-Schaltung sind durchweg direkt gekoppelt. Zur Eichung des Ablenkteiles vorgesehene quartzgesteuerte Zeitzeichengeber sind so verriegelt, dass sie einen Spannungsmassstab ergeben. Die Messgenauigkeit für Spannungen ist ±1%. Geringste Drift ist durch stabilisierte Hoch- und Heizungsspannungen gesichert.

EE 9753 *für weitere Einzelheiten*

DRAHTKONTAKT-RELAIS

(Abbildung Seite 305)

Woden Transformer Co. Ltd, Moxley Road, Bilston, Staffordshire

Das Woden-T.S.P. Drahtkontakt-Relais wurde eingeführt, um die Nachfrage der Rechner- und Reglerindustrie für ein mehrpoliges Schnellrelais zu befriedigen.

Ursprünglich ist es ein Einsteckrelais für Einzel- oder Mehrfachfassungen, die entweder direkt gelötet oder mit einem Schnellstecker versehen werden können.

Der Baustein besteht aus einem vierpoligen Umschaltrelais mit Doppel-Drahtkontakten, die leicht auswechselbar sind, und ist mit allen festen Kontakten in einen einteiligen Formling hoher Durchschlagsfestigkeit eingebaut. Alle feststehenden Kontakte werden aus einer Silberlegierung hergestellt.

Die Konstruktion des Relais erlaubt Schnellbetrieb mit Betätigungs- und Auslösezeiten, die bis zu 3 msec heruntergehen. Diese Type kann für Universalanwendung Gebrauch finden, hat aber besonderen Wert in Rechner- und Video-Schaltungen, wo die niedrige Kontakt-Kapazität besonders wünschenswert ist.

Die Kontaktlebensdauer des Relais ist voraussichtlich 3 000 000 Schaltungen. Die Kontaktlebensdauer kann durch Anwendung von Löschkreisen oder Unterbelastung wesentlich erhöht werden.

Die Abmessungen des vierpoligen Bausteines sind 44,5 mm × 15,9 mm × 38 mm. Sechs- und zwölfpolige Relais sind auch lieferbar.

Die Relais sind für eine maximale Umgebungstemperatur von 50° gebaut, können aber für Sonderverwendung höheren Temperaturen ausgesetzt werden.

EE 9755 *für weitere Einzelheiten*

RÖHRENVOLTMETER

(Abbildung Seite 305)

Brüel und Kjær, Nærum, Dänemark

Brüel und Kjær haben ein neues elektronisches Voltmeter Type 2409 entwickeelt, das für Spannungsmessungen aller Art von 1 mV bis 1 kV bei 2 Hz bis 200 kHz geeignet ist. Die beiden Spannungsskalen haben 10 V bzw 31,5 V Vollausschlag. Eine Dezibelskala ist für die Messung von Verstärkungs- oder Dämpfungsverhältnissen vorgesehen. In Verbindung mit einem geeichten Dämpfer können gemessene Spannungen direkt in Dezibel auf 1 V bezogen abgelesen werden.

Zusätzlich hat das Gerät eine dbm Skala für Messungen von NF Übertragungsleitungen, Fernsprechleitungen usw. Der Bezugswert ist 0,775 V und entspricht 1 mW in 600 Ohm.

Das Gerät kann Spitzenwerte, Mittelwerte und genaue Effektivwerte der Eingangssignale messen.

Zwei verschiedene Messgerätdämpfungen stehen auswahlsweise zur Verfügung, um die genaue Ablesung für hohe, als auch für niedrige Frequenzen zu erleichtern. Eine der Dämpfungskennlinien entspricht der Norm für VU (Spracheinheitsstärken) Messungen, während die andere für niedrige Frequenzen bestimmt ist und einen gleichbleibenden Ausschlag erzielt.

Bei Benutzung der Ausgangsklinke kann das Gerät als geeichter Verstärker eingesetzt werden, der bei geringer Ausgangsimpedanz eine 60 dB Höchstverstärkung gibt, die in 10 dB Stufen regelbar ist.

