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Commentary

IN July of this year a central organization to be known as the National Electronics Research Council, or in its more abbreviated form N.E.R.C., was formed as the result of somewhat lengthy discussions between the Government and the electronic industry of this country.

The general committee under the chairmanship of Earl Mountbatten, who himself put forward the idea as far back as November 1961, will consist of the Permanent Secretaries of the various Government Departments concerned with electronics, representatives of universities and technical colleges, together with a number of executives from the larger electronic manufacturers nominated by the Conference of the Electronic Industry.

Its broad principle, which should meet with general approval, is to initiate and co-ordinate electronics research in industry, universities and colleges of advanced technology, and the government research establishments to the ultimate benefit of all concerned.

Although not a research association as such, N.E.R.C. is proposing as one of its first tasks to make a thorough investigation of existing research programmes throughout the country in order to discover gaps in research, to find where additional effort is required and also to prevent unnecessary duplication of effort, in other words to ensure that the best use is being made of existing research and to expand and supplement it where needed.

A very large amount of research in electronics is carried out in this country by the various Government Research Departments, mainly for defence purposes, and much of this work comes under the heading of 'classified'.

Nevertheless with Government representatives on its council, N.E.R.C. hopes that new ideas arising from defence projects no longer secret can be released and handed over to industry for further development and exploitation. N.E.R.C. will then co-ordinate the efforts and endeavour to provide finance for those projects which are likely to be too expensive for individual companies or universities to undertake.

Another aim of N.E.R.C. is to carry out a closer study of university research in the electronic field. It is recognized that much of the work carried out in the universities is fundamental in nature and that it is not always suitable for immediate commercial exploitation, but here again closer co-operation between universities and industry, N.E.R.C. hopes, will be of benefit, and it proposes to encourage any steps to improve the sharing of knowledge and experience between universities, industry and the Government research laboratories.

It is generally acknowledged that the methods by which new work, new developments and techniques are disseminated are not as efficient as they might be and it frequently happens that a research worker having spent considerable time and effort, and money too, on what appeared to him to be a new problem is frustrated to find that the work has been already done elsewhere and the results published.

Thus the research worker starting out on a new project in electronics—and this applies equally well to all fields of research—should spend a considerable amount of time in finding out what has already been done although the time spent is not necessarily to his disadvantage for he may well acquire background knowledge in doing so.

One result of the attention the world is paying to electronics research is the vast amount of published information on the subject and it has been stated that in electronic research alone some 25 000 papers and documents were issued throughout the world during 1963.

There are already of course information sources such as the Science Abstracts published by the various learned societies and professional institutions but it would appear that these are not on a big enough or sufficiently world wide scale.

The Institution of Electrical Engineers has taken a step in this direction by the recent publication of 'Current Papers' which lists each month some 700 titles of papers and articles selected from about 150 journals throughout the world, translating the titles into English where necessary.

These papers and articles are concerned with electrical and electronic engineering so 'Current Papers' which admittedly is an experimental publication has a long way to go if it is intended to cover the ever increasing number of papers etc in electronics—25,000 last year alone as already stated.

In order to disseminate information on a proper scale N.E.R.C. is proposing to set up an organization for what it calls the retrieval of information.

The importance which some countries attach to this retrieval of information can be judged from the fact that in the United States alone some 3 500 personnel are engaged in this subject and the rather staggering figure of 30 000 has been given for the number of people in the U.S.S.R. doing the same task.

N.E.R.C. is but a few months old and at this stage is not more than a project itself. As it develops N.E.R.C. will have many things to do if it is to succeed.

One of the most important factors at the moment is where the money is coming from to finance its various activities. The only sources of income are from industry itself and from possible Government grants, so much will depend on the impact it makes on industry and Government.

Under its constitution N.E.R.C. is not to be a profit-making organization but if it is to operate fully there is no doubt that a large sum of money will be required—running perhaps into several million pounds each year.

Microelectronics in Equipment Design

(Part 1)

By S. S. Forte*, Ph.D., B.Sc., A.M.I.E.E., M.I.E.E.E.

The miniaturization of electronic equipment is developing at an unprecendented rate brought on by the evolution of the new solid-state technologies and by the emphasis on the need for increased reliability. It is proposed in Part (1) of this article to review the methods at present available to improve reliability and reduce size, following this with a description of an experimental digital computer which employs techniques applicable to this type of equipment. Part (2) will describe the stages involved in the development of linear circuits for an experimental microminiaturized airborne navigational aid equipment.

(Voir page 650 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 657)

IN considering the advantages and disadvantages of microminiaturization, the first question which arises is the fundamental one of whether it is really desirable at all. The answer to this, for reasons which will be given, is an unqualified ' yes '—but, paradoxically, not necessarily because of the need to make things smaller. It is the solid state technology in itself which brings the main rewards, to which the miniaturization aspect is a bonus and no more.

This is not to deny that, with the ever-increasing complexity of electronic equipment, practical economic benefits do, in many circumstances, accrue from a reduction in size. They manifestly do so, for instance in the case of large computers, in which field immediate dividends in terms of reduced delay times and increase in speed are reaped as a consequence of size reduction. Even more obvious is the importance of miniaturization in missile and space vehicle development when it is remembered that in present systems the ratio of vehicle weight to useful payload is over 1000:1.

One main influencing factor (but not the chief) in the trend towards increased microminiaturization is cost. Although costs are at present rather high it is being predicted that the actual cost of modules will fall at least to the level of their discrete component counterparts. Further savings are likely to be achieved in the reduction of drawing and assembly time; microminiaturization will also provide a major saving in operational cost by reducing spares holdings and maintenance time.

Reliability is the real key word in microelectronics, however. The more complex the electronic equipment becomes, the greater becomes the need for high reliability and easier maintainability, and it is a fact that properly engineered microminiaturized components do achieve a spectacular degree of reliability in comparison with conventional ones.

It is axiomatic with any electronic equipment, whether conventionally sized, miniaturized, sub-miniaturized or microminiaturized, that if the failure of any one component causes the entire system to fail, then the reliability of the system as a whole will equal the product of the respective reliabilities of each component. Thus, because miniaturizing techniques enable designers to introduce greater complexity, the smaller is the time interval which is to be expected between failures, assuming the component reliabilities to stay constant. The only way in which the mean time between failures (m.t.b.f.)* can be improved is to increase the reliability factor of each and every component.

Unfortunately, miniaturized (as distinct from microminiaturized) design involves several factors which can tend to reduce reliability, rather than improve it. The concentration of heat sources, the close proximity of components and wiring, and problems of assembly and mounting are inherent features of the miniaturizing process; each and every one of these operates toward the reduction of reliability and the best that can be done is to design the equipment to reduce the effects to a minimum. The logical outcome of this is microminiaturization, where the spectacular reliabilities achieved for components in this category more than compensate for the uncertainties which can be introduced in the course of miniaturization.

Until quite recently there was only one way of building the maximum degree of reliability into electronic equipment. This was by employing ' tested-in ' reliability procedures, which include the rigorous evaluation of every single component, inspection at every stage of construction and subjection of the completely assembled unit to every conceivable environmental condition that it might encounter during its operational life. This approach is still the only one possible for full-scale, miniaturized or subminiaturized components and units.

With the advent of microminiaturization an entirely new approach was opened up. This concept, termed ' built-in ' reliability, became possible through the use of microelectronic circuits which in themselves possess reliabilities which are several orders better than their conventional counterparts. Overall reliability is further improved by the simpler circuits, reduced number of connexions and reduced power involved.

To recapitulate briefly, the basic aims with any form of electronic equipment are to increase the system reliability and to decrease the cost of manufacture, installation and maintenance. Reduction in size and weight is of secondary interest in many applications, though important in certain fields.

The means by which microelectronic techniques aim to

* *The Marconi Co. Ltd.*

* *Failure rate of components is normally expressed as a percentage failing per 1 000 hours of operation at a certain level of confidence, determined by the number of samples available for reliability evaluation. This can be converted into m.t.b.f. by reducing the percentage to a decimal at 1 hour and taking the reciprocal.*

achieve these objectives are:

(1) By reducing the number and variety of manufacturing processes involved and by subjecting these to very close control.

(2) By using pure materials and employing means to restrict subsequent contamination.

(3) By manufacturing what is virtually a solid equipment of low mass in order to minimize the effects of vibration.

(4) By reducing the number of separate electrical connexions to an absolute minimum.

This article is concerned with the microelectronic approach to equipment design and not with specific details of microelectronic techniques, which have already been fully covered elsewhere. Some definitions of terms would perhaps be in order.

The microelectronic circuits under discussion are broadly divided into two categories:—

(a) *Thin Film Circuits*. This is the generic name for the class of microelectronic circuit in which all the passive components are in the form of very thin films deposited on an inert substrate, and in which the active components are discrete devices (usually in specially-designed packages) which are attached to the substrate by one of several available techniques.

(b) *Semiconductor Integrated ('Solid') Circuits*. This is one of the many names given to the class of circuits which consist of integrated components in a semiconductor substrate.

Considering each of these approaches in a little more detail, it is evident that the use of *thin film microcircuits* represents the lesser degree of departure from conventional circuit design, for, with the exception of inductors and large capacitors, all the passive components are deposited as discrete devices which can all be tested at some stage in the manufacturing process. This technique therefore has the advantage of permitting the direct translation of conventional circuit design into microcircuit form, and one in which design procedure will follow normal circuit design right through the breadboard stage. Once the component values and tolerances have been finalized, the thin film component sizes can be calculated, so that it only remains to lay out the circuits in a manner analogous to that used in the preparation of printed circuit boards as a first step in the production of the masks. Since the active elements have to be added to the circuit after fabrication, it is evident that thin film techniques are most applicable to circuits in which there is a considerable predominance of passive components over active components.

With regard to *semiconductor integrated solid circuits*, the ultimate aim in this direction is the completely functional or molecular circuit which would consist of a piece of semiconductor material with input and output terminals; a unit which, while having no separate identifiable component areas, would nevertheless perform the electronic function of its conventional-circuit counterpart. This goal has not yet been reached, but fully integrated circuits are presently being produced in which the complete circuit is formed on a single block of semiconductor material, but in which identifiable circuit components exist, albeit in a completely new form.

Three distinct lines of approach are possible for solid circuits, as under:

(a) The fully integrated circuit as described above, in which the circuit elements are all formed on a single chip of semiconductor material and subsequently interconnected by a matrix of evaporated conductors.

In the present state of the art with this type of circuit, the range of resistor values which can be obtained without compromising transistor design is very limited; the incorporation of capacitors involves a heavy penalty in semiconductor area; moreover, there is an upper frequency limitation caused by the relatively low isolation between 'components'. The risks of failure in manufacture due to the complex chain of operations is very high, as a consequence of which the yields are poor and the costs high in all but the simplest of circuits.

(b) The multi-chip solid circuit. This consists of resistor, capacitor and transistor elements, either manufactured separately or in standard groupings, each under the best appropriate conditions and subsequently interconnected within a single component package.

This technique provides better flexibility of design, enables last-minute changes to be made prior to encapsulation by the provision of redundant components, and enables individual components to be fabricated on materials selected for optimum characterization for the type of component concerned. It provides better isolation between components than can at present be achieved with a fully integrated circuit; hence it is more adapted to high frequency applications. It does, however, suffer from the disadvantage that a large number of connexions are required between the individual circuit elements in the package. Although these connexions use proven techniques such as thermo-compression bonds, a certain degradation in ultimate reliability is probable.

(b) The hybrid integrated circuit. This comprises active elements produced in a semiconductor slice, upon which resistors, capacitors and conductors are subsequently deposited by thin film techniques.

This method has several marked advantages, the most important one being the improved performance of the passive elements so constructed. The technique is of course equally applicable to a multi-chip configuration.

It is hoped that enough has been said to underline that the problems associated with the manufacture of solid circuits are quite different from those arising from the use of thin film techniques. In solid circuits, the distributed nature of the passive components, coupled with the poor tolerances normally achieved, the very high temperatures coefficient of the resistors and the voltage dependence of the distributed capacitance, all combine to complicate solid circuit design. Whereas thin-film techniques lend themselves to conditions in which the rate of passive to active components is high, solid circuits, for the reasons mentioned above, do not; their use is most justified where a reverse ratio obtains.

There is a further very important requirement in the manufacture of solid circuits, namely that at every stage in the design it is imperative that all the limiting factors be taken into account; it follows from this that the utmost liaison and co-operation is demanded between the circuit designer and the semiconductor manufacturer. Failure to effect this can only result in the manufacturer offering a 'black box' with only a functional specification—a circumstance which will never satisfy the equipment designer.

It is not proposed to enter here into the thin film versus solid circuit controversy; suffice it to say that, in the writer's opinion, both have a place in equipment design. The choice of technique will be influenced by various factors, including the nature of the circuits, the degree of reliability required and the total cost which the system can bear. In the hybrid solid circuit outlined earlier both techniques are amalgamated, and here one probably has the best of both worlds.

In the light of present knowledge and existing techniques the high costs involved in the preparation of the complete set of masks required to produce a fully integrated circuit, together with the low yield likely to be obtained with all but the simplest configuration, render these circuits prohibitively expensive except in cases where really large quantities of identical circuits are used. This at present virtually restricts the use of fully integrated circuits to the digital field, but this is not to say that a restriction will always apply, as research for improved techniques may yield more economic methods of production. In the present state of the art, however, under conditions in which a wide range of requirements exist (as in linear circuits), the thin film microcircuit or the multichip solid circuit provide, in many instances, the only practical approaches.

An example of the latter situation is seen in the normal range of radio communication equipment. In this, linear circuits are predominant and the demands cover a wide range. Frequencies used range from d.c. to the microwave bands, and for each application widely different bandwidth, linearity and power level specifications exist. In general, circuit tolerancing is considerably more critical. For these main reasons it is virtually impossible, when dealing with linear circuits, to achieve the degree of rationalization which has become possible in the digital field, and the many and varied requirements, even if they could be met by solid integrated circuits, could only be fulfilled at a prohibitive cost.

There are, however, applications in the telecommunications field where simple repetitive circuits are used in large quantities, and in which, therefore, solid integrated devices can compete. Automatic error correction for telegraph and data transmission and telephone channelling equipments are cases in point.

Outside the realm of telecommunications, an immediately obvious application is in large electronic computers. Computer design has as its ultimate aims a reduction in size (and consequent reduction in delay times and increase in operating speed), reduction in power consumption, increase in reliability, simplification of system design, greater ease of maintainability and (because of the number of logic elements involved) eventually a reduction in cost. All these factors make microelectronic techniques an attractive choice for computer work, and furthermore, because of the simple nature of the individual circuits and the numbers used, an application in which solid circuits, either in fully integrated or semi-integrated form, can figure.

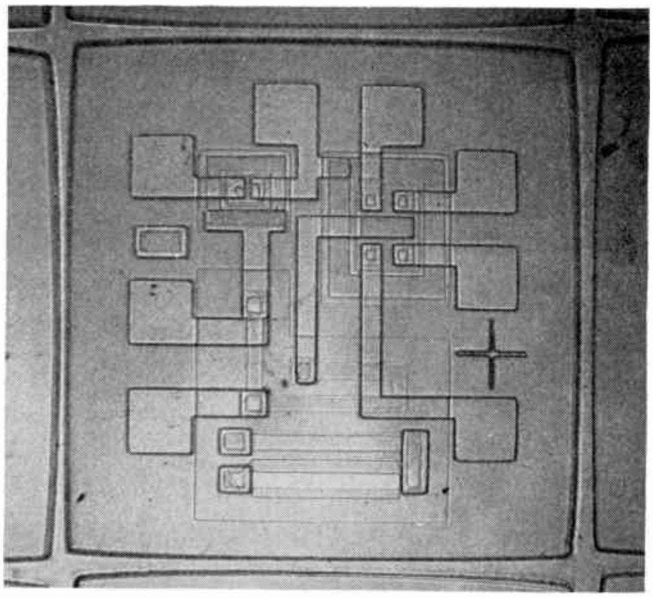
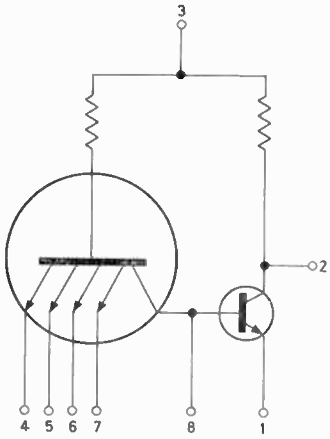


Fig. 1. Multiple emitter logic element

Fig. 2. Multiple emitter logic element equivalent circuit



Such then, is a brief resumé of the basic philosophy underlying the use of microelectronic circuits in equipment design. Two specific applications will now be considered as practical illustrations of a digital and a linear requirement, using semiconductor approaches.

Digital Computer

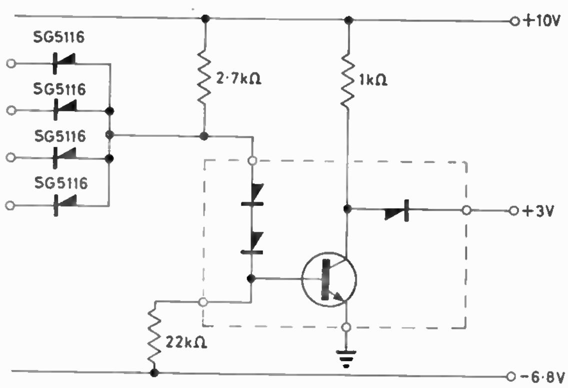
The computer described below has been designed in the Display and Data Handling Laboratories of the Company for digital data processing and 'real time' computation as required in conjunction with the latest radar techniques. The computer can be programmed to perform a wide range of calculations and is capable of very fast addition and multiplication.

The very high speed of operation made necessary by 'real time' computation has been achieved by combining short logic circuit delay with parallel operation of the arithmetic unit. The necessary degree of accuracy is provided by the 24 binary digit word length normally used.

The following examples of computation times demonstrate the speed of the machine.

Addition An addition order takes 2 to 5μsec to

Fig. 3. Semi-integrated diode transistor logic element



execute the operation involving two store cycles.

Multiplication The time for multiplication can vary between 9 and 12·5μsec (maximum) depending on the numbers involved.

In order to achieve these high speeds, it has been necessary to obtain a large degree of circuit compression. This has been achieved without sacrificing serviceability and accessibility by designing a new form of printed board assembly coupled with partial microminiaturization.

On considering various circuits for suitability as the general logic stage in the equipment, the main criterion is reliability. This must not only cover the life of individual components and joints, but must also include operational reliability in all possible system configurations. This has led to a considerable stress being placed on circuit noise immunity. From experience within the Company with a large system build-up using germanium logic, crosstalk voltages of up to 20 per cent of the logic circuit output swing are to be expected. This led to the increase of noise immunity on the fast logic circuit by including the voltage translating diodes in the base chain.

Turning to silicon logic, the crosstalk problem is aggravated since the silicon transistors, even the cheapest ones, will have f_T's in excess of 200Mc/s and will be in general 5 times faster than their germanium counterparts. The large increase in crosstalk which can be expected as a result of the faster transients can best be reduced by reducing path lengths in proportion. To this end the greatest volume compression at system level has been sought within practical engineering limitations. At present a 10:1 compression in volume has been achieved.

Since the logic circuit to be used in the computer must be cheap and compact, four of the most promising forms were chosen for evaluation. Three of these were in fully integrated solid circuit form, while the fourth is capable of full integration but was in fact considered in a semi-integrated form. Fig. 1 shows one of the elements of a slice carrying fully integrated logic components. This is a multiple emitter element and is illustrated in Fig. 2 as performing a NAND logic function with a 'fan in' of up to 4 and a 'fan out' of 3. It consists of a multiple gate followed by an invertor transistor.