EE 9757 *für weitere Einzelheiten*

SCHNELLBLITZRÖHREN

(Abbildung Seite 305)

Ferranti Ltd, Hollinwood, Manchester, Lancashire

Die Abteilung Elektronik der Ferranti hat eine neue Kathodenstrahl-Schnellblitzröhrenreihe eingeführt, deren Blitzhelligkeit in derselben Grössenordnung liegt wie die der für fotografische Zwecke verwandten elektronischen Gasentladungslampen.

Diese Röhren führen die Serienbezeichnungen CL 60 und CL 70. Im Gegensatz zum Normalverhalten von Kathodenstrahlröhren leuchten sie überall gleichzeitig auf und haben den zusätzlichen Vorteil, dass Blitze viel schneller aufeinander folgen können. Blitzfolgen bis zu 1 000 000 per sec können erreicht werden. Lange Blitze können auch hervorgerufen werden, sowie solche mit besonders feiner Innenstruktur, die aus einer Mehrzahl kleinerer Blitze besteht.

Drei verschiedene Röhrentypen sind jetzt lieferbar: Die CL 60 Serie besteht aus Trioden mit einem 99 mm Durchmesser; die CL 70 Serie ist eine Miniaturausführung mit 25,4 mm Durchmesser, und die CL 1 Serie umfasst Dioden mit einem 152 mm Durchmesser. Die Phosphorschicht der Röhren kann in jeder Farbe hergestellt werden, einschliesslich ultraviolett und infrarot. „Weisses" Licht kann ohne die von Gasentladungslampen bekannte Spektrallinienstruktur abgegeben werden und erstreckt sich über einen vorherbestimmten Spektralbereich.

EE 9754 *für weitere Einzelheiten*

MEHROKTAVEN NF FILTER

(Abbildung Seite 305)

Peekel, Laboratorium voor Electronics N.V., Alblasstraat 1, Rotterdam 8, Pays-Bas

Mit Hilfe des Filters kann im NF Bereich ein willkürlicher Frequenzgang eingestellt werden. Der Frequenzbereich liegt zwischen 30 Hz und 16 kHz und wird durch 8 Filter in enge Frequenzbänder unterteilt. Jedes Filter kann unabhängig durch ein in Dezibel geeichtes Dämpfungsglied eingestellt werden. Die Dämpfungsglieder sind so konstruiert, dass der Einstellknopf auf einer vertikalen Skala den dB Wert anzeigt und gleichzeitig die Lage in einem Koordinatennetz bestimmt, in dem die Frequenz als Abszisse und die Dämpfung als Ordinate erscheinen. Die Einstellknöpfe der acht Dämpfungsglieder zeigen daher augenfällig den eingestellten Frequenzgang.

Die Durchschnittssteilheit jedes Filters ist 30 dB/Oktave, und jedes Filter kann von 0 bis −55 dB in −2,5 dB Stufen geregelt werden. Einstellung aller Filter auf denselben Wert ergibt einen geradlinigen Frequenzgang von 30 Hz bis 16 kHz. Bei Einstellung aller Dämpfungsglieder auf 0 dB wirkt das Gerät als Verstärker mit Einheits-Verstärkung.

Die Eingangs- und Ausgangsimpedanz des Mehroktavenfilters ist 200 Ohm und die Eingangsspannung üblicherweise 1 V.

EE 9756 *für weitere Einzelheiten*

UNIVERSAL-OSZILLOGRAF

(Abbildung Seite 306)

Erskine Laboratories Ltd, Scalby, Scarborough, Yorkshire

Der Erskine Oszillograf Modell 5 ist ein Universal-Instrument für die Beobachtung von Wellenformen und deren Zeit- und Amplitudenmessungen.

Das Gerät ist leicht und kompakt, trotzdem aber gut entlüftet, und die Bauelemente sind äusserst leicht zugänglich. Die Vier-Anoden-Kathodenstrahlröhre liegt horizontal und wird mit Hilfe eines 45° Spiegels beobachtet. Diese Anordnung hat mehrere Vorteile. Die Lichtabschirmung kann eingebaut werden und liegt innerhalb der Gehäuseabmessungen. Die Schaltungen können funktions mässiger angeordnet werden, auf der Bedienungstafel sind die Knöpfe nicht mehr so zusammengedrängt, und die Tiefe des Gerätes wird nicht mehr durch die Elektronenstrahlröhre bestimmt.

Die notwendigen Eingangs- und Ausgangsverbindungen sind an der unteren Kante der Bedienungstafel angeordnet und liegen neben den dazugehörigen Schaltungen. Direkt darunter ist ein Aufbewahrungsfach für Verbindungsschnüre, das so lang wie das Gehäuse ist und gleichzeitig als Einlass für die Konvektionsbelüftung dient. Der Auslass liegt über der Tafel, die geneigt ausgeführt ist und mit ihrer oberen Kante hinter und unter der Gehäusekante liegt.

Der Oszillograf hat eine stabilisierte Hochspannungsquelle, die direkte und von

Netzschwankungen unabhängige Ablesungen auf kalibrierten Rasterskalen ermöglicht. Durch eine Verzögerungsschaltung kann jeder Teil einer sich wiederholenden Wellenform bis zu 20 msec von der Synchronisierungskante beobachtet werden. Das Ablenkgerät hat neun vorgewählte Geschwindigkeiten zwischen 3 μsec und 30 msec. Die X-Plattenempfindlichkeit ist 30 V/cm und die der Y-Platten 45 V/cm. Der Y-Amplifier hat eine Bandbreite von Gleichspannung bis 4 MHz und einen Verstärkungsgrad von 100.

EE 9758 *für weitere Einzelheiten*

MINIATUR TOPFKERN-BAUSTEINE

(*Abbildung Seite 306*)

Neosid Ltd, Stonehills House, Welwyn Garden City, Hertfordshire

Der von einem rechteckigen 12,7 mm Abschirmbecher umschlossene Baustein wurde für Transistorenschaltungen und besonders für Verwendung mit gedruckten Schaltungen entwickelt. Der Stiftsockel entspricht den Empfehlungen des Britischen Normen Institutes für gedruckte Schaltungen. Die Erdschelle des Abschirmbechers hat die Doppelaufgabe eines Erdungs- und Befestigungselementes Ein Gummidruckpolster hält die aufeinanderliegenden Kernflächen zusammen, und ein 4 mm Gewindestift ist für die Spulenabstimmung vorgesehen. Die verwandten Bauelemente arbeiten unter tropischen Bedingungen ohne Leistungsverminderung. Mehrfachwickelmethoden werden durch die Konstruktion erleichtert, und Teilmontage kann getrennt ausgeführt werden. Sorgfältige Beachtung aller Einzelheiten während der Entwicklung erleichtert das Umbiegen der tauchverzinnten Stiftanschlüsse vor dem Tauchlöten und ergibt gute Lötstellen.

Je nach Anwendungsgebiet und entsprechend der Spuleninduktivität und Frequenz, können Ferritkerne in verschiedenen Werkstoffen geliefert werden. Für niedrigere Frequenzen kann ein Q = 200 erreicht werden, und für 4 MHz ist ein Q=100 möglich. Entwicklungsarbeiten für eine Güteverbesserung auf höchsten Frequenzen sind im Gange. Die Topfkerne werden bis zu 40 MHz verwandt.

Bausteine für niedrigere Frequenzen bis zu 4 MHz können für 11 mH gewickelt werden und für entsprechend geringere Werte, wenn die Werkstoffpermeabilität für höhere Frequenzen reduziert wird.

Für den Baustein kann der Temperaturkoeffizient der Permeabilität geringer als $1 \times 10^{-4}/°C$ gehalten werden.