The basic circuits investigated were Multiple Emitter

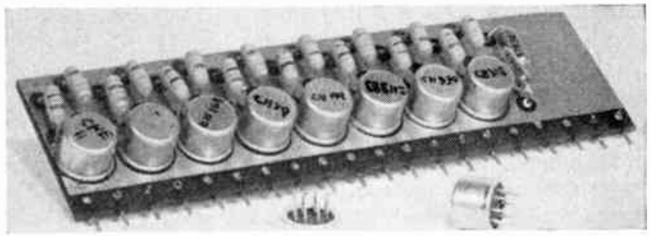


Fig. 5. Sub-board, consisting of 8 modules with external base chain resistors

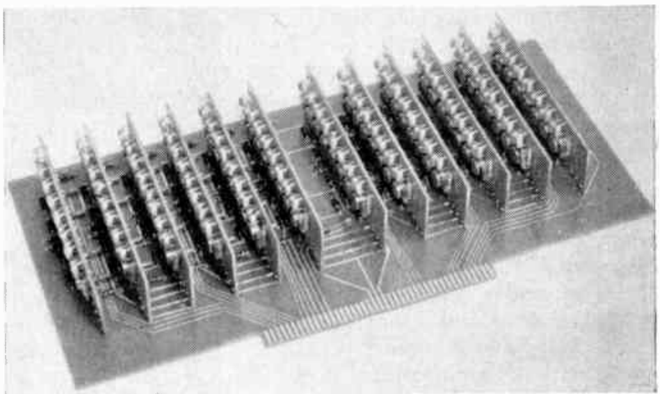
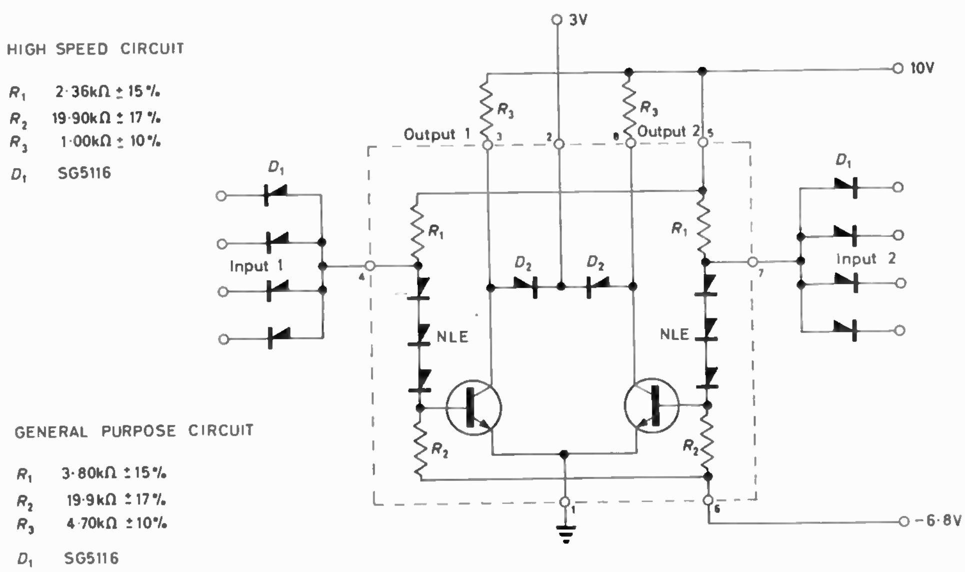


Fig. 6. Sub-boards mounted on a 'mother' board

Logic or Transistor Transistor Logic (m.e.l. or t.t.l.) (Figs. 1 and 2), Direct Coupled Transistor Logic (d.c.t.l.), Monolithic Emitter Coupled Logic (m.e.c.l.), and the semi-integrated Diode Transistor Logic (d.t.l.)

It was eventually decided that any circuit finally adopted for the general logic amplifier must have a noise immunity of at least one-third of the largest voltage swing commonly found in the build-up. This, together with other considerations, eliminated the first three logic systems, leaving the d.t.l. element as the form finally adopted. This noise immunity requirement is in the main due to the complexity of the build-up, and in general would not apply to small boxes of less than 9in cube. It is felt that, in the latter case, all the logic forms mentioned above would be satisfactory, provided that screening from external noise sources was adequate.

Fig. 4. Diode transistor logic double element



The d.t.l. circuit element used is shown in Fig. 3. The elements as used in this prototype machine are only partially microminiaturized since the resistors were left outside the T05 encapsulation, as were the input diodes. The exclusion of the input diodes and the collector loads was decided in order to achieve greater logical flexibility. The base chain resistors, on the other hand, were only placed externally as a matter of expediency at the time of initiation of the project, silicon diffused resistors of the desired tolerance not then being readily available, and, as is seen in Fig. 4, these have now been included within the T05 encapsulation in the later logic form. This, it will be noted, contains two logic stages except for the input diodes and the collector loads, which are still external.

Fig. 5 shows a sub-board consisting of 8 modules, with the external base chain resistors. The use of the new double stages with inclusion of base chain resistors will mean that the sub-board will only consist of four T05 headers. The boards use double sided printed wiring techniques with plated-through holes. The sub-boards are mounted on a 'mother' board as illustrated in Fig. 6 and these are plugged into the equipment. These 'mother' boards are interleaved to give a greater packing density.

The nearest equivalent to this computer performing the same a.t.c. and other radar functions is a computer which occupies four standard 7ft racks plus a control desk. As is seen in Fig. 7, the new computer is entirely contained within the control desk.

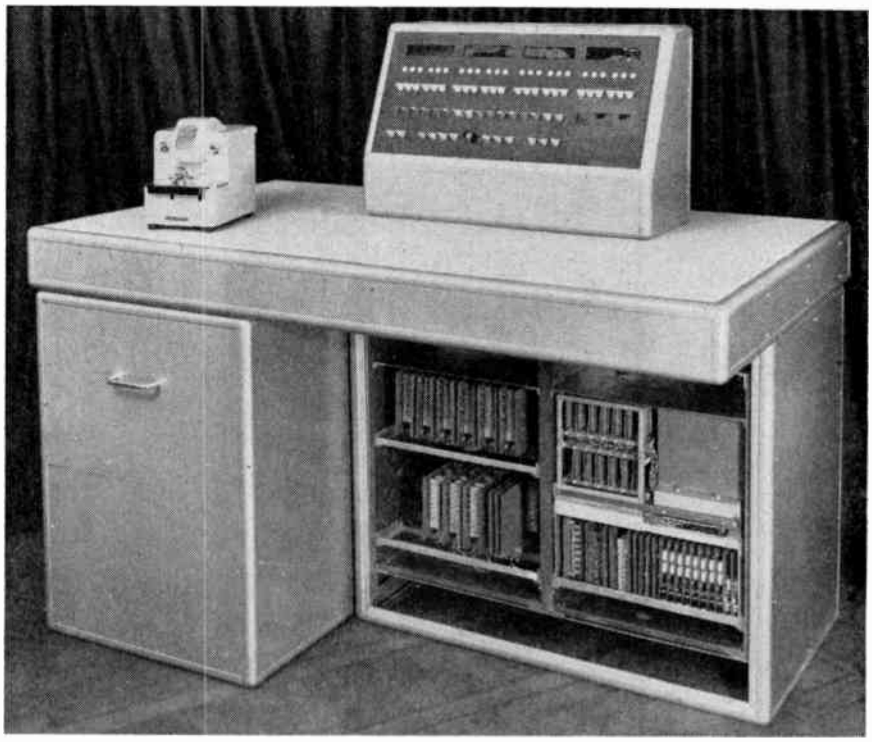


Fig. 7. Digital computer designed to incorporate microminiaturization techniques

The nearest equivalent using conventional techniques occupies four standard 7ft racks, plus a control desk. The new computer is entirely contained within the control desk and is 10 times faster than the conventional versions

The foregoing is an example of the operational use of one facet of solid circuit techniques, namely the multi-chip approach. In the equipment described, only partial micro-miniaturization was effected, but this was purely on grounds of expediency and no great technical barrier exists to the implementation of a wholly microminiaturized circuit. This has in fact been carried out in subsequent versions.

In Part (2) of this article it is proposed to outline some of the difficulties encountered in the application of micro-electronics to linear circuits, with particular reference to the design of an experimental microminiaturized 75Mc/s airborne marker receiver, which will be described in some detail.

(*To be continued*)

The Electronic Industry in Scotland

This month (from 3 to 19 September) the Scottish electronic industry will be putting on its biggest ever display at the *Enterprise Scotland '64* Exhibition in Glasgow. The industry's growth in recent years is one of the most striking success stories in Scotland since the war.

One of the major post-war objectives sought by the Scottish Council (Development and Industry) and also implicit in the Government proposals for the modernization of Central Scotland, has been the increase in opportunity for those trained in the sciences and technology. Research, and the industry which springs from research, holds the key to this need.

Thus two major schemes were initiated in the late 'forties. The first of these was the decision to set up the Government's main Engineering Research Establishment at the new town of East Kilbride. The second was the launching by Ferranti's, the Scottish Council and the Government, of an electronics scheme designed to train teams of engineers through Government contract work.

Since the days of the electronics scheme, the industry in Scotland has expanded rapidly. Over the last four years employment has grown by 64 per cent. This increase was nearly three times the proportionate increase in the same industry in other parts of Britain. It has taken output to an estimated £20 to £25M a year.

Part of the strength of the electronics industry in Scotland lies in the variety of companies which it contains. The only obvious absence is of companies making the simpler domestic appliances, such as radio and television receivers.

The largest and most sophisticated of all current projects is that of Ferranti's in the Edinburgh area. It employs around 7 000 people including hundreds of graduates engaged on research. Opened recently at Dalkeith, near Edinburgh, to manufacture electronic machine tool controls, it is the first plant devoted entirely to this kind of manufacture in Europe.

An American Company, Honeywell, is another of the big names in the Scottish electronic industry. This firm began their Scottish operations in 1948 in Lanarkshire with 60 workers. The initial line of manufacture was industrial control systems which have remained its main product. The Company recently announced its decision to introduce the assembly and subsequently the complete manufacture of a range of electronic computers. Over the next five years this line of manufacture will increase its labour force by an estimated 2 000.

At Glenrothes in Fife two other American electronic companies, Beckman and Hughes International, have both recently announced expansions. In another town, East Kilbride, Standard Telephones and Cables are going into electronic manufacture, while Birmingham Sound Reproducers are planning a Scottish project which might ultimately employ several thousand. Dobbie McInnes (Electronics) Ltd, of Glasgow, is another of the Scottish based electronic companies.

This growth and influx of new electronics manufacture in Scotland has already benefited from the standard grants made available by the Government in the Budget of 1963. The whole process should benefit further from the modernization scheme being put into operation throughout Central Scotland. Scotland's industrial belt is already undergoing rapid change—road and bridge building, new power stations, accelerated house-building and many measures designed to improve the amenities of the area.

The supply of trained people at university level is proportionately greater in Scotland than in other parts of Britain. An increasing number of technical colleges are catering for the demands created by electronics firms.

Automatic Switching of Sensitive Thermocouples

By J. L. Goldberg* and H. M. King*

An apparatus is described for the automatic switching of sensitive thermocouples at intervals continuously variable from a few seconds to several tens of minutes. The mechanical part consists of a telephone uniselector pawl and ratchet system geared to the shaft of a high quality thermocouple switch. An electronic controller periodically supplies a number of impulses to the uniselector coil in accordance with the gear ratio, so that correct indexing of the thermocouple switch is effected.

The apparatus has been developed for automatically recording the temperatures and temperature gradients along line standards and standard surveying tapes by using thermocouples.

(Voir page 650 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 657)

AUTOMATIC measurement of temperatures and temperature gradients with thermocouples is sometimes required in precision length metrology. Although high quality thermocouple switches are obtainable with residual e.m.f.'s less than $0{\cdot}01\mu V$, as far as the authors are aware no automatic switch is available which in addition operates continually at rates say from a contact every four seconds to a contact every twenty minutes.

The apparatus described here has been developed to fulfil this requirement.

Mechanical Construction

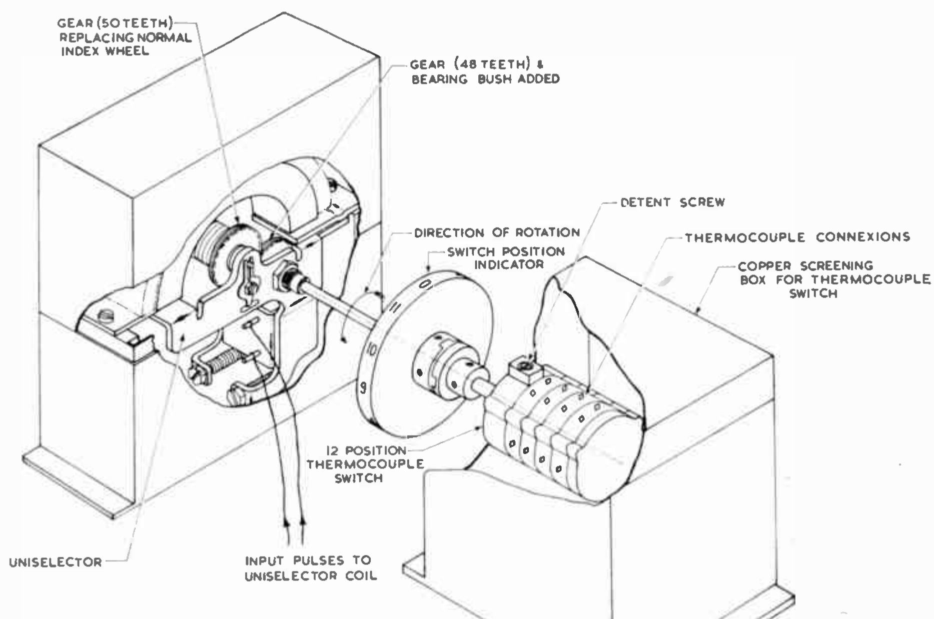
The well-known telephone uniselector automatic switch is unsatisfactory for switching very low-level d.c. voltages. This is due mainly to heat from the relay coil causing thermal e.m.f.'s in the switch[1]. The uniselector, however, does possess a well-constructed pawl and ratchet for driving the contact assembly and has a satisfactory torque characteristic. At average tensions in the restoring springs, the torque available at the mainshaft without the contact assembly is about 16oz/in, and this can be increased by a factor of almost two without causing failure of operation. If the contact assembly is retained, the excess torque available at the drive shaft is still approximately 12oz/in at normal spring tensions. Therefore, under suitable conditions, the uniselector driving mechanism can turn a switch shaft externally coupled to it.

For thermocouples a suitable automatic switch assembly is a 'Leeds and Northrup' twelve-position rotary switch coupled to the driving mechanism of a uniselector as shown in Fig. 1. This switch is accommodated in a copper box and a terminal block is provided for external connexions. This block is mounted on the lid of the copper box and provision is made for easy access to the detent adjusting screw of the switch.

The uniselector has a 50-tooth ratchet wheel driven by a pawl coupled to an electromagnet. Four impulses to the electromagnet rotate the ratchet wheel 4/50 of a revolution (28·8°). As the thermocouple switch shaft must be rotated through 30° to change contacts, a 50-tooth gear is mounted on the uniselector shaft remote from the ratchet. This gear meshes with a 48-tooth gear on the drive shaft to the thermocouple switch, and the drive shaft itself is mounted in a bronze bearing bush fitted to the uniselector frame.

The torque at the drive shaft of approximately 12oz/in is sufficient to turn the thermocouple switch after adjustment to its detent screw. In this regard, however, as the mechanical positioning of the switch shaft by means of the

Fig. 1. Mechanical arrangement of uniselector and thermocouple switch



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selector gearing is assured by the correct number of impulses to the magnet, the detent device on the switch is redundant and is therefore adjusted to provide minimum opposition to the driving torque.

The Electronic Timing and Control Circuits

The control circuits are designed to deliver a group of four pulses to the uniselector coil at about intervals of 0·5sec in order to step the main thermocouple switch one contact. The dwell time of the switch then commences. Means are provided to set this dwell time smoothly in the range 4sec to 22min. There is provision for manual starting of a timing period and for manual selection of a particular contact.

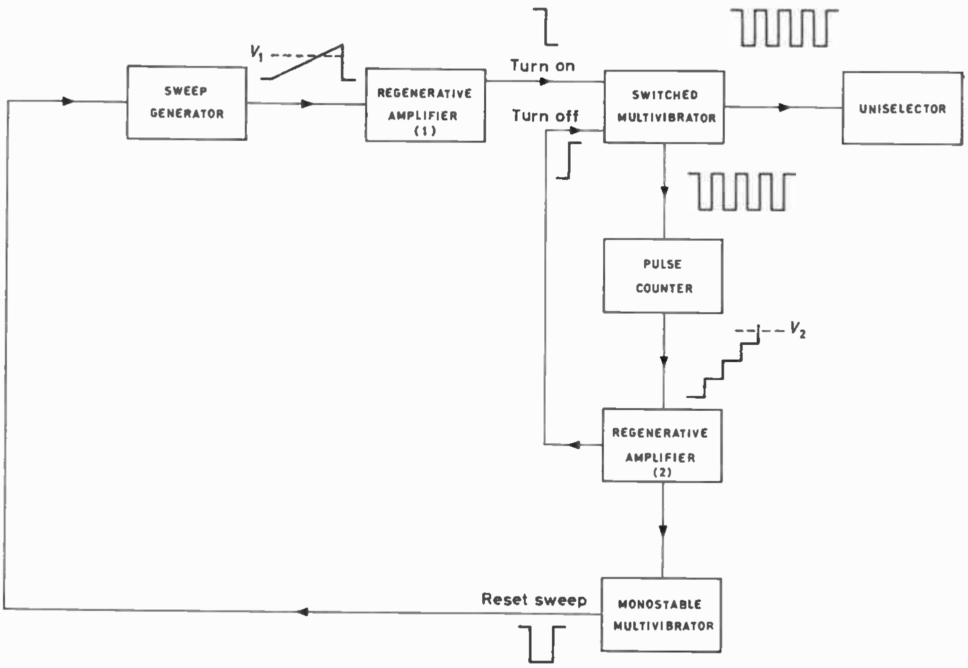


Fig. 2. Arrangement of switch controller

Description of Operation of the Electronic Controller

Fig. 2 shows an outline diagram of the system. A sweep circuit generates a substantially linear voltage waveform. At a certain voltage level V_1, which may be varied, a regenerative amplifier is triggered to give a sharply defined step of voltage. This allows an astable multivibrator to start and pulses are delivered from it to the uniselector coil. At the same time these pulses are counted in a capacitive counter and the waveform of the counter is made to trigger a second regenerative amplifier at a selected voltage, V_2. This voltage is proportional to the number of pulses required from the astable multivibrator. The output of this second regenerative amplifier performs two functions; first it stops the astable multivibrator, and secondly it generates a gating waveform which is used to reset the sweep circuit and thereby restart the operating cycle. The regenerative amplifiers are then automatically returned to their original states.

The astable multivibrator circuit is started and then stopped only after four pulses have been generated, and any variation in the multivibrator period has no effect on the number of pulses delivered. If an independent timing circuit was used for controlling the 'on' period of the multivibrator, the problem of ensuring stability of timing of two independent circuits would arise. This approach was found to be unsatisfactory.

Detailed Description of the Electronic Circuits (refer to Figs. 3, 4, and 5)

Sweep Generator

To obtain a linear sweep of long duration with a relatively low voltage supply, an electrometer valve V_1 is used as a cathode-follower in a Bootstrap circuit[2]. This type of circuit has an additional advantage in that the sweep voltage is produced at sufficiently low impedance to drive transistor circuits.

When relay contacts A_1 are closed at the start of a sweep, the d.c. conditions are arranged in the electrometer circuit in Fig. 3 so that the 'cathode' (actually the filament) potential is near zero volts. When these relay contacts open, C_1 charges via R_1, and if the gain of the electrometer circuit is nearly unity, the charging current to the capacitor C_1 will be substantially constant. If E is the battery voltage, C_1 charges at a rate of approximately E/R_1C_1 V/sec. The charging rate of C_1 is varied by switching in one of four resistors selected for the particular sweep rate required. Capacitor C_1 is a high quality oil-filled paper capacitor and the resistors R_1 are high-stability carbon.

The voltage waveform (a) shown in Fig. 5 is taken at the output of an emitter-follower VT_1 shown in Fig. 3. An npn transistor is used here in preference to a pnp transistor so that it can supply adequate current fo an external circuit when the voltage input to the base goes positive by a large amount with respect to earth.

Methods of Defining on-off Voltage Levels in the Transistor Circuits

Various methods are used to obtain well-defined 'on' and 'off' states independent of transistor characteristics.

In Figs. 4(a) and (b) the 'on' condition is obtained by feeding sufficient current to the base of the transistor to cause saturation. The 'off' condition is defined by supplying an input voltage of the correct sign so that the base-emitter diode of the transistor is cut off. When this occurs, the collector potential rises to the supply voltage in Fig. 4(a), or to clamping voltage in Fig. 4(b).

Where only a limited amount of voltage is available to supply the required turn-on current, the on-off states can be defined with the aid of a diode MR_1 and a base current E_s/R as in Fig. 4(c).

It is desirable to have non-saturating circuits with well-defined states. The 'on' state can be arranged by supplying sufficient input current V_1/R as in Fig. 4(d) and using current feedback through a diode MR_2 (Goldberg[3]). To achieve the 'off' state, the input current must be sufficiently reduced to ensure clamping of the collector voltage to a more negative level $-E_0$ volts, with the feedback diode MR_2 then non-conducting. Alternatively, a non-saturating stage can be turned off by applying voltages V_2, V_3, to the diode gating circuit shown in the dotted square of Fig. 4(d).