EE 9759 *für weitere Einzelheiten*

DIODEN-MESSGERÄT

(*Abbildung Seite 306*)

R. E. Thompson & Co. (Instruments) Ltd. Hersham Trading Estate, Walton-on-Thames, Surrey

Die Strombereiche des Gerätes gehen von 50 μA bis 5 A und die Spannungsbereiche von 3 V bis 1,2 kV. Strom und Spannung können gleichzeitig auf verschiedenen Messern abgelesen werden. Alle Halbleiterdioden, einschliesslich Zener-Typen, können vollständig geprüft werden, da die Bereiche sowohl für Fluss- als auch Sperrkennlinien eingesetzt werden können. Ein Fluss/Sperrung Umschalter kehrt die Ausgangspolarität um.

Ein vierstelliger Schalter kontrolliert Strom und Spannung an den Ausgangsklemmen und stellt die zugehörigen Messbereiche ein, wodurch Dioden und Messgeräte einer geringeren Beschädigungsgefahr ausgesetzt werden. Hilfsschalter ermöglichen die Erweiterung der Messbereiche und zwar für die Spannung auf das vierfache und für den Strom auf das 1 000 fache (in drei Stufen: 10, 100, 1 000). Kleine Signallampen zeigen die in Betrieb befindlichen Messbereiche an. Strom und Spannung können bei jeder Schalterstellung mit Grob- und Feinabstimmung geregelt werden. Die Reglung erfolgt über den eingeschalteten Bereich mit zehnstufiger Grobabstimmung und stufenloser Feinreglung.

EE 9760 *für weitere Einzelheiten*

OSZILLOGRAF FÜR FERNSEH-MESSUNGEN

(*Abbildung Seite 306*)

Marconi Instruments Ltd, St. Albans, Hertfordshire

Der Sonderzweckoszillograf Type TF 1277 ermöglicht genaue Messungen der für die Prüfung von Fernseheinrichtungen—sendern und—funkbrücken gebräuchlichen Wellenformen. Das Gerät kann für Farb- und Schwarz-Weiss-Fernsehen mit 405, 525 oder 625 Zeilen eingesetzt werden und auch als Studio-Wellenformmonitor oder hochwertiger Universaloszillograf Verwendung finden.

Das Gerät ist besonders für quantitative Gütebeurteilung eines Fernsehbildes nach dem K-Faktor geeignet. Die Genauigkeit, mit der die notwendigen Prüfwellenformen dargestellt werden, ermöglicht die Messung von K-Werten, deren Grösse unter den Werten liegt, die normalerweise als annehmbar für die Entwicklungsziele bezeichnet werden.

Freilaufende und getriggerte Ablenkung sind wahlweise vorhanden, und der Beginn der Sichtanzeige kann um einen veränderlichen Wert verzögert werden. Der Ablenkteil kann auch so geschaltet werden, dass die Darstellung von Puls- und Stangenwellenformen möglish ist, oder dass ein Zeichengeber in einem Zeilenstroboskop eine zu untersuchende Zeile im Monitorbild so moduliert, dass sie leicht im Raster erkenntlich ist.

Der gleichstromgekoppelte Y-Verstärker hat einen im wesentlichen geradliniegen Frequenzgang von Gleichstrom bis 10 MHz mit einer Höchstempfindlichkeit von 100 mV/cm. Eine Empfindlichkeit von 5 mV/cm wird bei Verwendung eines wechselstromgekoppelten Vorverstärkers über einen Bereich von 3 Hz bis 10 MHz erreicht.

Zwei koaxiale Y-Eingangssockel können entweder unabhängig oder für Differenzmessungen Verwendung finden. Bei unsymmetrischen Eingängen wirde jede durch die Erdung der Signalquelle und des Oszillografen auftretende Brummspannung ausgeglichen und dadurch die Messung von rückständigen Hilfsträgern und Geräuschen bis zu-50 dB hinab auf einem 1 V Signal ermöglicht.

Der Frequenzgang des X-Verstärkers erstreckt sich von Gleichstrom bis 1 MHz. Die Dehnung ist zwischen 1x und 50x variable mit einer Ablenkbreite von 10 cm für die Mindesteinstellung.