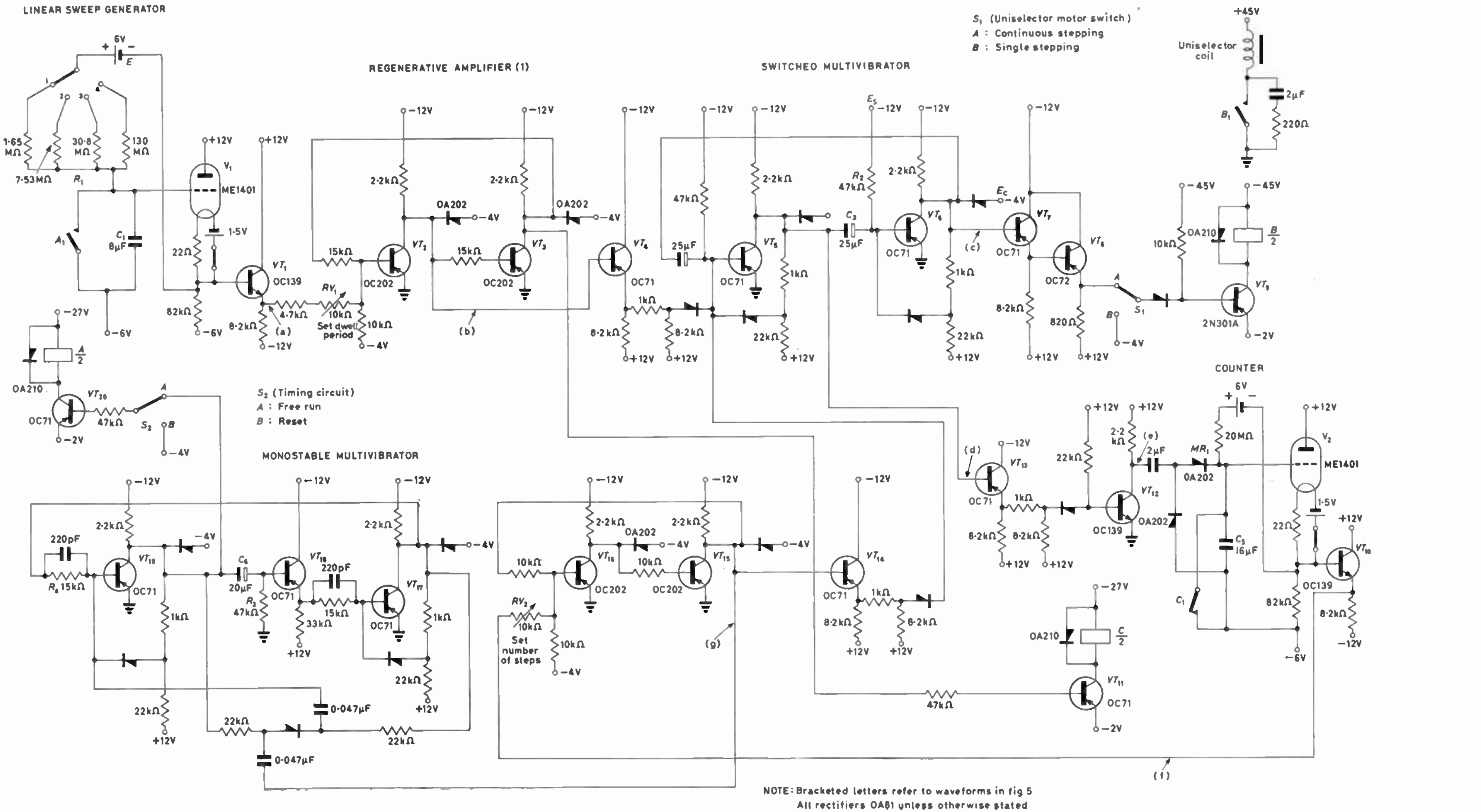


Fig. 3. Circuit of controller

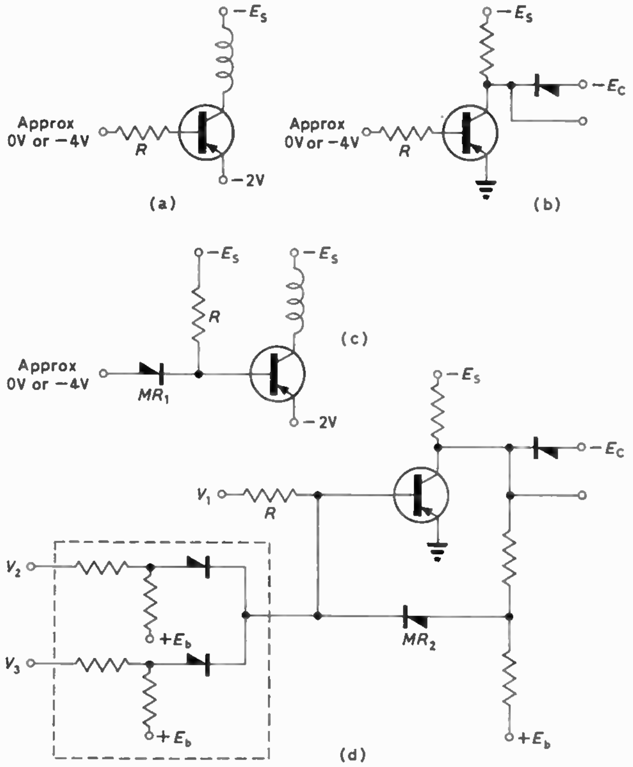


Fig. 4. Illustration of methods used to obtain defined "on" and "off" states in the transistor circuits

Regenerative Amplifiers

The regenerative amplifiers in Fig. 3 are formed by cross-connecting in a regenerative manner two circuits of the type shown in Fig. 4(b). A trigger action is then obtained. Sufficient current is supplied to one side of the circuit, i.e. to transistor VT_2, initially to bias this stage in the 'on' state. At a certain value of current input, defined by the voltage from the sweep generator and the series resistor RV_1 (marked 'set dwell period' in Fig. 3), the circuit reverses its state. The current at which this reversal of state occurs is a constant value for a given transistor, so that alteration of the input resistor RV_1 may be used to alter accurately the time required for the voltage applied to it to reach the trigger current level.

Table 1 shows the variation in dwell period which is obtained with the resistors R_1, by altering RV_1 (maximum value 10kΩ) alone.

Switched Multivibrator (VT_5 and VT_6)

Two non-saturating circuits are regeneratively coupled through reactive elements to form an astable multivibrator. An approximate expression for the period $2t_o$, sufficiently accurate at low frequencies, is:

$$2t_o = 2RV_2C_3 \ln\left[\frac{|E_s|}{|E_s| - |E_c|}\right]$$

If $RV_2 = 47\text{k}\Omega$, $C_3 = 25\mu\text{F}$, $E_s = -12\text{V}$, $E_c = -4\text{V}$, then $2t_o = 0{\cdot}96\text{sec}$ which agrees well with the experimental value of one second. Parameters E_s and RV_2 are chosen to ensure

TABLE 1

Range	Value of R_1 (MΩ)	Time Interval (sec)
1	1·65	3·8 to 16
2	7·53	16 to 71
3	30·80	67 to 304
4	130·00	298 to 1480

sufficient base current to each stage to establish a defined 'on' state[3].

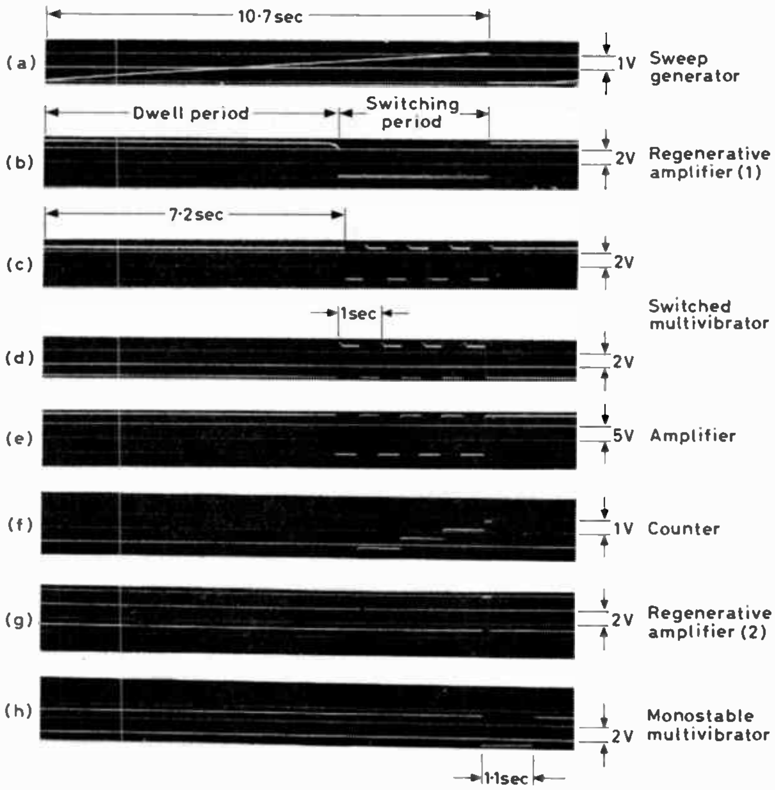
The multivibrator is normally held 'off' by the potential of VT_2 collector operating through the emitter-follower VT_4. The change of state in the regenerative amplifier (1) when triggered causes the multivibrator to start and oscillate continuously. This method of starting the multivibrator has been found to be always reliable when non-saturating circuits are used for the basic elements.

In Fig. 5, waveforms are shown of the regenerative amplifier (a) and the astable multivibrator. Triggering of the regenerative amplifier in waveform (b) is shown to occur here after 7·2sec, the dwell period of the controller. The switching period of 3·5sec then commences, the multivibrator delivering four pulses during this period (refer to waveforms (c) and (d).

Uniselector Switching Circuits

The collector voltage waveform of VT_6 is taken via a double emitter-follower VT_7, VT_8, and this waveform switches VT_9 in and out of conduction. Relay B operates and releases, switching current on and off in the uniselector coil. To handle the rapidly changing currents of considerable magnitude in the uniselector coil, and the consequent high induced voltages, relay B has Elkonite contacts of large diameter set in substantial leaf springs. This relay requires a power transistor VT_9 to operate it satisfactorily.

Fig. 5. Waveforms of the controller during one cycle of operation



The collector voltage of VT_5, the other phase of the multivibrator, is used to monitor the number of switching cycles that have occurred. This monitoring waveform is amplified in an npn transistor VT_{12}. The collector waveform of VT_{12} is thus in the same phase but lies between 0V and +12V, i.e. the voltage is always positive with respect to earth. Refer to Fig. 5(e).

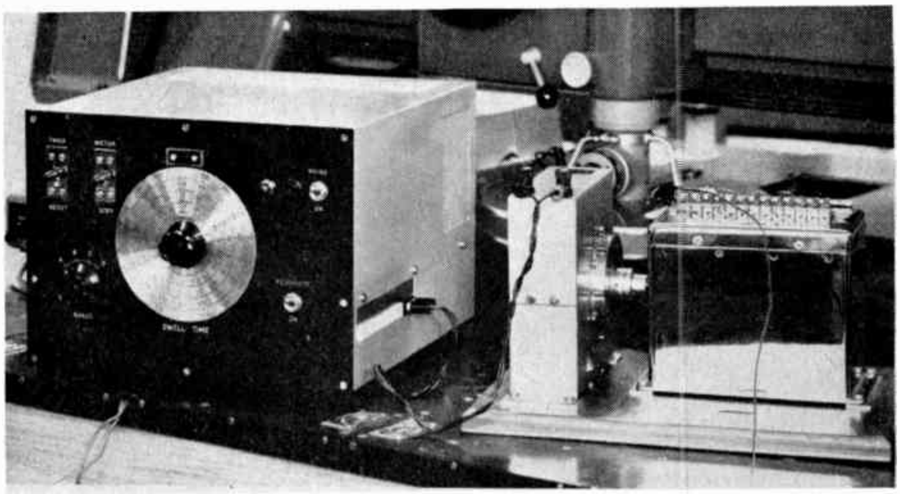


Fig. 6. A production model of the apparatus

COUNTER CIRCUIT

This circuit, which consists of capacitors C_4 and C_5 and diodes MR_1, MR_2, is used to generate a staircase waveform and thus indicate the number of pulses applied to it. When the regenerative amplifier (1) changes state, it causes relay C to release by cutting off transistor VT_{11}. The normally closed contacts, which ensure that the starting voltage on C_5 is zero, now open and the voltage on C_5 is thus free to change.

When the input waveform is negative-going, C_4 discharges through MR_2, and MR_1 is cut off. During positive-going excursions, MR_1 conducts and C_4 receives the same charge as C_5 although the voltage on each is different, C_4 being of smaller capacitance than C_5. The voltage is built up on C_5 in discrete steps on each successive pulse. The magnitude of each successive step, however, becomes smaller as the number of pulses increases because the potential at MR_1 anode must reach the potential on C_5 before charge transfer can take place. It is important in practice to ensure that the pulse amplitude supplied to the circuit is stable and that capacitors C_4 and C_5 are stable components because their value determines the magnitude of the steps of voltage. In this apparatus C_4 and C_5 are oil-filled paper capacitors; the +12V supply of VT_{12} comes from a Zener diode. A clear description of this type of counter circuit is given by Earnshaw[4].

The staircase voltage waveform which is shown in Fig. 5(f) is transferred to a second regenerative amplifier VT_{15}, VT_{16}, by means of an electrometer cathode-follower V_2 and emitter-follower VT_{10}. With the contacts of relay C closed the output voltage level of VT_{10} is near earth potential. When the waveform at the emitter of VT_{10} is undergoing transition from the third to the fourth step, the regenerative amplifier VT_{15}, VT_{16} changes state.

The triggering voltage level is determined by resistor RV_2 which should be adjusted so that the level lies in the middle of the transition mentioned. There is then a tolerance available to take account of changes in the starting level of the staircase waveform. These changes may occur owing to run down of the electrometer filament batteries and will alter the grid-filament potential of V_2.

A step waveform taken from the collector of VT_{15} turns off the astable multivibrator by biasing the base-emitter diode of VT_5.

MONOSTABLE MULTIVIBRATOR

The voltage step from the regenerative amplifier (2) which turns off the astable multivibrator is also used for triggering a monostable circuit for resetting the sweep generator. This monostable circuit is formed by coupling two non-saturating circuits regeneratively. One of the couplings is direct, the other is through a circuit C_6, R_3 and an emitter-follower VT_{18}. With these connexions as shown, VT_{19} will be high conducting and VT_{17} will be low conducting. The positive-going step of voltage from the regenerative amplifier (2) is applied from VT_{15} collector via a gating network to make the circuit respond only to positive-going inputs. This gating network is essential because of the sequence of events which occurs soon after the monostable circuit has been triggered: relay A operates to reset the sweep circuit; this in turn causes regenerative amplifier (1) to reset. Relay C also operates and resets regenerative amplifier (2). The output of this latter circuit is negative-going and, in the absence of the gate, would re-trigger the circuit so as to shorten the period. (Refer to Figs. 5(g) and 5(h)). The period of the monostable circuit can be calculated approximately from the formula

$$t_0 = R_3 C_6 \ln \left[\frac{|E_c| \, . \, \alpha'}{I_c R_4} \right]$$

where E_c is the clamping voltage, and α' is the earthed emitter current gain at a collector current I_c. In this case

$$E_c = -4\text{V}, \ \alpha' = 50, \ I_c = 5\text{mA}$$

$$R_4 = 15\text{k}\Omega, \ R_3 = 47\text{k}\Omega, \ C_6 = 20\mu\text{F}$$

so that t_o = 1·01sec. The measured value was 1·1sec. The purpose of the emitter-follower VT_{18} is to isolate the timing network C_6 R_3 from the neighbouring transistor stage VT_{17}.

POWER SUPPLIES

The power supply is mains-operated and is conventional. Zener diodes are used to stabilize the various voltages with the exception of the supply for the uniselector.

Conclusion

The prototype apparatus has performed successfully in the Division for a period of six months and has proved satisfactory for mass production.

A photograph of the production model is shown in Fig. 6.

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A Simple Capacitance Measuring Circuit with Digital Presentation

By S. L. Hurst*

A simple, very rapid and foolproof method of measuring capacitance values in the range of from say 0·001μF to 10μF is described, with an accuracy of measurement that is within ±2 per cent. Within smaller ranges of values very much higher accuracy of measurement is possible. The method may also be extended to the measurement of resistor values, but with certain restrictions and possibly inferior accuracy. The matching of capacitors and resistors may also be made to a very high order of accuracy.

(Voir page 650 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 657)

IN very many circumstances, such as in the development laboratory, test departments and works receiving and acceptance departments, the need for a rapid measurement of capacitor values is encountered. In many of these cases, a very high accuracy of measurement is not required, and neither is the loss angle or power factor of the capacitor; instead a quick check to ascertain that capacitors are of the value marked upon them to within say ±5 per cent is often perfectly adequate.

With the increasing availability of commercial digital-type counter/timer instruments as a stock item of equipment in many laboratories and industrial organizations, the facility of utilizing this instrument for the direct presentation of capacitor values is available, with the addition of only a simple self-contained transistor monostable multivibrator circuit external to the timer. The only requirement necessary on the counter/timer is that the time measurement shall be capable of being started and stopped by pulses or d.c. voltage steps fed into it, this however being a standard facility on most commercial instruments except the simplest and cheapest ones.

Theory of the Test Circuit

The circuit of the simple transistor monostable circuit is given in Fig. 1. The collector supply voltage V_{CC} is a negative voltage if pnp transistors are in use, while V_{BX} is a bias supply of opposite polarity. The normal (stable) state of the circuit is VT_1 non-conducting ('off') and VT_2 fully conducting ('on').

If this circuit is pulsed to its unstable state of VT_1 'on'/VT_2 'off' by an appropriate input signal, the circuit relaxes to its stable state after a time dependant upon the time-constant C_1R_2 the text-book formula for this time, assuming negligible voltage-drops across a transistor when 'on', being $t = C_1R_2 \ \ln 2, = 0{\cdot}693C_1R_2$ microseconds, where C_1 is in microfarads and R_2 is in ohms.

This formula indicates that the timing is independent of supply voltage values and transistor parameters and this is true to a very large degree provided the supply voltage V_{CC} is large in comparison with the conducting transistor collector-to-emitter and base-to-emitter voltages V_{CE} and V_{BE}, and also providing the values of R_2 and R_3 are such as to allow VT_2 and VT_1 to be driven into the fully 'saturated' condition.

Thus if (a) C_1 is made the unknown capacitor, (b) R_2 is preset to an appropriate value, and (c) the counter/timer is arranged to start timing on triggering the monostable circuit to its unstable state and stop timing when the circuit relaxes 0·69 C_1R_2 microseconds later, then under these conditions the count registered on the timer can be read off directly as the value of C_1. The time taken to present this answer will of course be merely the relaxation time of the circuit, which for all normal values of capacitance will be a small fraction of a second.

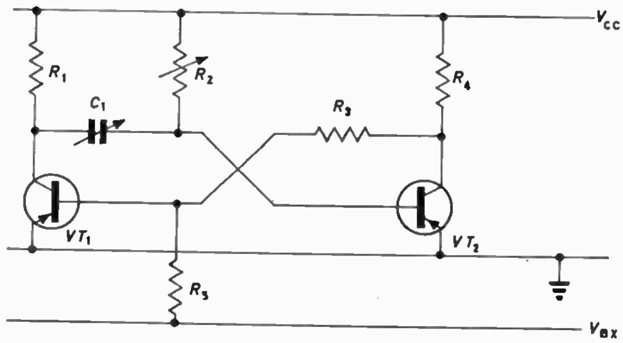

Fig. 1. The basic test circuit

Circuit Details

The monostable circuit used is similar to the basic circuit, but with the addition of a small 'speed-up' capacitor C_2 connected across R_3 in order to sharpen the transit time of the circuit. The timer start/stop signals are both taken from the collector of VT_2, the timer being set up to start on receipt of a negative wavefront and stop on receipt of a positive wavefront. Resistor R_2 consists of one fixed plus two variable resistors in series, the two variable ones being a 'coarse' and a 'fine' preset adjustment, with the fixed resistor as a safeguard for VT_2 in case both presets are inadvertently adjusted to zero ohms.

A collector supply voltage V_{CC} of −12V and a bias supply voltage V_{BX} of +6V were chosen; as will be detailed below variation of V_{CC} over a wide range can be accommodated with little variation in timing, while V_{BX} can if desired be dispensed with at normal room temperatures if R_5 is returned to the zero voltage line. The presence of V_{BX} however does ensure improved timing stability for the circuit under increased ambient temperature conditions, or if VT_1 tends to exhibit a high leakage current.

To trigger the circuit from its stable to its unstable state and hence initiate the timing count, two means are provided. The first is a manual 'one-shot' arrangement, whereby depression of a push-button produces one cycle only of the monostable circuit; the alternative is an 'auto' arrangement, whereby a slow-speed astable multivibrator repeatedly triggers the monostable circuit approximately once every second.

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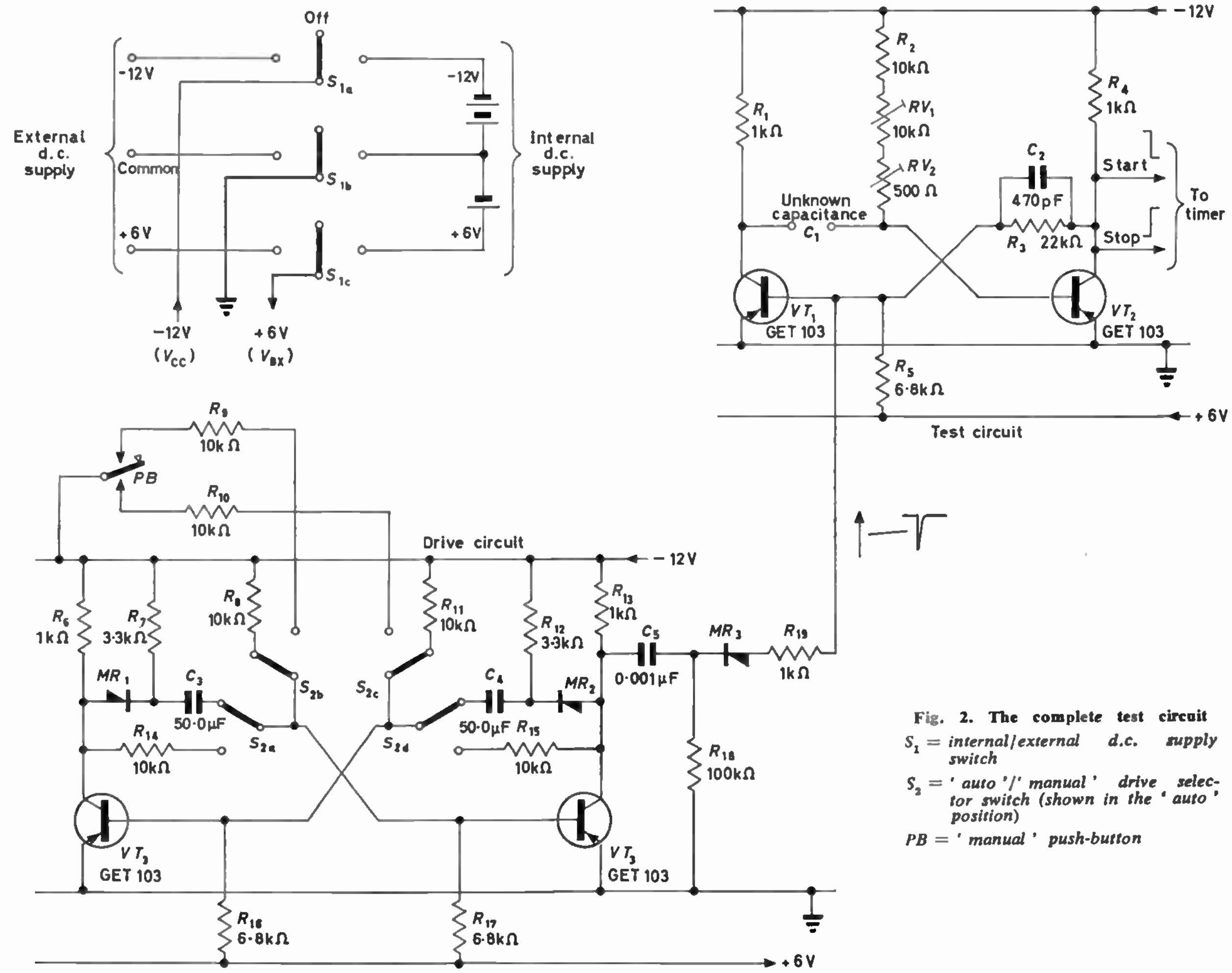

Fig. 2. The complete test circuit

S_1 = *internal/external d.c. supply switch*

S_2 = *' auto '/' manual ' drive selector switch (shown in the ' auto ' position)*

PB = *' manual ' push-button*

The astable multivibrator used in the 'auto' condition is a common-emitter collector-coupled circuit, but with diode isolation between the transistor collectors and the cross-coupling capacitors in order to give faster rise times to the collector waveforms. This circuit modification[1] allows positive and negative rise times of one or two microseconds to be achieved with even low-frequency pnp transistors in the 1c/s multivibrator, this feature being particularly important for this application as a well defined short-duration pulse to trigger the final monostable circuit is desirable.

In the manual condition care has to be taken that one depression of the manual push-button triggers the timing circuit once only, as erratic multiple triggering may produce cumulative or spurious answers on the timer instrument. A simple connexion from the push-button to trigger the monostable circuit is thus unsatisfactory due to the possibility of contact bounce on the push-button, and the most economical circuit arrangement to guard against this possibility is to re-connect the astable multivibrator driving circuit so as to form a bistable circuit, the push-button now 'setting' and 'resetting' this bistable circuit to suitably produce one and only one output pulse. The push-button should have a break-before-make action to ensure this condition is always absolutely maintained.

The complete test circuit, consisting of the monostable circuit containing the unknown capacitance, and the driving circuits is therefore as shown in Fig. 2.

In order that the resultant digital readings on the timer may be read off directly as the value of the unknown capacitance C_2, it is first necessary to preset R_2 accordingly. As the timing is theoretically given by $t = 0{\cdot}693\, C_1 R_2$ this therefore requires $0{\cdot}693\, R_2$ to be some factor of ten, i.e. R_2 must be $(1{\cdot}44 \times 10^x)$ ohms, where x is an integer.

Now the satisfactory operation of the monostable circuit imposes restrictions on the value of R_2, too low a value resulting in excessive base current for VT_2, too high a value resulting in inadequate base current to switch VT_2 fully 'on'. With the collector loads and type of transistors employed, R_2 is restricted to the theoretical value $1{\cdot}44 \times 10^5\Omega$ for optimum conditions.

In practice, the measured total value for R_2 set up in the Operating Details covered below is found to be about $14{\cdot}25\text{k}\Omega$, thus indicating close agreement between theoretical and actual results.

Operating Details

A capacitor of known value is first connected to the test circuit, and the preset coarse and fine resistors R_2 adjusted such that the digital reading on the timer numerically equals this value of capacitance. As will be seen in the Accuracy of Measurement section below, the accuracy of measurement of the circuit falls off above and below the capacitance at which the test circuit was preset; it is thus advantageous to use a capacitance for setting of about the mean of the values likely to be measured on the unknown capacitors.

The settings on the timer itself should be adjusted as necessary, and the significance of the decimal point (if any) on the display noted. Exact details will of course vary depending upon the make and type of counter employed. However, if the counter is set for example to directly display a count of microseconds then with R_2 set to its optimum value of about 14kΩ, the microsecond count will be equal to microfarads × 10^{-5} e.g. a count of 10 000 on the timer = 1·0μF.

One fairly obvious precaution is necessary if the test circuit is used on 'auto' and that is the timer itself must not automatically count and reset such that it sums more

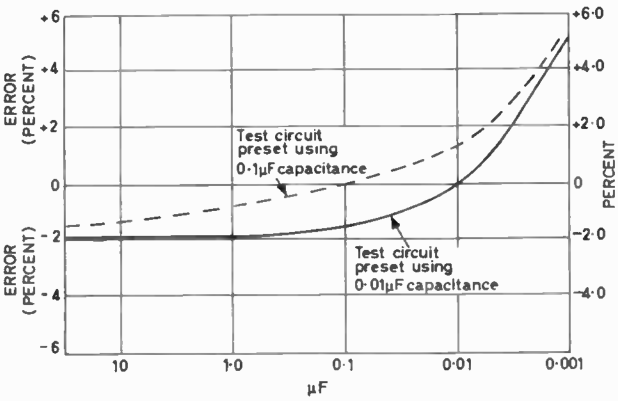

Fig. 3. Accuracy of resultant measurement against value of capacitance being measured

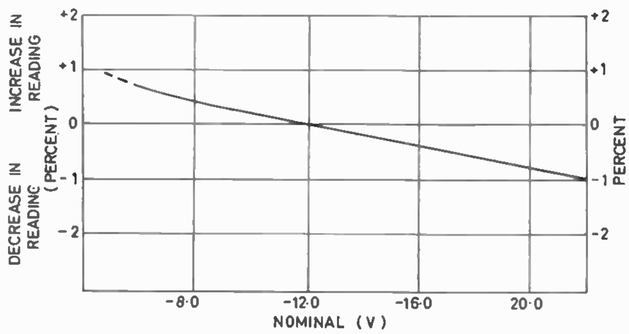

Fig. 4. Variation in measurement against variation of negative supply voltage V_{cc}

than one cycle of the monostable test circuit. As the test circuit on 'auto' is being driven only about once per second, it is quite straightforward to observe that this is not happening and that the timer starts from a zero count for each test cycle.

Accuracy of Measurement

With the nominal values of voltage applied to the circuit, namely $V_{CC} = -12·0$V and $V_{BX} = +6$V, and with the necessary preset adjustments made using a 0·1μF known capacitance, the accuracy of measurement over the unknown capacitance range from 0·001μF to 10μF, i.e. four decades, is as shown in Fig. 3.

As will be seen from this graph, an accuracy of measurement within ±2 per cent is obtained except for values below about 0·003μF, below which point the error begins to increase rapidly. The logarithmic scale of the horizontal axis tends of course to accentuate the fall-off in accuracy at the low-value end. Greater accuracy at these low values of capacitance could be obtained by minimizing the stray capacitances of the test circuit layout, or by setting up the preset adjustments using a 0·001μF capacitance instead of 0·01μF.

With the preset adjustments set up using 0·1μF, the accuracy of measurmeent graph is shifted generally upwards by one decade of capacitance value, as shown dotted in Fig. 3. Again a fall-off in accuracy will be noted when the capacitance being measured approaches the 0·001μF region.

As will be apparent, where it is required to pair or match capacitors against another capacitor, this test circuit enables such matching of values to be readily undertaken with exceptional accuracy. A matching accuracy of, for example, two parts in a thousand would normally be available from a four-digit time display.

The effect of variation of d.c. supply voltage V_{CC} on the accuracy of measurement is shown in Fig. 4. As will be seen from this graph, a −50 per cent to +100 per cent variation in V_{CC} produces less than 1 per cent change in timing of the test circuit. Variation of the positive bias

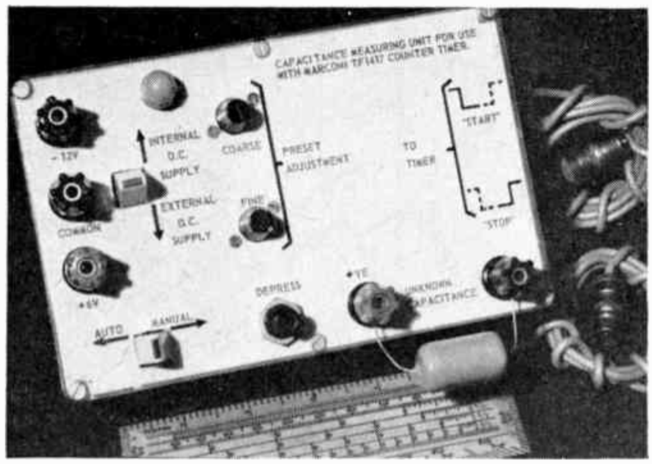

Fig. 5. Compactness of complete test circuit

supply V_{BX} has no measurable effect whatsoever on the time as would be expected, provided transistor VT_1 remains fully cut-off in the stable state of the monostable circuit. At an ambient temperature of 20°C with a healthy transistor in position VT_1, reduction of V_{BX} from +6·0V nominal to +1·0V can be readily accommodated.

Measurement of Resistor Values

The basic principle employed in the foregoing capacitance measurements can also be utilized for the measurement of values of resistance. In this case C_1 has to be preset and R_2 becomes the unknown component.

Two serious practical limitations are however involved. Firstly C_1, which requires to be set to approximately 0·0142μF in order that microseconds on the timer may be read off directly as ohms, is difficult to preset, and secondly permissible upper and lower limits must be laid down for R_2. Too low a value for R_2 will damage transistor VT_2, while too high a value will introduce inaccuracies in the timing, due to VT_2 not being held fully conducting in the stable state of the circuit. A practical range of resistor values for R_2 would possibly be 1kΩ to 50kΩ.

In general, except possibly for the matching of resistors in the 1kΩ to 50kΩ range, the use of the circuit for resistance measurements is inconvenient and limited.

Conclusions

The test circuit described enables simple and rapid measurement of capacitor values to be made with an

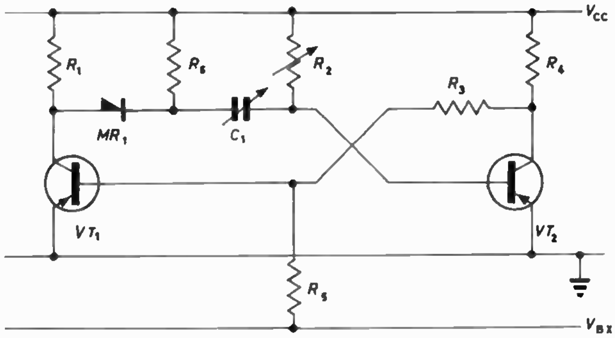

Fig. 6. Modification to basic monostable timing circuit to enable VT_1 collector waveform to be utilized

accuracy adequate for the majority of acceptance testing and production or circuit experimental use. It would seem particularly suitable for use by unskilled personnel, provided the preset adjustments and timer controls were set by appropriate staff.

Fig. 5 indicates the possible compactness of the complete test circuit, which can be made self-contained with its own batteries if desired, or supplied from some external d.c. supply.

Minor modifications to the test circuit to modify the polarity of the 'start' and 'stop' control signals fed to the timer may of course be necessary to suit specific types of timers that cannot accept the negative 'start' and positive 'stop' wavefronts used in the circuit detailed above. In particular, if an opposite polarity of either or both control signals is necessary, then the modifications shown, in Fig. 6, to the monostable timing circuit may be necessary, in order that the collector waveform of transistor VT_1 may be 'square up' and hence utilized for start or stop control purposes.

Acknowledgments

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The British National Hydrogen Bubble Chamber at C.E.R.N. Geneva

THE 150cm Bubble Chamber, a description of which was given in 1963* has recently been brought into operation at C.E.R.N. Geneva where it is working in conjunction with the 28GeV proton-synchrotron and the first series of experimental results have now been obtained.

It will be recalled that the bubble chamber consists essentially of the following parts:

(1) The chamber vessel itself, of forged aluminium alloy, fitted with two vertical glass windows.

(2) The expansion system for the chamber.

(3) Hydrogen shields, to contain liquid hydrogen if a leak or fracture of a window should occur.

(4) A liquid nitrogen shield to screen the chamber from extraneous heat.

(5) A vacuum tank with associated pumping equipment to enclose the parts above.

(6) A hydrogen liquefier.

(7) An electromagnet system.

(8) An optical and photographic system for recording the particle tracks in the bubble chamber.

(9) Overall control system for the expansion of the chamber, together with all the instrumentation.

The body of the chamber is shown in Fig. 1 and above it is part of the expansion system. The square frame surrounding the chamber is that of the vacuum jacket which is designed to contain 500 litres of liquid hydrogen at a temperature of −250°C.

The three circular holes at the side of the body are for the mounting of the light sources.

The proton-synchrotron which was officially inaugurated in February 1960 is some 200 metres in diameter, and is primarily a circular accelerator designed to give protons—the nuclei of hydrogen atoms—the very high kinetic energies required for present-day research into the fundamental structure of matter.

The protons are produced initially from hydrogen atoms by an ion source and are introduced to a pre-accelerator where they attain an acceleration up to 500keV. Further acceleration to the protons is given in successive steps by a linear accelerator up to 50MeV, corresponding to a speed of about one-third that of light.

At this point they are injected by an inflector into the circular vacuum chamber of the proton synchrotron in pulses of six microseconds duration.

The circular vacuum chamber of 200 metres diameter has a cross section of 14cm by 7cm and 16 accelerating cavities are located round the chamber.

As the beam of protons accelerate round the chamber a magnetic field, produced by 100 magnet units, increases in step so as to maintain the beam in its orbit.

Pick up electrodes determine the position of the beam relative to the side walls of the chamber and are connected to a feedback system to control the orbit.

The duration of the acceleration cycle is about one second during which time the beam has travelled round the vacuum chamber of the synchrotron some 480 000 times and has attained an energy of 28GeV.

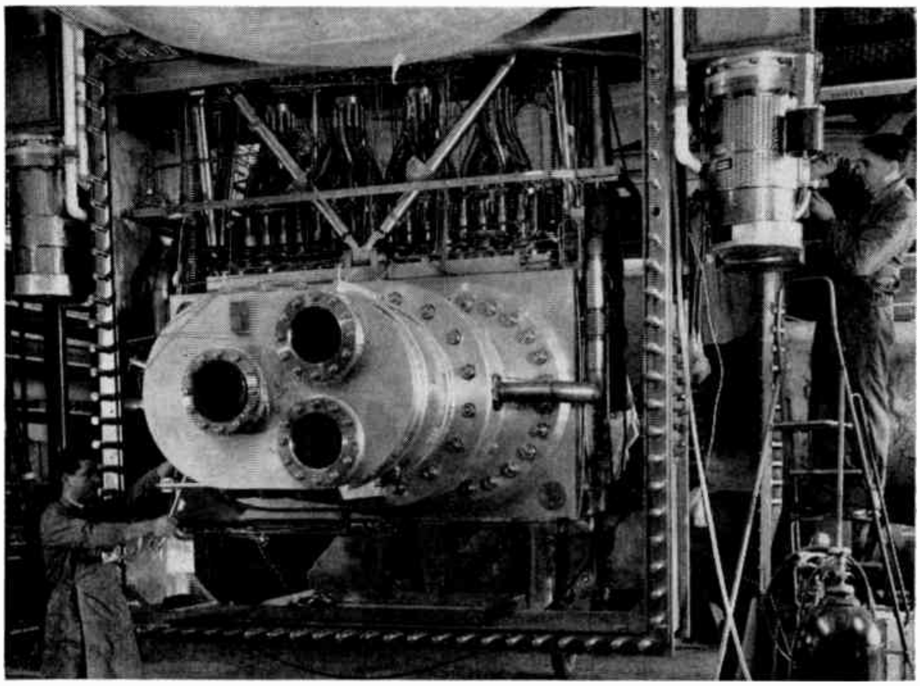
The beam is then directed along a 180 metre path from the proton-synchrotron to the bubble chamber by a number of horizontal and vertical bending magnets and remote-controlled adjustable collimators maintain the beam dimensions. This 'beam-line' was specially designed to produce more kaons of a higher energy.

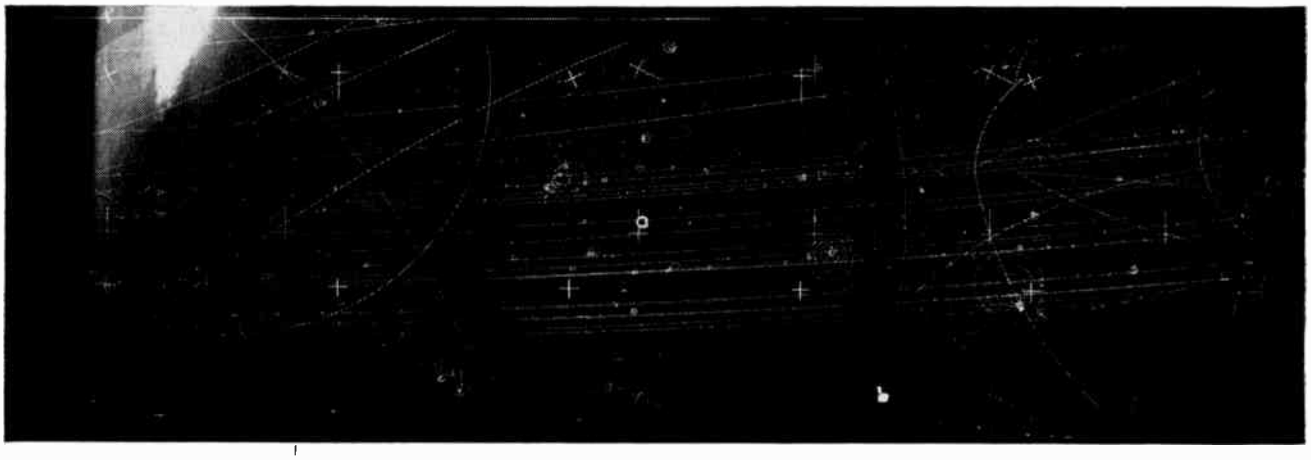
Particle tracks in the bubble chamber are photographed

* *The British National Hydrogen Bubble Chamber. Electronic Engng.* 35, 245 (1963).

Fig. 1. (right) Assembly of the British National Hydrogen Bubble Chamber at C.E.R.N. Geneva

Fig. 2. (below) Sample photograph of collisions of two incoming negative kaon particles. The spiral shaped tracks are those left by electrons trapped in the magnetic field (up to 15 000 gauss) of the bubble chamber. They have no particular meaning for the experiment





by three separate cameras giving a stereo record of each event which lasts for a fraction of a millisecond.

The first experiments involve the study of the interaction of high energy negative kaons in the hydrogen of the bubble chamber. It is hoped that further evidence may be found for a new particle called the omega-hyperon and detailed studies of various resonance states will also be made. The large chamber, combined with a very good beam of K^-—particles, should provide an immense amount of data for analysis.

A series of experiments intended to provide accurate information about the interaction of high-energy pions and protons will also be carried out.

It is hoped that during the next year about one million pictures will be taken at C.E.R.N. at a rate of up to 25 000 per working day. Specially designed equipment has been installed at C.E.R.N. and in many laboratories in member States of C.E.R.N. for examining and measuring the photographs.

Experimental runs were carried out during July this year with the proton-synchrotron operating at 19·2GeV and the photograph shown in Fig. 2 is one of the series taken during the run.

The purpose of the experiment, labelled T49 in C.E.R.N.'s experimental schedule, was two-fold:

(a) to produce and study the omega-minus particle, discovered in February in the U.S.A.;

(b) to investigate the production of other "strange particles", such as the lambda and sigma hyperons.

The photograph of the particle track in Fig. 2 shows the collision of two incoming negative kaon (K^-) particles with protons in the liquid hydrogen. The probable explanation is as follows: The bottom left-hand side of the photograph shows the production of an electrically charged strange particle, a positive sigma particle. After travelling about 1cm the sigma particle decays into a proton (p) and a neutral pion. The latter leaves no track because it is electrically neutral.

On the right-hand side four charged particles are produced with at least one neutral 'V' particle. One of the charged particles (a positive pion) comes to rest after travelling about 25cm in the chamber and then decays into a negative muon (the very short sections of tracks in the upper right-hand corner of the photograph) and a neutrino (not seen). The muon in turn decays into a positron (e^+), and a neutrino/antineutrino pair (not seen).

Integrate-and-Dump Operator for Serial Correlators

By R. M. Seeley*

A need has existed for a finite-time integrator to be used in serial digital correlators. The principal requirements are ability to accept a high bit rate with stable integration for random rates, fast and complete dumping, and stable transfer of the total to a low impedance output. A device meeting the need is described here; in addition a novel fast flip-flop is disclosed. Overall accuracy averages 2 per cent with a 2Mc/s data stream input, using a 300μsec integrate and 2μsec dump period.