EE 9761 *für weitere Einzelheiten*

ABLENKUNGSOSZILLATOR UND VIBRATIONSREGLER

(*Abbildung Seite 307*)

Sir W. G. Armstrong Whitworth Aircraft Ltd, Baginton, Coventry

Dieses Gerät wurde für den Vibratorverstärkerantrieb entwickelt. Es hat einen Frequenzbereich von 10 Hz bis 32 kHz und kann ±40 *g* Spitzenbeschleunigung genau regeln. Der automatische Frequenzdurchlauf arbeitet mit sieben Geschwindigkeiten für jeden Frequenzbereich. Der Durchlauf kann umgekehrt, an jeder gewünschten Stelle angehalten, und innerhalb des benutzten Bereiches beliebig begrenzt werden. Druckknopf-Handbedienung kann jederzeit den automatischen Durchlauf übersteuern. Signallampen zeigen die verschiedenen Arbeitsweisen an. Ein Rechteckausgangssignal kann ein Stroboskop-Flutlicht mit Oszillatorfrequenz oder mit einem kleinen Inkrement von 0,5 Hz, 1 Hz oder 2 Hz treiben. Bei mechanischer oder elektrischer Resonanz nimmt der Vibrationsreglerteil des Gerätes automatisch die Spitzenbeschleunigungsreglung des Tisches durch Einreglung des Vibrationsverstärkereinganges vor. Sind statt konstanter Beschleunigung andere Funktionen vorgeschrieben, so wird zwischen dem Wandler und dem Gleichheitsprüfer ein Funktionsverstärker in die Reglerschleife eingeschaltet.

Der Regler selbst ist fähig, innerhalb des 10 Hz—32 kHz Bereiches einzuregeln. Die obere Frequenzgrenze in diesem Bereich wird jedoch durch die Vibrationsanlage bestimmt. Er kann bis zu ±40 *g* regeln; aber bei niedrigen Frequenzen ist der Tischausschlag bestimmend.

EE 9762 *für weitere Einzelheiten*

NIEDERSPANNIGE UNIVERSAL-UMSPANNER

(*Abbildung Seite 307*)

Radiospares Ltd, 4-8 Maple Street, London, W.1

Diese niederspannigen Universal-Umspanner sind für abnormale Spannungen in Laboratorien und Fabriken konstruiert. Sie sind vollständig gekapselt und mit Harz ausgegossen; alle Verbindungen gehen zu Kabelschuhklemmen auf einem hochgültigen Hartpapier-Schaltbrett. Der

Umspanner hat sieben Sekundärwicklungen, und die Primärwicklung hat fünf Anzapfungen. Bei entsprechender Serien- und Parallelschaltung der verschiedenen Sekundärwicklungen können alle Spannungen von 1 V bis 40 V in 1 V -Stufen abgenommen werden. Die Gesamtbelastung des Umspanners darf 90 W nicht übersteigen.

EE 9763 *für weitere Einzelheiten*

AUTOMATISCHER ZÄHLER

(Abbildung Seite 307)

Società Elettronica Lombarda, 70 via Teodosio, Milan

Das Gerät besteht aus einem Schnelluntersetzer, einem automatischen Probenwechsler einfacher, stabiler Konstruktion und einem Zeitschreiber.

Der Untersetzer, in Bausteinkonstruktion ausgeführt, besteht aus einem Netzteil (300-2 000 V, Stabilität ±0,05%, Rückholgenauigkeit ±0,1%), einem Eingangsverstärker (Verstärkungsgrad 100, Ansteigezeit unter 0,2 μsec), einem Schnelldiskriminator oder einem Einkanal-Analysator (5-100 V Schwelle und 1, 2, 4 oder 8 V Kanalbreite), drei elektronischen Dekaden mit 1 μsec Auflösungszeit und einem automatischen Zähler (vorgewählte Zeit 10 bis 9 999 sec in 1 sec Stufen; vorgewählte Zählung 1 000 bis 10^6).

Der patentierte automatische Probenwechsler ist mechanisch sehr einfach und stabil. Er kann auf 25 mm Tischen 30 Proben und auf 50 mm Tischen 12 Proben unterbringen.