(Voir page 650 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 657)

SERIAL digital correlators are widely used because of the availability of serial delay lines that are able to hold many bits—up to several thousand. The three operations in digital correlation are variable delay, multiplication, and finite-time integration. These operations are shown in the equation:

$$c(t) = 1/T \int_{T/2}^{T/2} f(\tau)\; g(t-\tau)\, d\tau$$

The variable delay and the storage of many bits is accomplished by the DELTIC[1] method; multiplication by a mod-2 adder, and finite-time integration by one of several devices—for example, a binary counter. The integration is over a finite time because a finite number of bits is held in the serial memories; it is desirable to integrate over the time of one 'revolution' only, with the start synchronized to the point at which the DELTIC acquires a new sample.

A true integrate-and-dump operator best fills the DELTIC application. The operation in time indicated by:

$$b(nT) = 1/T \int_{(n-1)T}^{nT} a(x)\, dx; \quad -\infty < n < \infty$$

$$= 1/T \sum_{t_i=(n-1)T}^{t_i=nT} a(t_i)\, \delta t_i$$

has an impulsive response $h(t) = 1/T$ from 0 to T; $h = 0$ elsewhere. The corresponding spectral operation is that of a filter with true $\sin X/X$ response followed by a sampler for evaluation at nT, which is not the spectral operation of an ordinary linear filter. This point is made because linear filters are sometimes used as approximate integrating devices in this application.

The integrate-and-dump operator described here makes use of principally analogue circuits (Fig. 1) to decrease size and cost. A binary counter capable of summing over 1 000 bits, for example, would require at least four times the space and cost required by the analogue circuit. The analogue precision averages 2 per cent over the range, which is quite sufficient for correlator applications.

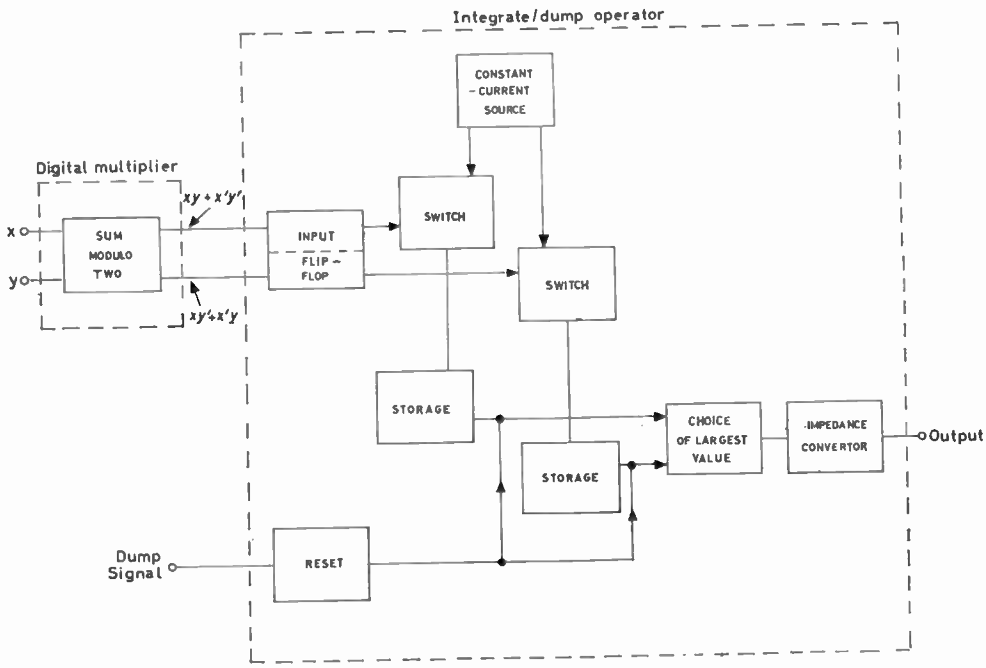


Fig. 1. Multiply and integrate/dump functions

Fig. 1 shows the principal parts of this analogue integrate-and-dump device. This particular form is adapted to correlation processing on a band-pass signal[2] where the phase-sensitive properties of the correlation function are not desired; that is, an absolute-value output is desired independent of whether the functions undergoing correlation are in or out of phase with each other. The basic idea is to convert the 1, 0 information to corresponding time lengths, to generate unit currents during each time length, and to sum the resulting charges.

Since either all 1's or all 0's are a desirable correlation

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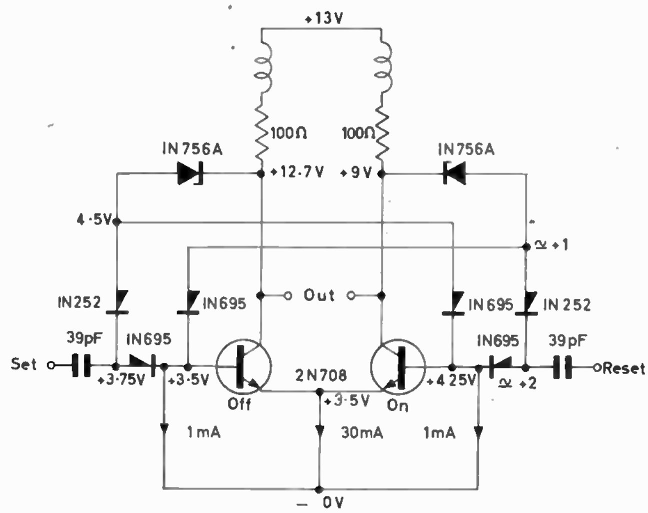

Fig. 2. Flip-flop circuit

The inductors are a National Cail Co. type R45-1 ferrite r.f. bead choke; l.f. inductance 0·3μH

output, each condition is provided with its own path of charge summation and the greater of the two is read out. A condition of partial 1's and 0's causes each path to charge only partially toward full-scale. At the no-correlation condition, each charges half-way; the output scale has the bottom half suppressed so that the no-correlation condition reads out as 0 volts and full-scale condition as +10V, even though internally the charging capacitors are reset to −10V at the end of each interval (T) of computation.

Input Flip-Flop

The delayed-and-multiplied information is presented to the integrator as a binary pulse stream with a 2Mc/s clock rate. Since the pulses as presented are asymmetrical in duty cycle, the integrator contains an a.c.-triggered input flip-flop (Fig. 2) capable of actuating only on one edge of the incoming pulses—the edge produced by pnp transistors turning on. This flip-flop* has several interesting features: it can only be actuated by positive-going input transitions; output transition times and input charge requirements are inherently small; the base circuit is automatically temperature compensating so that a known 'off' voltage can be designed-in; and the design is suitable for small collector swings relative to the collector standing voltage.

Fig. 2 shows the stable state with typical d.c. voltages present. Note that the combination of diodes from collector to base on the 'off' side is such as to keep the input diode 'on', so that a positive input swing wastes little voltage in causing transistor turn-on. The diode combination makes the voltage difference between 'on' and 'off' bases equal to the drop across the silicon 1N252; since this is nominally equal to transistor 'on' V_{BE}, the 'off' V_{BE} is nominally zero. Choice of other diode types would allow apportioning the 'off' V_{BE} to suit the designer.

On the 'on' side, the diodes are all nominally 'off' in the string from 'on' collector to 'on' base. Therefore, no input pulse—either positive or negative—can trigger the flip-flop on the 'on' side, and obviously only positive transitions will have an effect on the 'off' side. The independence of negative input pulses is important. Many designs have diode steering so that there is nominal protection against transient inputs of the wrong polarity; however, often these diodes carry current, and when the undesired pulse turns off that current, the transistor input circuit is sufficiently upset to cause a change of state. Here the diode carries no current and cannot upset anything.

There are two limitations to this flip-flop design:

(1) The output swing should be greater than about 2V

(2) The maximum collector emitter voltage should be greater than 9V.

The first limitation results because the total manufacturing voltage variation of avalanche diode and steering diode is about 0·9V. There must be an additional amount of swing to ensure that the proper set of diodes controls the 'off' base—say 1V; the sum, then, is 2V. The second limitation occurs because it is desirable to use sharp breakdown diodes; therefore, breakdowns of not less than 8V are necessary. Of course, a more complicated design using round-shouldered diodes would be possible.

This structure has an unusual transient input impedance: the positive-going transition on the waiting input is lightly loaded because there is no coupling to the opposite collector until the output transition turns 'on' the 1N695 that is connected to the 'off' base (i.e., when the loop is fully active); coupling to the near collector is present only for a period long enough to turn 'off' the near 1N252 (which requires about 25pC). The last coupling time can be considerably reduced by using a better diode—for example, the HD5000, for which 5pC are required in this circuit.

The charge required by each side to change state totals about 80pC, which must be provided from input and regeneration; of the total 35pC are due to transistor transition capacitances and wiring capacitance, while the remainder is due to the required charge change in the base to control 30mA collector current, plus the 1N252 turn-off charge.

Loop coupling during transitions is provided by the common emitters plus the breakdown diode back capacitance in series with the stored charge of the cross-coupled 1N695. The 1N695 in cascade with the input capacitor is also useful for its stored charge (25pC/forward mA); this charge serves to define the voltage at the diode-capacitor junction, which would otherwise be indeterminant.

The output transitions are clean, with 5nsec transition times and 5nsec delay when non-marginal input charge is supplied. Relatively slow input transitions (50nsec long) require 70pC input charge; the amount remains less than 100pC (with faster input transitions) to 20Mc/s. The repetition capability of this flip-flop does not parallel the resolution capability principally because of the relatively large breakdown diode capacitance.

Switched Integrator

The timing from the input flip-flop is within 1 or 2nsec of being perfectly symmetrical, and the next integrator stages are driven with fast, even transitions. However, the integrator stages in Fig. 3 are not sufficiently fast to maintain the timing definition supplied by the input. This is because the design took place when the 2N1259, with a 60Mc/s f_t, was the fastest pnp silicon transistor available.

Silicon transistors were necessary to hold capacitor leakage to negligible levels; pnp seemed necessary to maintain the voltage levels symmetrical around zero volts (a convenient and stable power supply for setting one of the output limits). With the output swing desired, a value of V_{CB} greater than 20V was necessary.

The switched integrator (Fig. 3) is a symmetrical grounded-base pair fed from a constant-current emitter

* *Patent applied for.*

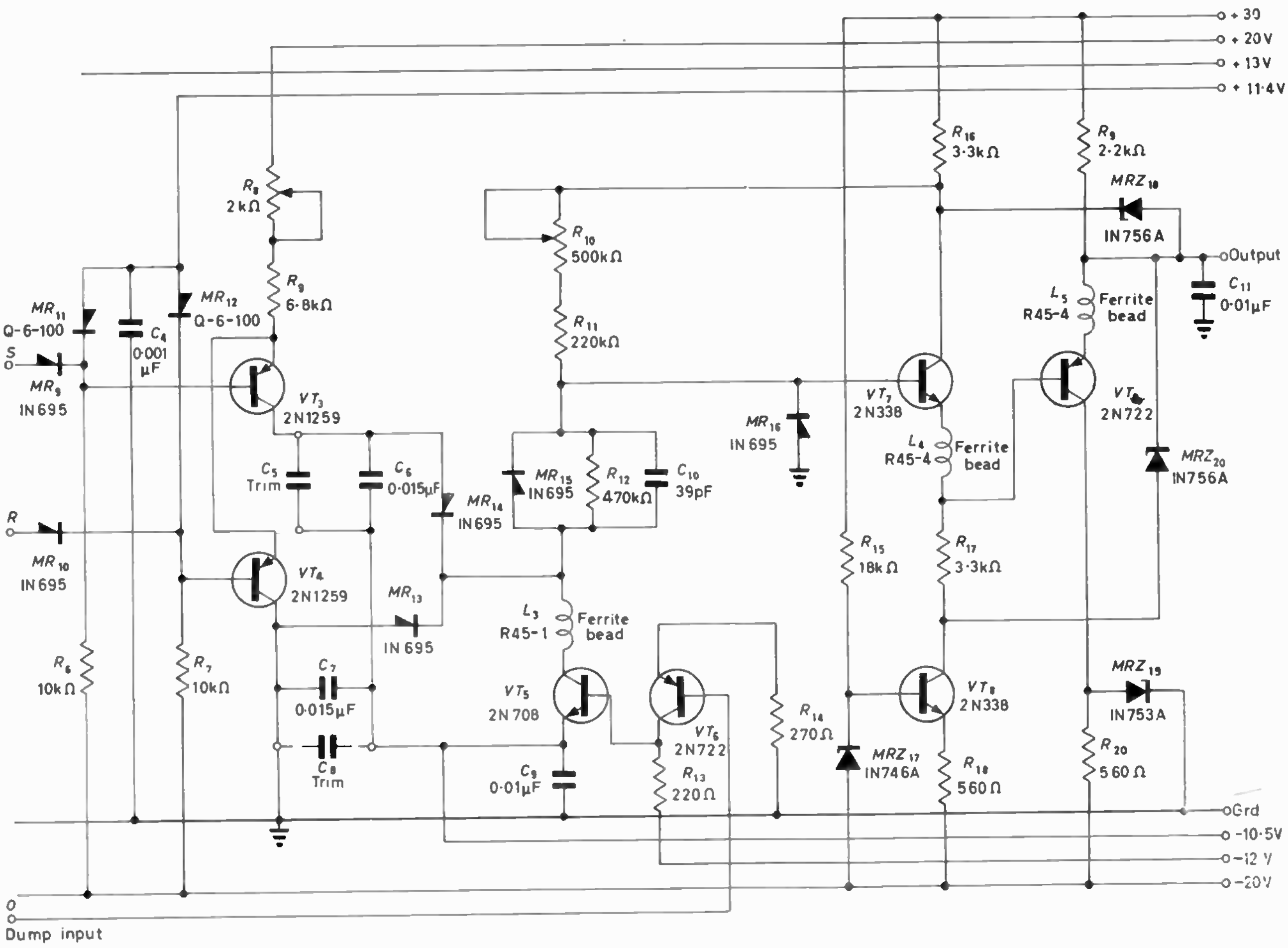

Fig. 3. The complete circuit

source to obtain the fastest speed and best definition of charging current possible with the 2N1259. The variation in speed among randomly chosen pairs causes corresponding variation in charge alternately transmitted to the integration capacitors. The total charge required by emitter transistor capacitance and diffusion capacitance in each transistor is variable over perhaps a 20pC range from unit to unit. Thus, there is a differential-control-charge uncertainty of about 4 per cent (compared to the 500pC charge to be transferred every bit time) and an additional 4 per cent static uncertainty due to variations of α_0 and V_{BE}.

The static effects can be balanced by adjusting the relative size of the two integrating capacitors for equal full-scale values with in-or-out-of-phase 100 per cent input correlation. The dynamic effect is most apparent at 0 per cent input correlation, where one would expect a 4 per cent offset referred to a 20V swing, or 8 per cent on the suppressed read-out scale. Note that increased input correlation decreases the average bit rate. The average bit rate, even at 0 per cent input correlation and 2Mc/s DELTIC rate, is still only 1Mc/s because the data is random. Thus, the worst error in actual usage with a 2Mc/s DELTIC is 4 per cent, which decreases rapidly with increased input correlation. The static linearity, offset, and stability is well within 1 per cent (the theoretical non-linearity is less than 0·1 per cent for a transistor having an r_c value greater than 1MΩ).

The Q 6-100's at the 2N1259 bases are supplied to clamp the bases to a known source for the 'on' condition with very small charge demanded from the flip-flop to turn them 'off'. The 1N695's supply all charge required for turn-off (some 70pC).

The 2N708 in the dump circuit has a difficult job to perform, because it must remove 0·3μC in 2μsec; the corresponding average current is 150mA and the peak 250 to 300mA. The 2N708 turns on in a short time—about 30nsec—to a peak current of several hundred milliamperes, while the V_{CE} could still be 20V if it were not for the R45-1 choke, which quickly knocks down the V_{CE} and also restrains current build-up in the collector circuit. Without the coil, the transistor locally heated in about 0·5μsec to such an extent that the β could drop to near zero; the time required for the heat to move out to the case and dissipate was as much as 50μsec. In addition to this effect, only one manufacturer's version of the 2N708 could support the currents demanded, even at low power, apparently because of epitaxial construction. The circuit shown accomplishes dumping to within 0·1V of the reference in 2μsec.

Impedance Convertor

The purpose of the impedance convertor circuit (the right-hand portion of Fig. 3) is to transfer the integrated voltages to a low-impedance output stage without causing appreciable charge leakage from the integrating capacitors. The two cascaded emitter-followers have two separate feedback paths to maintain the required base current and the required emitter current constant, no matter how the input voltage varies. It happens that one path constitutes negative feedback; the other, positive. Assuming the adjustment of base current supplied from the collector junction point is made correctly at 0 volt input, so that no additional input current is required, then none should be required anywhere in the voltage range. Actually, the best low-input-current capability of the device is not needed in this application, where there is only 30mV integrated offset produced by each microampere of unbalance current in 300μsec.

The incremental input impedance of this portion of the circuit is several megohms to 1Mc/s, being $\beta_1\beta_2R_L$ to $f_{\beta 1}$ or $f_{\beta 2}$, whichever is lower, where R_L is the total incremental impedance as seen looking out from the second emitter, approximately 1kΩ in this case. Of course, it is possible to increase the impedance or bandwidth by choosing higher values for R_L or wider-band transistors. The offset of the circuit is nominally zero, and temperature compensation is within a few millivolts. The input current can be adjusted to a small part of a microampere, but seems to hold only to the order of several microamperes in long-time usage because of β_1 ageing and temperature effects (which could be compensated by an emitter network in the first stage).

At present, a similar integrate-and-dump device is being developed for operation from 16Mc/s DELTIC rates.

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2. Ackerman, C. L., Miller, C. S., Brown, Jr., J. L. Theoretical Basis and Practical Implications of Band-Pass Sampling. *Proc. Nat. Electronics Conf. 18*, 1 (1962).

A 100V 100mA Regulated Power Supply Using Transistors

By R. E. Aitchison* and W. S. Lamond*

This article describes a 100V 100mA regulated power supply module. The design illustrates some principles which can be applied in the design and protection of transistor power supply circuits when the supply voltages exceed the voltage breakdown ratings of the transistors. A number of these modules can be cascaded to give a higher voltage supply.

(Voir page 651 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 658)

When the supply voltage exceeds the voltage breakdown rating of the transistors, the design of the circuits must incorporate special features to ensure that both during normal circuit operation and during switching transients the transistors cannot be damaged by voltage breakdown, which occurs virtually instantaneously and cannot be prevented by the use of fuses.

The basic circuit is a series degenerative type of d.c. regulator and is illustrated in Fig. 1. The pre-regulated supply V_1 is used to give high performance over a range of output and input voltages, and a separate reference V_2 so that a number of identical units can be controlled by the one reference and so cascaded and give a range of output voltage.

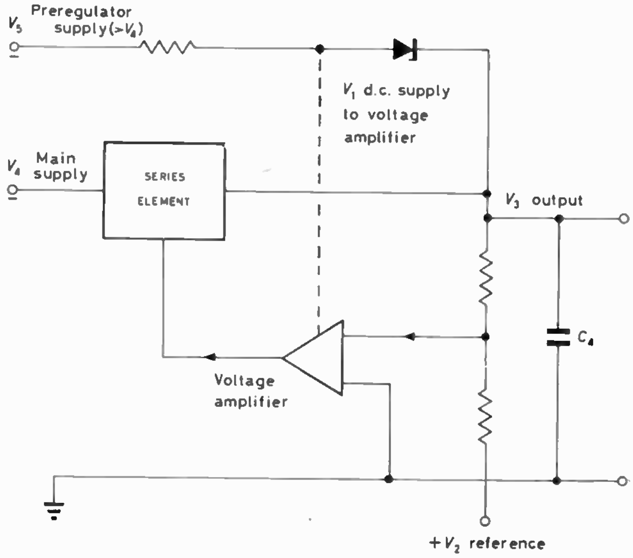
However, main interest is in the problems introduced by voltage breakdown. The voltage amplifier must operate at the voltage V_1 which is higher than the full output voltage V_3. The series amplifier must operate at the difference between V_4 and V_3, while during switching-on transients it must cope with the full supply voltage V_4, as the output capacitor C presents a virtual short-circuit across the output terminals during the initial charging period. Similarly, a momentary accidental short-circuit across the output terminals has the same effect. When $V_4 \simeq 100$V or more, it is clear that transistors used in the circuit will be exposed to voltages in excess of the breakdown ratings.

In a practical circuit the methods used to cope with voltage breakdown are as follows:

(a) Use of a number of transistors in series, with appropriate voltage sharing, to give higher breakdown voltages.

(b) Where necessary, a resistance between base and emitter chosen to give an adequate compromise between voltage breakdown and current gain. In this way the higher collector-base breakdown voltage is approached, while the common emitter gain is not appreciably reduced.

Fig. 1. Basic circuit arrangement of a series degenerative power supply with an external reference V_2 and a pre-regulated supply V_1 to illustrate voltage breakdown problems



* University of Sydney, Australia.

(c) Identification of switching transients and their suppression of limitation using a Zener diode which is inoperative during the normal operation of the circuit.

Practical Circuit

The circuit of a particular 100V 100mA unit incorporates these principles as shown in Fig. 2.

The voltage amplifier is formed of a cascade of transistors in series (*VT*$_7$ to *VT*$_4$, 2N591, V_{CB} 32V). The voltage divider ensures equal voltages across each transistor, and the emitter-base resistors of 3·3kΩ give a high collector-emitter breakdown voltage. This combination of four transistors acts as a single transistor of much higher breakdown voltage. with the input to the base of the first transistor (*VT*$_7$) and output from the collector of the last (*VT*$_4$). The same principle is used for the series transistors *VT*$_1$ and *VT*$_2$ (ASZ16, V_{CB} 60V, V_{CE} 48V), but the voltage divider is of lower impedance, and again resistors (of 470Ω) are placed between base and emitter to ensure a high breakdown voltage.