Die folgenden drei Nachweiselemente können durch einfaches Auswechseln Verwendung finden: ein Zählrohr mit Endfenster, ein Zählrohr mit extra dünnem Fenster (0,9 mg/cm²) oder Proportionalzähler, oder ein Szintillationszähler mit Alpha, Beta- oder Gammaphosphor.

Für die Zählrohre wird das billige Argon-Butan Gasgemisch benutzt.

Die Bausteinkonstruktion macht nicht nur die Wartung sehr leicht, sondern erlaubt auch spätere Erweiterungen und Änderungen, ohne dass das Gerät veraltet.

EE 9764 *für weitere Einzelheiten*

PROPORTIONAL TEMPERATURREGLER

(Abbildung Seite 307)

Ether Ltd, Caxton Way, Stevenage, Hertfordshire

Mit Einführung des fortlaufend gesteuerten, proportional wirkenden Reglers mit Rückstellvorrichtung Typennummer 995, hat die Ether Ltd ihre Serie der Transitrol Temperaturregler auf sechs erweitert.

Transitrol Type 995 ist mit einer 178 mm Anzeigeskala ausgerüstet, arbeitet im Zusammenhang mit einem motorgetriebenen Mischregler und regelt Gas- und Ölfeuerungen mit grösster Genauigkeit. Er kann auch eingesetzt werden, um Dampfdruck trotz Wechselbelastung gleichförmig zu halten.

Zur Temperaturkontrolle wird in der Messeinrichtung ein normales Galvanometer verwandt. Eine Zeigernase betätigt eine einfache fotoelektrische Vorrichtung, die ihrerseits den Motor für die Reglereinstellung kontrolliert. Verwandt wird hier ein für 24 V 50 Hz gewickelter, selbst anlaufender Umkehrmotor mit zwei Statoren und zwei Läufern, die koaxial auf einem gemeinsamen Schaft sitzen und die Leistung durch geräuschlose genau gefertigte Getriebe auf den endgültigen Schaft übertragen.

Das Gerät eignet sich für Wand- und Schalttafelmontage. Es ist als Einschiebeeinheit ausgeführt, um leicht aus dem Gehäuse herausgezogen werden zu können.

EE 9765 *für weitere Einzelheiten*

Zusammenfassung der Wichtigsten Beiträge

Anwendung gedruckter Schaltungen für Breitband-Mikrowellenbrücken von R. Rowland

Zusammenfassung des Beitrages auf Seite 256-261

Der Beitrag beschreibt die Verwendung gedruckter Schaltungen für die Konstruktion der Trägerfrequenz- und ZF-Geräte für Breitband-Mikrowellenbrücken und bespricht die dadurch gewonnenen Vorteile. Es wird gezeigt, dass die Bausteine billiger herzustellen sind und trotzdem erhöhte Leistung, Stabilität und Gleichförmigkeit aufweisen. Ausserdem ist die Wartung vereinfacht und der Prüfgerätbedarf wesentlich verringert. Anwendungsbeispiele für Bausteine mit gedruckten Schaltungen werden beschrieben.

Lawinen-Transistoren: Eine Auswertung ihrer Eigenschaften und Anwendung von R. C. V. Macario

Zusammenfassung des Beitrages auf Seite 262-267

Die Untersuchung der Eigenschaften einiger versuchsmässig hergestellter Legier-Transistoren ergab Resultate, die einen Überblick über ihre Kennzeichen ermöglichten. Die Kennzeichen sind einwandfrei den Eigenschaften der Flächenhalbleiter verwandt, was für die Entwicklung eines Elementes mit gegebenen Eigenschaften sehr vorteilhaft ist. Konstruktionsdaten werden besprochen und Anwendungsmöglichkeiten zusammengefasst.

Ein Kardiotachometer-Schnellschreiber von A. W. Melville und J. B. Cornwall

Zusammenfassung des Beitrages auf Seite 268-271

Das Gerät wurde für Einsatz im Operationssaal entwickelt. Es besitzt eine ausgezeichnete Trennschärfe gegenüber Störungen, eine wirksame Kontrollschaltung für die Impulsamplitude und einen Schreiber mit wahlweisen Zeitkonstanten. Eine der letzteren ermöglicht Schnellaufzeichnungen, die für die Bewertung schnell wirkender Anregungsmittel erforderlich sind.