However it would be impractical to design the series amplifier to cope with the switching-on transient of approximately 140V across the series combination *VT*$_1$*VT*$_2$.

A Zener diode *MRZ*$_1$ of 56V breakdown is connected across *VT*$_1$ and *VT*$_2$, and a series limiting resistor (1kΩ) keeps the peak current through this Zener diode within its ratings. During switching-on, the peak voltage across *VT*$_1$ and *VT*$_2$ combined is kept to less than 70V, but during

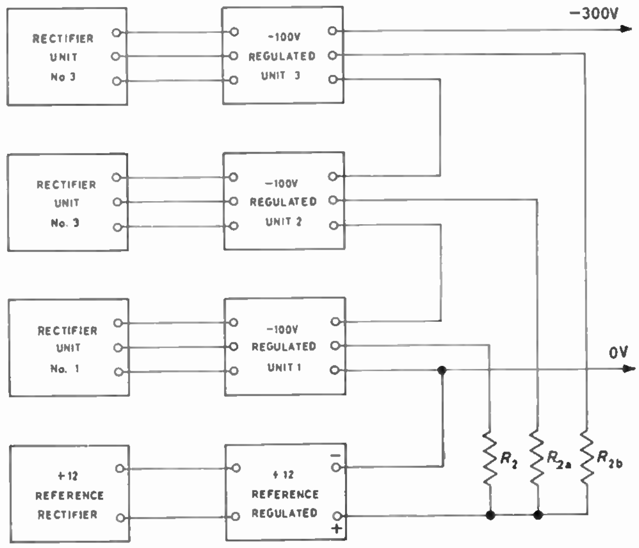


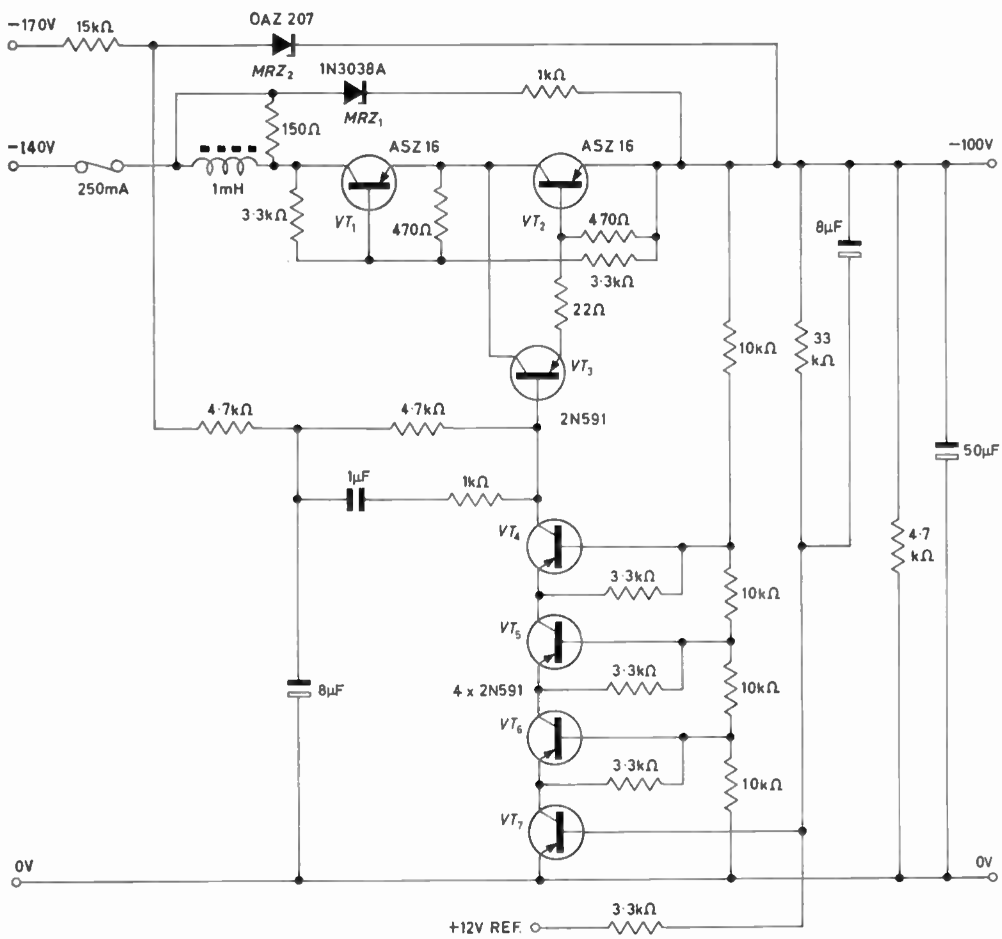
Fig. 3. Method of grouping three modules to give a 300V 100mA supply

normal operation *MRZ*$_1$ is non-conducting and has no effect on the circuit performance. A small 1mH choke and damping resistor (150Ω) also help reduce the switching transients.

The remaining details of the circuit follow accepted and routine design procedures. With a bridge rectifier power supply giving 10V ripple (peak-to-peak), the output ripple at full load is less than 3mV peak-to-peak, and is mainly determined by 50c/s hum introduced by earth wiring. The output impedance for a no load to full load change of output current is less than 20mΩ, and the mains regulation factor (change in mains voltage/change in output voltage) for a ±10 per cent change in mains is in excess of 1000:1. For completeness the method of cascading three units to give a 300V 100mA variable (270 to 330V) supply is shown in Fig. 3.

Although the principles are illustrated by their use in this particular power supply, designed around inexpensive and readily available germanium transistors, the same principles can be used in any similar application in circuits in which the supply voltage exceeds the particular transistor rating, and dangerous switching transients can occur.

Fig. 2. The circuit of the 100V 100mA power supply



Demodulation Circuits for PAL Colour Television Receivers

(Part 2)

By W. Bruch*

(Voir page 576 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 583)

Demodulation with Delay Lines with Respect to Arbitrary Axes, e.g. $(R' - Y')$, or X', Z'

It is evident from the NTSC equations that every colour difference signal can be obtained from a linear combination of two other such signals. Thus I' and Q' can be derived from a combination of $(R' - Y')$ and $(B' - Y')$ and vice versa, and similarly X' and Z' are related to I' and Q'. A similar possibility of combination can also be expected to exist for the component carriers F_I' F_Q' produced by the signal separator of Fig. 1. A linear combination of these quadrature signals would result in a phase-modulated signal of the NTSC type. If, however, the components are made to be in phase they can be added and subtracted like signals provided that the phase reversals of I', $(I', -I', I', \ldots)$, introduced on alternate lines at the transmitter are cancelled. An electronic switch, such as was described above for reference phase switching, can also be used to cancel these phase reversals in the F_I' channel. A wideband 90° phase shifting network is then used to turn the colour carriers F_I' and F_Q' into phase alignment so that they can be added and subtracted like voltages of video frequency thus producing a new carrier of the same frequency and phase which now contains only the required colour difference modulation. For greater clarity in designing the carrier frequency matrix, the NTSC equation may be rewritten as

$$R' - Y' = 0{\cdot}62Q' + 0{\cdot}96I'$$
$$G' - Y' = -0{\cdot}65Q' - 0{\cdot}27I'$$
$$B' - Y' = 1{\cdot}70Q' - 1{\cdot}11I'$$

or as a further simplification with all signals referred to $+Q$

$$1{\cdot}61\ (R' - Y') = +Q' + 1{\cdot}54\ I'$$
$$-1{\cdot}55\ (G' - Y') = +Q' + 0{\cdot}42I'$$
$$0{\cdot}59\ (B' - Y') = +Q' - 0{\cdot}65I'$$

This leads immediately to the circuit of Fig. 16. One of the signals, here F_I', is required both as a positive and a negative quantity. The centre tap of the push-pull secondary is therefore kept at F_Q' potential.

Unlike the NTSC demodulator outputs, the signals from the demodulators following the matrix network of Fig. 16 are not of equal amplitude. These differences are equalized in the output amplifier stages. If equal phosphor sensitivities are assumed, i.e. the colour tube will give a 'white' picture on all three channels with equal control grid potentials, the gains in the three colour output amplifiers will have to be

$$|V_{(R'-Y')}| = |V_o/1{\cdot}61| = -0{\cdot}62V_o$$
$$|V_{(G'-Y')}| = |V_o/-1{\cdot}55| = -0{\cdot}65V_o$$
$$|V_{(B'-Y')}| = |V_o/0{\cdot}59| = 1{\cdot}70V_o$$

or, expressed in terms of $(R' - Y')$

$$|V_{(G'-Y')}| = |-1{\cdot}04V_{(R'-Y')}|$$
$$|V_{(B'-Y')}| = |2{\cdot}74V_{(R'-Y')}|$$

The following expressions would serve to obtain X' and Z' as defined by Carnt and Townsend[7]:

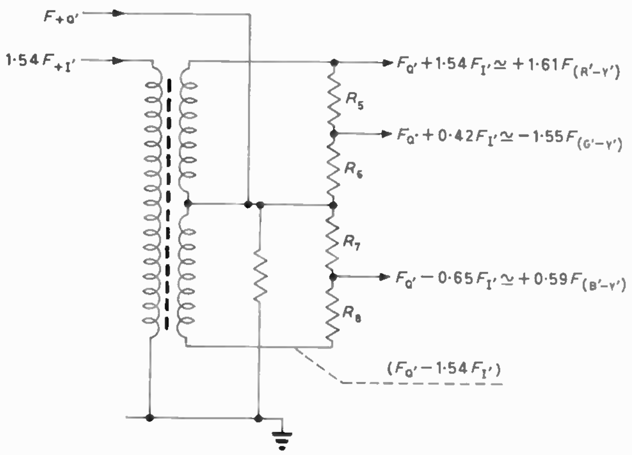
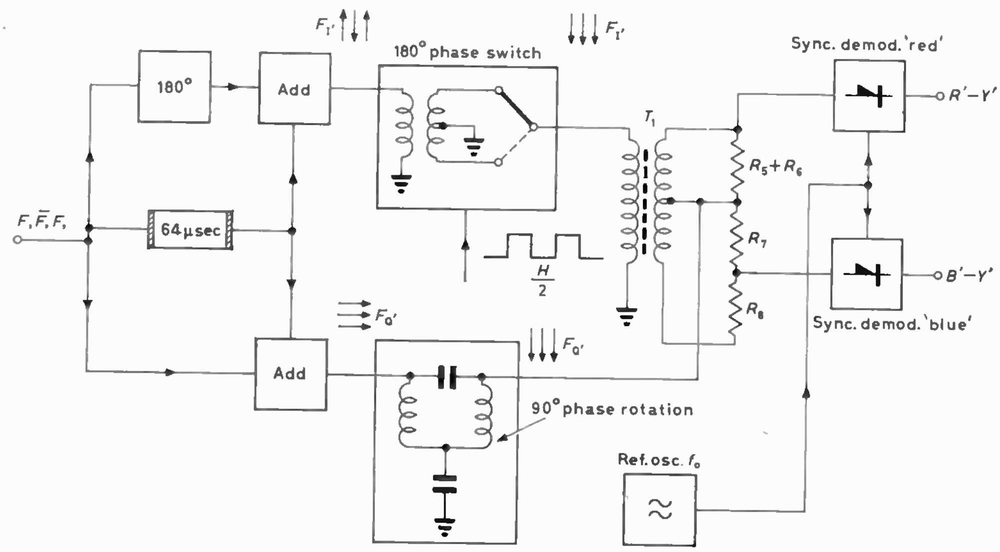


Fig. 16. Carrier frequency divider matrix

Fig. 17. (R′−Y′) and (B′−Y′) demodulator for PAL$_s$ using matrix of Fig. 16



* *Telefunken A.G., Hanover.*

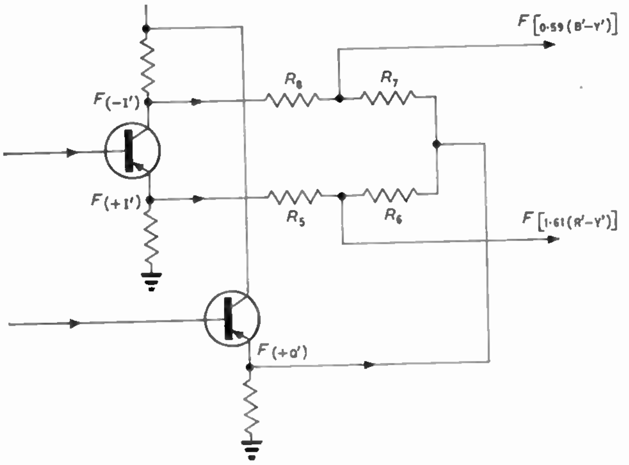


Fig. 18. Carrier frequency resistance bridge matrix

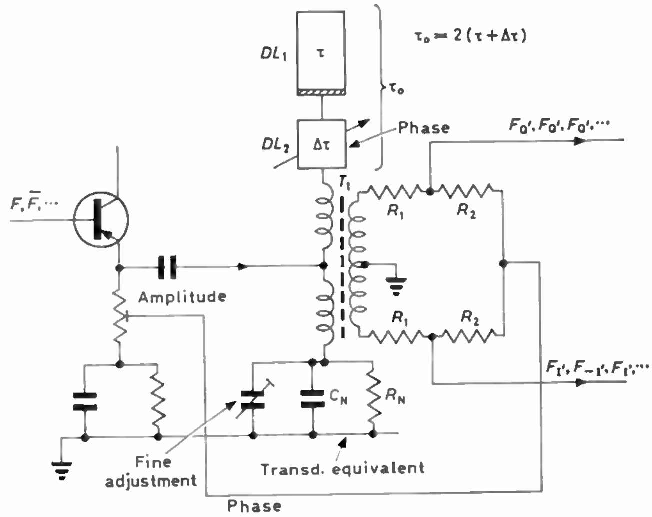


Fig. 20. PAL$_{DL}$ demodulator with reflection type delay line and phantom circuit to separate forward and return signals

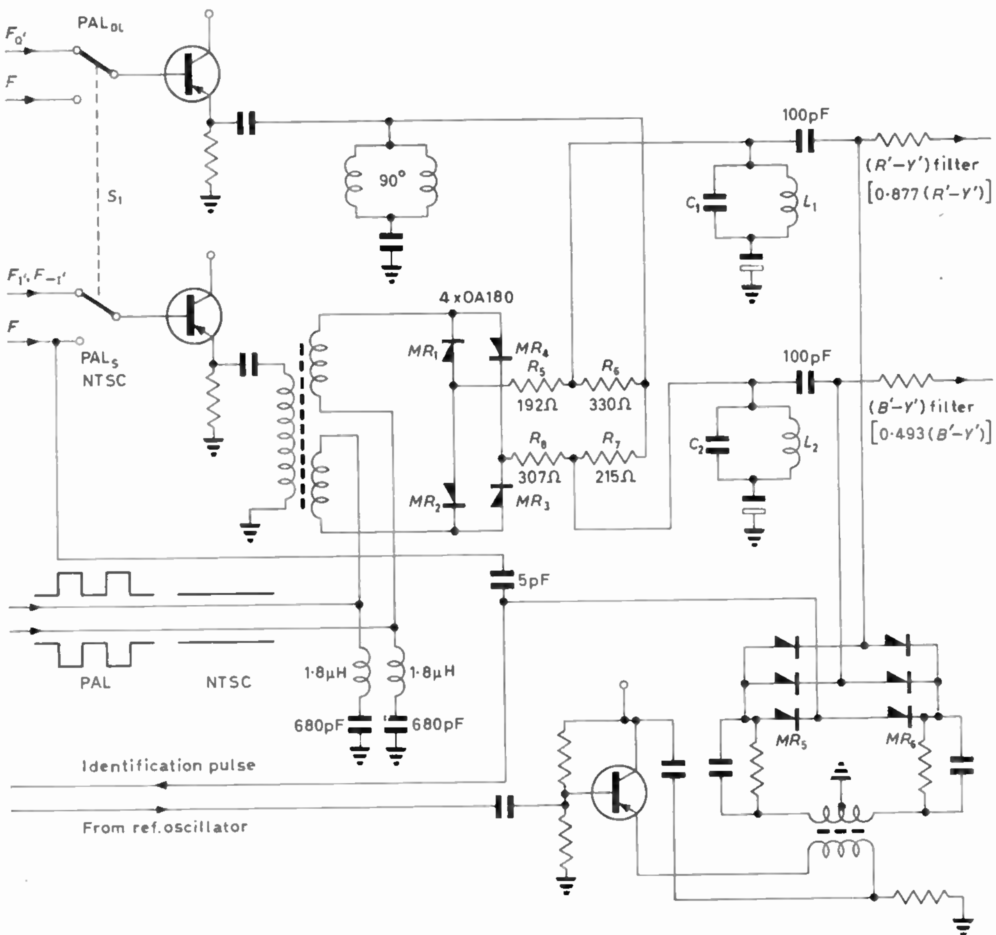


Fig. 19. Universal PAL decoder for (B'−Y'), (R'−Y'), X'Z' or I' and Q' demodulation with PAL$_{DL}$ and selection facility for PAL$_S$ and NTSC operation

$$0{\cdot}90\,X' = Q' + 1{\cdot}07\,I'$$

$$0{\cdot}60\,Z' = Q' - 0{\cdot}28\,I'$$

The block diagram Fig. 17 shows such a demodulator with push-pull transformer for $(B' - Y')$ and $(R' - Y')$. The two carriers are rotated into phase, by means of a shifting network, after the electronic switch in the F_{I}' signal path has cancelled the 180° reversals. The ratio of R_7/R_8 determines the amount of F_{I}' mixed with F_{Q}' to produce $F_{(B'-Y')}$. If a switch S_1 is additionally provided in the signal separating circuit (for instance for demonstration and comparison purposes) the receiver can also be used for PAL$_S$. Even NTSC signals can be received if the diode switches are kept continuously in their appropriate state (see S_2 in Fig. 4). The resulting NTSC circuit was described[9] although it is not used in practice. A universal demodulator switchable to PAL_{DL}, PAL_s or NTSC can be designed on the principle described in which the demodulation axes can be chosen arbitrarily merely by changing a few resistors. The same circuit is then used in such a demodulator for $I'-$, $Q'-$ demodulation as for $(R'-Y')$, $(B'-Y')$ or X', Z'. The same design considerations for the resistance bridge matrix (Fig. 1(b)) also apply to the high frequency matrix. F_Q' is fed into the bridge at the right-hand junction of the bridge (Fig. 18) while the arms are supplied with F_I' and F_{-I}' in equal amplitudes. Polarity reversal of F_I' is achieved by the addition of a further diode pair MR_3, MR_4 to the F_I' commutating switch which also passes F_I' in the reverse direction as is seen from Fig. 19. In computing the potential divider ratio of the matrix network the internal resistances of the switches must be added to the resistances R_5 and R_8 respectively. For the diodes a forward resistance of 23Ω was calculated. The complete circuit with component values for demodulation with respect to $B'-Y'$ and $R'-Y'$ is given in Fig. 19. Only the values of the resistors R_5, R_6, R_7 and R_8 need modification if demodulation with respect to X' and Z' or I' and Q' is required. As in the circuit of Fig. 17 the reference carrier $\angle I'$ is here also supplied to the two synchronous demodulators in the same phase. A third pair of diodes MR_5 and MR_6 serves to rectify an identification pulse carried by $\angle Q'$ which is derived from the separator circuit. A simple LC arrangement is used to turn the F_Q' signal into phase alignment with F_I' and thus into phase with the reference carrier. It is seen therefore that the reference carrier is fed into all three synchronous demodulators in the same $(\angle I')$ phase. Two parallel resonance circuits L_1C_1 and L_2C_2 are turned to the chrominance carrier frequency and are positioned so as to avoid short-circuiting the switch at this frequency.

This circuit was successfully used for several EBU demonstrations and, more recently, also for a number of propagation surveys both with $(B'-Y')$, $(R'-Y')$ and $X'-$ and $Z'-$ demodulation. Of all the PAL circuits it has so far proved the most adaptable to the combined requirements of NTSC, PAL_{DL} and PAL_s. By exchanging but four resistors the demodulation axes can be shifted into any desired direction. Such a universal decoder can be operated with any output circuit, since any combination of colour-difference carrier signals can be formed in the resistance matrix R_5, R_6, R_7, R_8. Which combination of I' and Q' is required for any particular circuit can easily be ascertained without a laborious determination of the demodulation axes in the NTSC system (as is required for instance with X' and Z').