Transistoren und Kerne in Zählschaltungen von F. Rozner und P. Pengelly

Zusammenfassung des Beitrages auf Seite 272-274

Die Verwendung magnetischer Kerne mit rechteckiger Hystereseschleife und Transistoren als Bauelemente in Digital-Rechnerschaltungen hat nun allgemeine Anerkennung gefunden. Der Beitrag zeigt, wie diese beiden Bauelemente zusammen eingesetzt werden können, um nichtkritische, zuverlässige Zählschaltungen für Eingabefrequenzen bis zu 750 kHz zu bilden. Eine beschriebene Abänderung erlaubt die Verwendung der Schaltung als Umschaltzähler.

Die Speicherladung eines Flächentransistors Während des Abschaltens in der Verstärkerzone von R. S. C. Cobbold

Zusammenfassung des Beitrages auf Seite 275-277

Durch die Lösung der eindimensionalen Diffusionsgleichung für einen Flächentransistor Können Gleichungen für Emitter—und Kollektorströme abgeleitet werden. Diese Gleichungen zeigen eine ziemlich gute Übereinstimmung mit praktischen Ergebnissen.

Magnetische Kerne mit rechteckiger Hystereseschleife als Schaltelemente von J. F. Kaposi

Zusammenfassung des Beitrages auf Seite 278-283

Magnetische Kerne mit ungefähr rechteckiger Hystereseschleife können wegen ihrer zwei stabilen Remenanzpunkte als Schalt- oder Speicherelemente verwandt werden. Der Kurvenlauf der Schleife für einen Kern dieser Art hängt z.T. von den magnetischen Kerneigenschaften und z.T. von der schaltenden Wellenform ab. Die theoretischen Grundlagen für Kernschalten mit Spannungs- und Stromimpulsen werden in dem Beitrag behandelt, Versuchsresultate mit verschiedenen Kernen und einige Spezialanwendungen beschrieben.

Genaue Messungen der Transistor-Grenzfrequenz von Yasuo Tarui

Zusammenfassung des Beitrages auf Seite 284-287

Das beschriebene Gerät ermöglicht schnelle und genaue α—Stromverstärkungsmessungen an HF Transistoren zwischen 1 MHz und 20 MHz. Der absolute Wert und die Phase können mit einer Genauigkeit von 1,5% gemessen werden. Eine schnelle Bestimmung der α—Grenzfrequenz ist möglich.

Ein Miniatur Funksprech-Uberleitgerät von E. J. Allen

Zusammenfassung des Beitrages auf Seite 288-290

Das Miniatur Funksprech-Uberleitgerät ist trotz vergleichbarer Leistungsfähigkeit sehr viel kleiner als herkömmliche Geräte und erlaubt die gleichzeitige Führung von vier Gesprächen üblicher Qualität.

Ein elektrophysiologischer Verstärker für Studenten von P. E. K. Donaldson

Zusammenfassung des Beitrages auf Seite 290-291

Ein kompakter Einseitenband-Verstärker für die Aufzeichnung der Aktionsspannung ausgeschnittener Nerven und Muskeln ist beschrieben. Das Gerät wird vollkommen vom Netz gespeist. Es wird angenommen, dass diese Tatsache, zusammen mit dem einfachen Aufbau, das Gerät auch für andere biologische Laboratorien interessant machen wird.

Ein neuartiger Ringzähler von P. J. Westoby

Zusammenfassung des Beitrages auf Seite 295-298

Mehrstufige Ringzähler werden vielfach benötigt. Die üblichen Anordnungen sind aus den angeführten Gründen auf ungefähr 12 Stufen beschränkt. Der Beitrag beschreibt eine Anordnung, die die Benutzung einer unbegrenzten Stufenzahl ermöglicht, ohne dass die den früheren anhaftenden Nachteile auftreten. Die Arbeitsweise wird beschrieben und einfache, die Entwicklung beeinflussende Uberlegungen werden angedeutet.