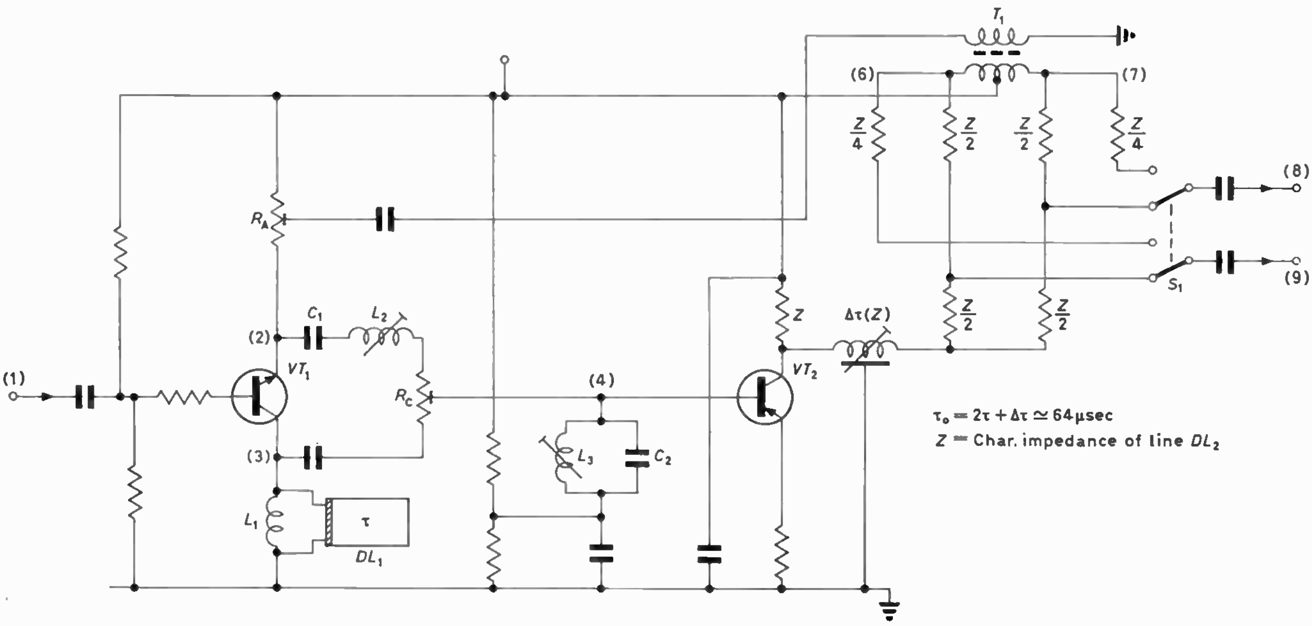


Fig. 21. PAL_{DL} demodulator with reflection delay line using I′ and Q′ separation without isolation of forward and return signals at delay line

Circuits with Simplified Delay Line

The ultrasonic delay line is the only non-standard component in a PAL_{DL} (and also SECAM) colour receiver. With mass-production methods it should be possible to produce these lines economically. Nevertheless considerable thought has been given to the possibility of simplifying and cheapening the delay lines. It would be feasible to cut a line, consisting of a glass rod and two terminal transducers, in half and to polish the free end. This would give two new reflection type lines. The wave applied through the transducer to one end of the rod will travel to the polished free end where it undergoes total reflection to return to the transducer where it can be picked off. Since the glass rod is traversed twice the total delay thus achieved with a line of half the original length is still near enough 64μsec. If this type of line is made part of a bridge circuit, energized at colour carrier frequency, its length can be ground accurately by a fully automatic servo system. This is not only an economical production method but, because of the resulting accuracy, the fine adjustment line DL_2 can be dispensed with. Fig. 20 shows a reflection line used with a phantom circuit well known in telephony. To separate the direct and delayed signals a carefully balanced push-pull transformer T_1 and an accurately adjusted transducer equivalent C_NR_N to neutralize the direct signal must be used. Furthermore, the temperature dependence of the transducer self-capacitance must be corrected. It was therefore found rather difficult to make an accurate equivalent of the delay-line input impedance of adequate band-width and attention was turned to other possibilities.

It was suggested by W. Scholz[4] that it is not necessary

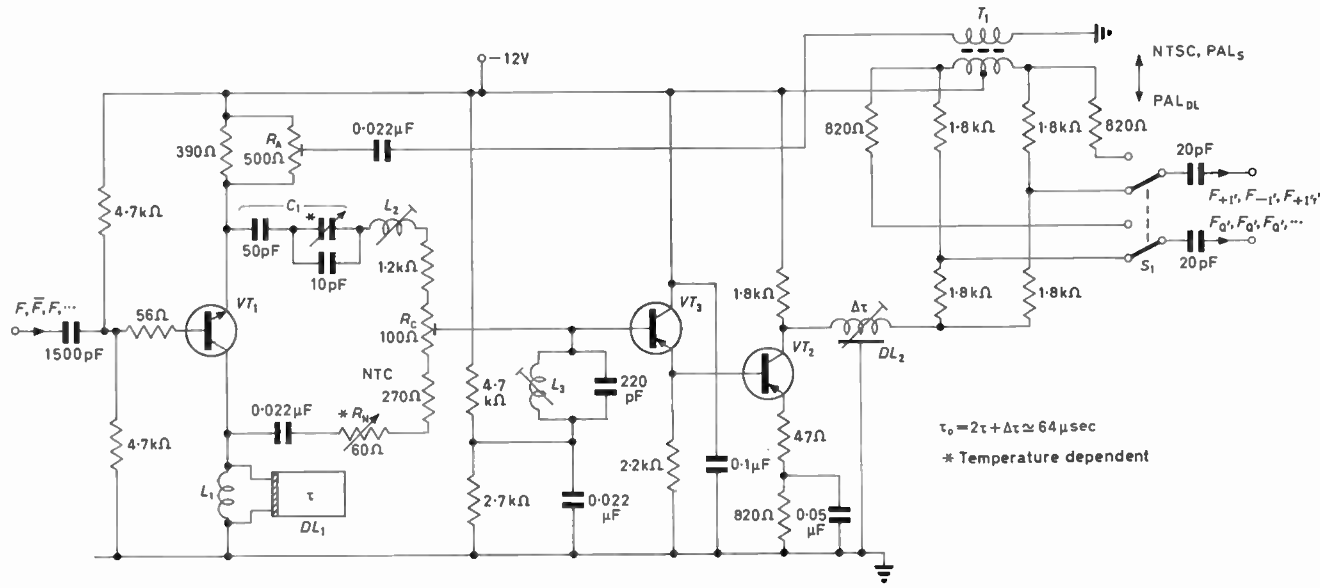


Fig. 22. Practical arrangement of circuit of Fig. 21

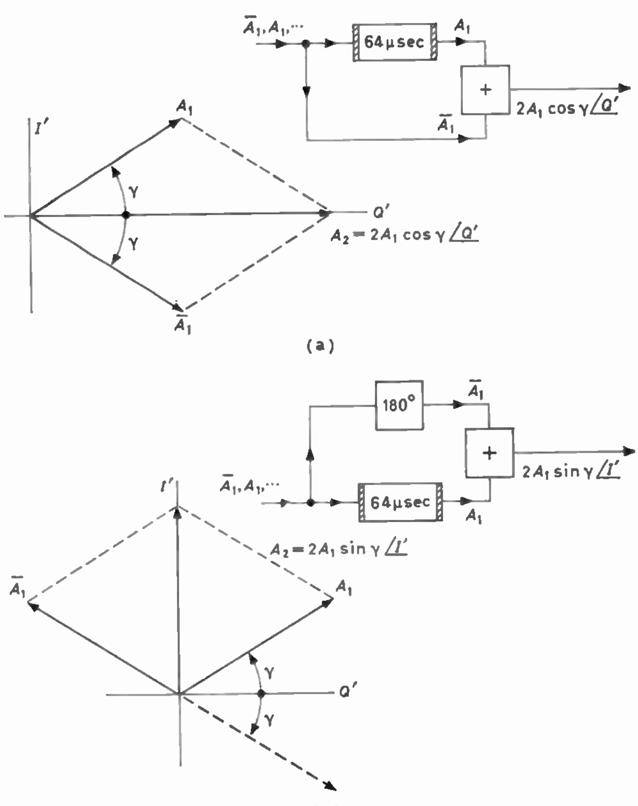


Fig. 23. PAL$_{DL}$ demodulation with carrier injection ahead of delay line
(a) Reference phase mirroring for Q' channel
(b) Reference phase mirroring for I' channel

with a PAL-delay-line demodulator to have the direct and delayed signals completely separated. At the delay-line transducer input end reflected signals such as for instance $(F + a_1\bar{F})$ exist simultaneously, where F and $\bar{F}$ are consecutive line signals and a_1 the attenuation constant. If from this signal $(F + a_1\bar{F})$ the direct signal is subtracted namely with amplitudes $(1 + a_1)$ and $(1 - a_1)$ respectively the following relations hold, provided the phase delays are accurate,

$$F + a_1\bar{F} - (1 + a_1)F = a_1\bar{F} - a_1F = -2a_1F_{I}'$$
$$F + a_1\bar{F} - (1 - a_1)F = a_1\bar{F} + a_1F = +2a_1F_{Q}'$$

yielding the required $I'-$ and $Q'-$ signals.

For the first practical tests of this method no delay-line of the required length was available and a short line had to be used. Experiments were therefore made with a circuit as shown in Fig. 21 and 22, in which the direct signal is subtracted from the mixed signal at the line transducer (point 3 in Fig. 21) in such a way that only the delayed signal remains.

L_2 and C_1 are used to match the direct signal from point 2 to that of opposite phase at point 3. Thus a tapping point must exist along the length of resistor R_o at which the direct signal vanishes and only the delayed signal remains. Slight phase correction can be achieved by means of L_2. The delay time is set accurately, as in previous PAL$_{DL}$ demodulators, with an additional time adjustment line DL_2. In future, when the half-length delay line ground to fine tolerances becomes available, inclusion of the fine adjustment line DL_2 should be unnecessary. The remaining parts of the circuit are similar to those used with the full-length delay line. The circuit of Fig. 22 shows the first experimental prototype in which no effort was made to minimize the number of transistors and other components.

PAL$_{DL}$ Demodulation with Carrier Injection Ahead of the Delay Line

As mentioned above, demodulation can be achieved either with synchronous demodulators or with standard rectifiers and an added reference carrier signal, the two methods being equivalent. It is possible, furthermore, to supply the two demodulators with their reference signal via the common chrominance path ahead of the delay line. The reference carrier need then be fed into the system at a single point only, though its phase has to be switched over during alternate line intervals. As is seen from Fig. 23(a) the reference carrier may be injected over a wide phase range in the four quadrants between the I' and Q' signals. Addition in the Q'- carrier channel of the direct and delayed components will always result in a vector along the $(\pm Q')$ axis if the reference signal is mirrored in the Q' axis during alternate lines. If, for instance, the

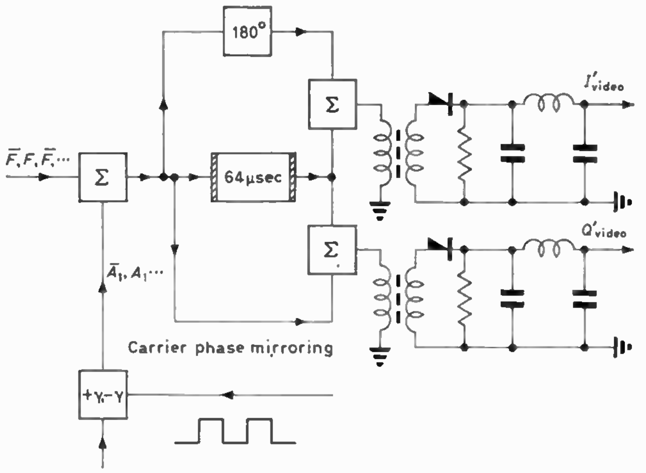


Fig. 24. PAL$_{DL}$ demodulator based on Fig. 23

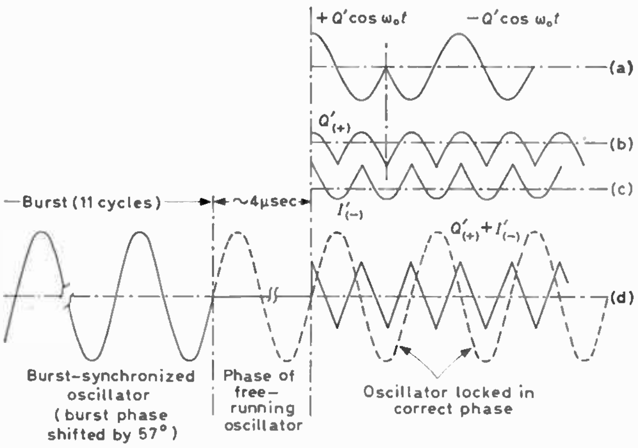


Fig. 25. Slave oscillator locking by burst signal

(a) Phase-shift due to polarity reversal of Q' to −Q'
(b) Elimination of phase-shift by full-wave rectification with pure Q' modulation
(c) As (b) for pure I' modulation with reversed diodes
(d) Signal for slave oscillator (sum of burst and signals (b) and (c))

injected reference carrier $A_1\angle\gamma$ is turned to $A_1\angle-\gamma$ during the following line interval, the resultant at the output of the adding circuit is $A_2 = 2A_1 \cdot \cos\gamma/Q'$, that is, the resultant reference is in phase with Q' as it is required for the Q' rectifier*. It is necessary to ensure, however, that A_2 is always larger than Q'_{max}. The vector diagram for the I' channel is shown in Fig. 23(b). In this case the vectors are subtracted so that $A_3 = 2A_1 \sin\gamma\angle I'$. This is the vector of the reference carrier for the I' rectifier.

If $\gamma = 45^\circ$, A_2 and A_3 become equal. In this case the reference carrier phase must be switched through $2 \times 45^\circ = 90^\circ$ on alternate lines. According to the law of image formation the phase should in general be switched through an angle of 2γ. In practice the mirroring of the reference carrier can be achieved, for instance, by the method illustrated in Figs. 14 and 15. Fig. 24 shows a block diagram of the method outlined. The fact that this method functions with a single carrier voltage and a simple phase switch could lead to considerable economies. Using the circuits described[3] for regeneration of the reference carrier from the burst signal, the electronic phase switch could be omitted if the reference carrier was injected at 45° and a special burst signal was transmitted which was advanced by 90° for alternate lines.

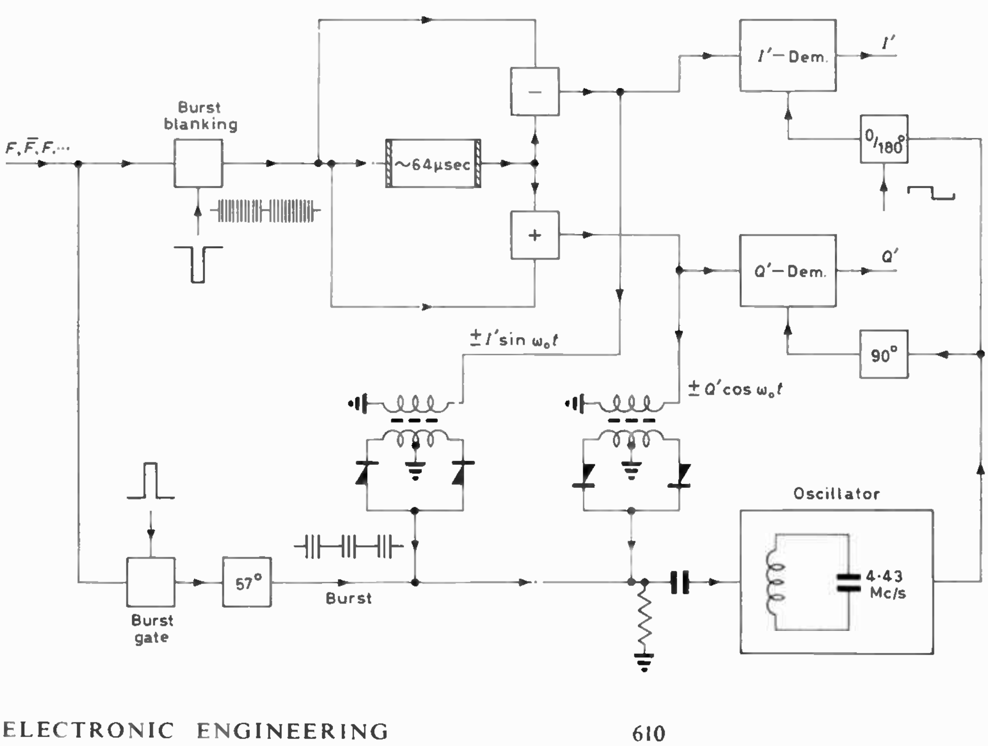
New Approaches to Subcarrier Regeneration from the Chrominance Signal

Remanent variations in colour saturation due to phase errors β can be eliminated if the reference carrier phase is made to vary synchronously with the errors. If the phase modulation $e^{j\alpha}$ of the chrominance signal could be removed from the colour carrier at the receiver, a purely amplitude modulated carrier would remain which could be used to synchronize a reference carrier oscillator in the required fashion. For the NTSC system this problem remains unsolved to date. For the PAL system with delay lines two methods are described which have proved successful in practice. The first of these methods uses the signals

$$F_Q' \text{ and } F_I'$$

separated by the delay line demodulator to synchronize the reference carrier. These two amplitude modulated signals no longer contain the chrominance phase modulation β but comprise a phase displacement $e^{j\beta}$ to which the reference carrier can be related. They are however not yet suitable for the required purpose as Q' and I' may occur after synchronous detection both as positive and negative signals. A change of sign is equivalent to a 180° phase shift of the corresponding carrier (Fig.

Fig. 26. Reference carrier generator synchronized by chrominance carrier



* *This argument is valid if the reference carrier is injected in the first or fourth quadrant, if injection takes place in the second or third quadrant reversal of the rectifier connexions will produce the correct signal phase at the output.*

25(a)) which is removed by full wave rectification. A half-wave signal of twice the colour carrier frequency results which can be used to pull a local slave oscillator into phase. There remains a phase uncertainty of π due to the double frequency signal, which is in turn removed by a starting signal derived from the burst at the beginning of each line. For instance a start-stop oscillator may be triggered by the burst and continue to oscillate under the control of the half-wave signal. Stability of this oscillator need only be sufficient to prevent it from drifting by 180° for most of the line duration. The circuits at present in use comprise a continuously running *LC* oscillator which in addition to the half-wave signal is supplied with the burst signal of an amplitude sufficiently high to ensure reliable locking-in. For this oscillator to have a sufficiently fast response, under differential phase conditions for instance, it must have a low *Q*-factor such as is given by a circuit capacitance of, say, 1000pF. To ensure reliable phase synchronization with any colour it is necessary to derive a half-wave signal from both Q' and I' and to use the sum of the resulting half-waves. For the summation to be performed one of the signals must, of course, first be delayed by 90°. The synchronizing signal amplitude which is now proportional to $(Q'+I')$ is always present, even when either Q' or I' pass through zero.

Another signal suitable for synchronization can be obtained by summation of the I' and Q' voltages with initial 90° phase shift with the diodes of the two full-wave rectifiers respectively connected in the opposite sense. Both a.c. voltage curves (Figs. 25(b) and 25(c)) are of basically the same shape as is their sum (Fig. 25(d)). The synchronizing signal supplied to the slave oscillator (Fig. 26) is the sum of the burst, phase displaced by 57°, and of $Q'_{(+)}+I'_{(-)}$ as shown in Fig. 25(d). Tests were made with an artificial sawtooth modulation of line frequency fed into the encoder with phase displacements of from 0° to a maximum of 80°. With these extreme phase displacements useless pictures were obtained with the NTSC system and with PAL circuits, employing crystal controlled regeneration, progressive loss of colour, reaching practically monochrome reproduction was observed; perfect colour images were however maintained throughout the phase error range with the new controlled oscillation circuit. In this "Improved-PAL" receiver component costs are reduced by at least the price of the quartz crystal and it may therefore be expected that the total cost, which is determined largely by the delay-line, can be reduced by almost 30 per cent.

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Electronic Machine Tool Control for Concord Project

A large installation of machine tools equipped with Ferranti numerical control is building up in the plant of Sud Aviation, Toulouse, in readiness for the production phase of the Anglo-French supersonic airliner, the Concord. A proving programme started two years ago on parts for the Caravelle and Mirage IV. These tests were so conclusive that already four Cramic routers and a Huron vertical miller have been installed. A fifth Cramic router is on the point of delivery and within the next few months a Ferranti controlled plotting machine also designed and manufactured by Cramic Engineering will follow.

Although assembly of Concords will take place at Sud Aviation and at the British Aircraft Corporation factory in Bristol, each company is responsible for the entire manufacture of certain parts of the aircraft. For Sud these are the wings and centre section of the fuselage. When Sud Aviation first considered numerical control for machining wing ribs, spars and fuselage bulkheads, two major factors influenced their decision to adopt the Ferranti magnetic tape system.

Firstly, the capital cost of the equipment is lowest, for unlike punched paper tape systems which predominate in the U.S.A., an interpolating computer is not needed on each machine. The major advantage claimed for paper tape systems is the facility for hand programming, which can eliminate dependence on computing services. But Sud considered this of dubious benefit since hand programming is uneconomic for continuous path work anyway, and the use of a computer is to be encouraged, not discouraged. It was thus reasonable to take advantage of the lower capital cost of the Ferranti system and the inherent greater reliability that must come with a simpler mechanism. The second factor stems naturally from the first. By using central computing services the latest improvements in data handling can be adopted as they become available, without affecting the machine tool control.

Translation of design data to the finished product need only take weeks, instead of months as with normal methods, since the long and costly task of designing and making jigs, fixtures and templates is eliminated. Only simple holding devices are needed on the machine.

Additionally, the number of machine hours required for a set of parts is drastically reduced. Since operating efficiency is in excess of 90 per cent, the time actually spent cutting metal rises steeply. Cases can be cited where machining time compared with conventional methods is cut by as much as 85 per cent. In one particular example from the proving tests carried out, the machining time of the finished component was reduced from 18·28h to 4·2h.

Accuracy is another factor which encourages a trend to numerical control, for much time is saved by reducing hand fitting and finishing during final assembly. Accuracy is far greater than with conventional methods. In practice, accuracy on the numerical machines is between ±0·002in in relation to the theoretical profile, but more importantly mating profiles are repeated with ±0·001in.

Design data defining the profile of wing spars and ribs is supplied to the production department in mathematical form as tables listing the co-ordinates and interpolating curve required to create the wing aerofoil. Hence information needed to machine the profile can be transferred directly to the programming sheet.

Once change points have been transferred to the programming sheet and milling information added, a paper tape is punched on a Creed teleprinter. The paper tape produced is despatched by normal airmail services to Ferranti, England, for computing and processing on to magnetic tape. This chain of communication may well be shortened in the future, however, because with the growing use of systems on the Continent, Ferranti and Sud Aviation have decided to establish a magnetic tape centre in Paris, in the Courbevoie factory of Sud Aviation.

Transmission of data in this manner is generally confined to wing components where the design originates from mathematical calculations. But many components, fuselage bulkheads for instance, are designed graphically on full size drawings and often altered artistically when the wooden mock-up of the aircraft shows need. There is no mathematical data to work from.

When the form of such a component constitutes straight lines and circles the programmer redraws the profile and, in number sequence, marks out the change points required to produce it. In conjunction with the original part drawing the programming sheet is then made up listing the co-ordinate positions and the required milling information against each change point on the X, Y and Z axes.

When the curves involved constitute irregular profiles however, it is necessary to make a full size drawing of the component on sheet alloy metal, sheet metal being used to give a stable description of the part. From this, co-ordinate points at specific intervals along the form are plotted off and the measured dimensions transferred to the programming sheet.

Experience has been building up on components for the Caravelle and the Mystere 20. To date, 90 different jobs have been programmed needing over 300 tapes. It is estimated that in Toulouse one programmer has a weekly output of about one hour's running time on the machine. By using the profile data this figure will be multiplied by about five.

Sequential Access Ferrite Store with Stepping Switch Addressing

By M. D. A. B. Rackowe*, B.A.

A stepping switch is used in conjunction with a ferrite core matrix and semiconductor circuits to form a low cost digital store. The store is of the word access type, and the one described here has a capacity of 10 words each of 10 bits.

The simplicity resulting from the use of stepping switch addressing is only achieved by sacrificing speed, and random access would also be difficult to obtain.

(Voir page 651 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 658)

A SMALL-CAPACITY sequential word access ferrite core store is described in this article. One-hundred cores are arranged in a single plane matrix to give 10 words each of 10 bits, the words being selected by a uniselector stepping switch instead of by one of the more usual semiconductor or magnetic addressing systems. The use of a stepping switch means that the store is only useful where neither high speed nor random access are required. However, a considerable saving in complexity and cost is achieved. The uniselector is driven round by a separate timing circuit and amplifier which constitutes an analogue store governing the time spent at each uniselector position.

The storage system used here is of the word access[1] type, since the inherent simplicity justifies its use in small capacity stores. Also this system is most suitable for use with the uniselector, which performs the addressing function and the overall timing of the store.

The addressing system used with a ferrite core memory has to be able to handle fast, high current drive pulses. Moreover these pulses must be able to pass in either direction along the matrix line. These conditions are difficult to meet with solid state components, and such systems are complex and costly. Where random access or high speed access are necessary there is at present no alternative. However, in some applications advantage may be taken of the simplicity and cheapness of stepping switch addressing, since it is clear that a bank of contacts can perform all the functions necessary for routing core drive pulses.

The core matrix is in a single plane, with three wires threading each core, one in the Y direction and two in the X direction (see Fig. 1). One X wire is used to pass current during the write operation, and the other is used to collect the read signal when a core is switched.

For the matrix the sequence is as follows. A full read current pulse passes down the selected Y line, and a '1' is read out from each core, which was previously in the '1' state. All the cores are then in the '0' state. A half write current of opposite sign then passes down this Y line. If it is desired to write a '1' into a particular core, a coincident half current is made to flow down the appropriate X line.

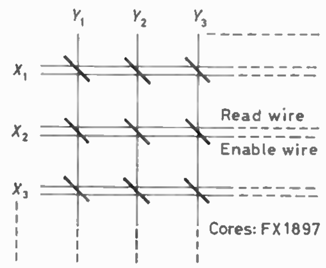

Fig. 1. Matrix arrangement

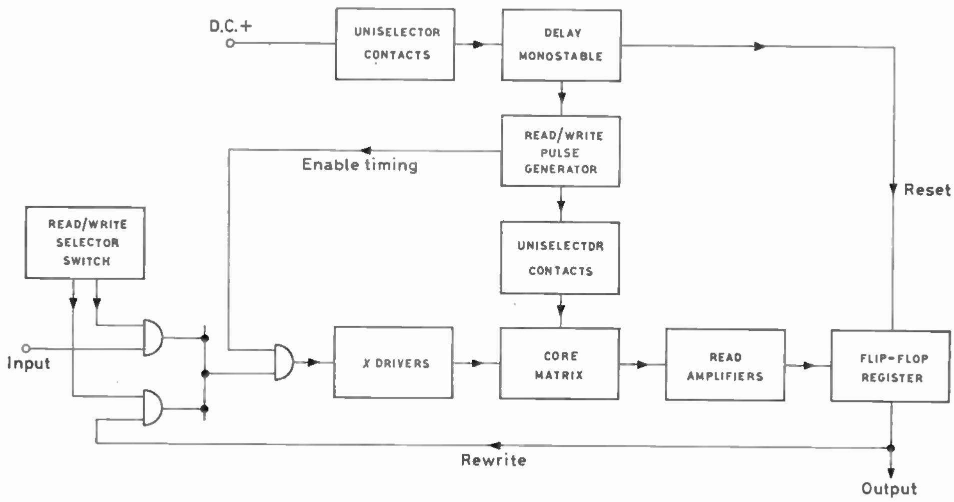

Fig. 2. General arrangement of store

Stabilized power supplies are used. These provide +12V and −12V, at a maximum current of 2A, of which 1·5A are required by the uniselector coil.

General Operation of System

Each cycle of operation of the store is initiated by the uniselector wipers moving from one position to the next. The uniselector itself is controlled by a separate timing circuit and amplifiers. Timing within each read/write cycle is achieved by using a chain of monostable delays, so no clock pulse is required.

The general arrangement of the store is shown in Fig. 2. A pulse from one bank of uniselector contacts triggers the delay monostable. At the end of the delay period a pulse from this monostable triggers the read/write pulse generator. The read/write current pulse is routed to the

* *Coutant Electronics Ltd, formerly AMF International Ltd.*

selected Y line via another bank of uniselector contacts. The word is read out via the read amplifiers and stored in a register of flip-flops. These flip-flops were all previously reset to zero by a pulse from the delay monostable occurring at the start of the delay period. The delay is only incorporated to allow the uniselector wipers to come to rest before the read/write current pulse occurs. Otherwise since a certain amount of contact bounce takes place, the current pulse is interrupted.

After the read pulse has occurred, all the cores on the selected Y line are in the '0' state. When reading from

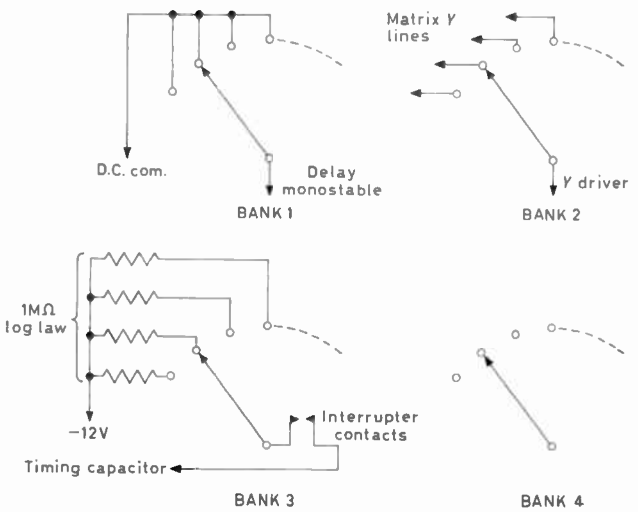

Fig. 3. Uniselector contact arrangement

the store, it is then necessary to rewrite the information into the cores. To achieve this the flip-flop outputs are fed via a series of gates to the X line drivers. One of these gates ensures that the X line pulse coincides with the Y line half write pulse. On write, information from the store input, instead of from the flip-flop register, passes through the gates to the X drivers.

Delay Monostable

The second bank of uniselector contacts in conjunction with the monostable input circuits, produces a square wave pulse each time the uniselector steps round (Figs. 3 and 4). This square wave has a fast leading edge and a slow trailing one. Differentiation produces a positive going pulse suitable for triggering the monostable.

The monostable output is differentiated and clipped, so that the leading edge gives rise to a negative going pulse. This is used to reset the register flip-flops to '0' at the start of each cycle. The trailing edge of the monostable waveform triggers the first blocking oscillator in the Y line pulse generator.

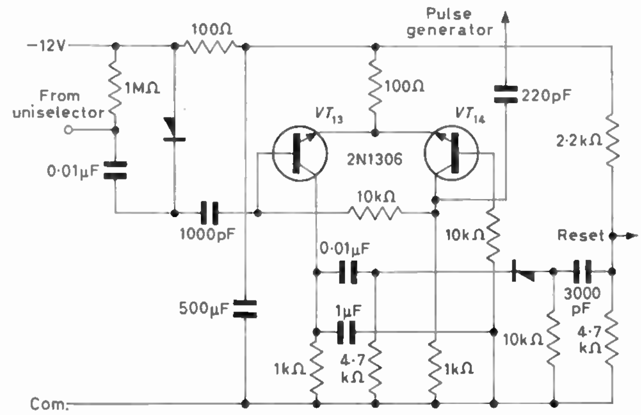

Fig. 4. Delay monostable

Y Line Pulse Generators and Drivers

The circuit is shown in Fig. 5. Transistor VT_6 and the associated circuit form the first blocking oscillator. This produces a pulse of length 8μsec, with 0·2μsec rise time. The trailing edge of this pulse triggers a second identical circuit. The two consecutive pulses so produced are taken via invertors VT_8 and VT_9 to the Y drivers VT_{10} and VT_{11}. The outputs are transformer coupled to the matrix Y line in such a way that the read and write pulses flow in opposite directions in the output winding. The turns ratio is chosen so that the write current is half the read current. All the transistors operate in the saturated mode, so dissipation is low. The voltage output from the drivers is converted to a current output by a resistor in series with the matrix line. Since the number of cores on any one Y line is small, effects on the driving voltage due to back e.m.f. of disturbed cores, wire inductance and resistance are negligible.

VT_{12} is another invertor which provides an input to the

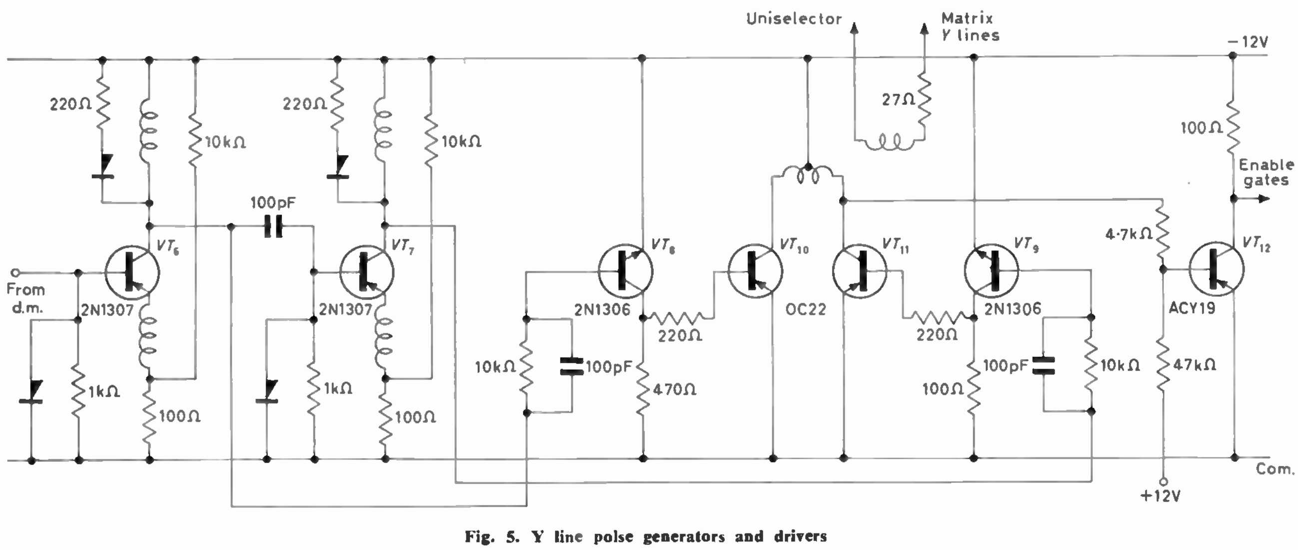

Fig. 5. Y line pulse generators and drivers

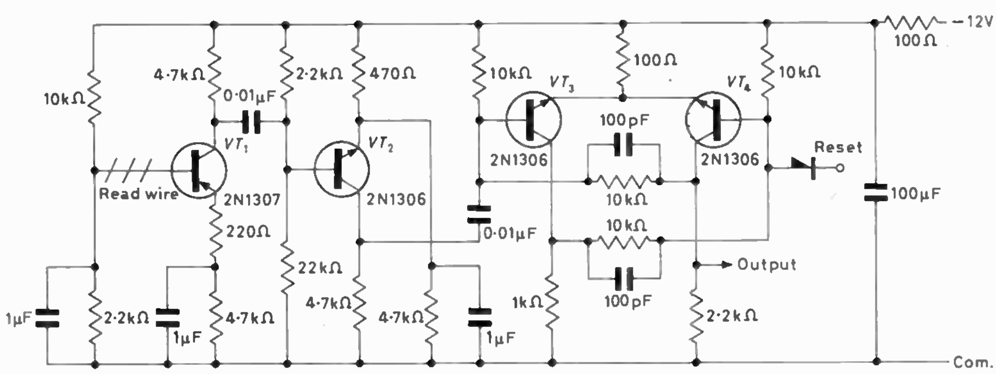

Fig. 6. Read amplifier and flip-flop

AND gate used for synchronizing the enable pulse (see Figs. 2 and 7).

Read Amplifiers and Flip-Flop Register

The output from each X line read wire is amplified, and the pulse resulting from a switched core on that line sets a flip-flop to the '1' state. The information is held in the flip-flop until a reset pulse sets it to '0'.

Referring to Fig. 6, transistor VT_1 behaves as a linear amplifier. Each read line is decoupled separately close to the amplifier input transistor to avoid the ingress of stray pulses. Potential dividers in the base and emitter circuits of VT_2 are arranged so that normally this transistor is biased off. Only negative going pulses greater than 500mV will drive this transistor on. Thus unwanted disturb signals are discriminated against and strobing of the amplifier is unnecessary.

The output from the read amplifier is taken to one input of the flip-flop. The other input is diode coupled to the source of reset pulses. The diode is back biased until a reset pulse occurs, thus avoiding coupling between the individual flip-flops in the register.

The store input and output levels are between 0 and −2V for a '0', and between −9 and −12V for a '1'.

Enable Gates and X Line Driver

The enable gates comprise three AND gates and an OR gate arranged as shown in Fig. 2. The circuit diagram for the gate and the X driver is shown in Fig. 7.

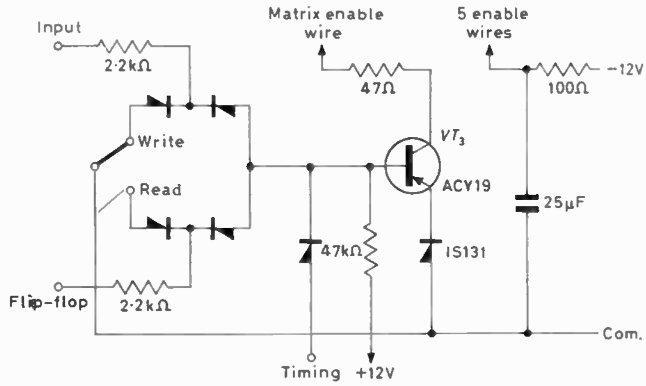

Fig. 7. Enable gates and X driver

The first two AND gates select the input or the rewrite signal according to the position of the read/write switch. The outputs are taken via the OR gate to a third AND gate which has an input from the Y line pulse generators. This gate synchronizes the pulse from the X driver which is a single transistor operating as a saturating switch. The X drive pulse is 250mA, and all the X drivers are pulsed together. This would mean a current drain of 2·5A for the pulse duration. The drivers are therefore divided into 2 groups of 5, each group being supplied from a capacitor, which is recharged at the end of the pulse.

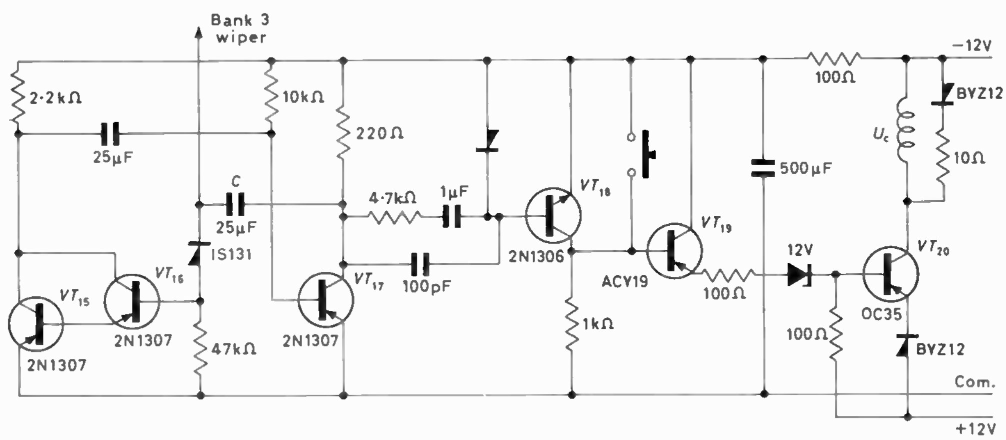

Fig. 8. Uniselector timing and drive circuit

Uniselector Timing and Drive

The uniselector is driven by pulses which are derived from the astable multivibrator consisting of transistors VT_{15}, VT_{16}, VT_{17} and associated circuits. The output is shaped and amplified by VT_{18} and VT_{19}, and applied to the power driver VT_{20}. This transistor switches the supply to the uniselector coil, which operates at 22V and 1·5A. The correct d.c. level for the power transistor drive is obtained with a Zener diode.

The period between pulses supplied to the uniselector is governed by a set of variable resistors which are selected by a bank of uniselector contacts (see Fig. 3). One resistor is used for each position, so that the time between two successive steps may be independently varied. In order that this should behave correctly the charging current to timing capacitor C (see Fig. 8) must be interrupted before the end of the pulse. Otherwise timing of the following interval will be effected by the wrong resistor. This is achieved by including the uniselector interrupter contacts in the charging circuit; these contacts open as soon as the solenoid pulls in. The duration of the pulse is made sufficient for these contacts to have opened before it finishes.

On write, the charging current to capacitor C is interrupted by contacts on the read/write switch. Stepping of the uniselector is then controlled by the push-button switch.

By the use of further banks of contacts, the uniselector cycle may be modified in various ways. Thus unused positions can be rapidly passed through, and a switch may be included which will halt the uniselector at a preselected position.

Conclusions

The limitations of this store are mostly imposed by the uniselector. However it replaces some fairly complex electronic circuits. The store cycle time is about ½sec, whereas with static addressing it would be about 25μsec.

The number of words in the store may be increased up to the number of uniselector positions, merely by increasing the number of matrix cores in the X direction. If it is required to increase the number of bits of each word, the matrix must be extended in the Y direction. Each bit requires a flip-flop for the register, a read amplifier, an X line driver and a set of enable gates. In the present unit these are built on to one plug in printed circuit board. The number of these boards required will therefore be equal to the number of bits per word.

REFERENCE

1. Quartly, C. J. Square Loop Ferrite Circuitry. (Iliffe Books Ltd., London, 1962).

A Pulse Phasemeter

By R. E. King†, Ph.D., M.Sc., M.I.E.E.E.

A simple circuit using semiconductors giving an output indication proportional to the incidence time-difference between two binary input pulse trains is described. The circuit involves a conventional bistable and a differential gate to give a ternary output. The addition of a pre-amplifier to each input channel converts the unit into an audio-frequency phasemeter.

(Voir page 651 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 658)

THE case often arises in electronic and control systems where it is necessary to give an output indication directly proportional to the magnitude and polarity, or phase, of the time difference τ between two binary pulse trains having the same period T, as shown in Fig. 1.

Where an indication of magnitude only is required (as used in certain phasemeters) this is in general no problem since the two pulse trains, denoted A and B, can be used directly to trigger a bistable circuit resulting in an output C^* if the bistable is triggered on the positive-going edges of the input waveforms. The signal C^* can then be suitably averaged to give an output proportional to τ. It is obvious that for such a circuit if the pulse train B occurs prior to A (with respect to some known time reference, i.e. $t_B < t_A$) the output signal of the bistable would be identical to the former case for $t_A > t_B$, as shown in Fig. 1. This simple arrangement is thus insensitive to the relative time of incidence of the input pulses i.e. is 'polarity' insensitive giving in effect an output indication proportional to the absolute value $|t_A - t_B| = |\tau|$.

By a relatively simple modification, described below, however, the desired 'polarity' sensitivity as depicted by the ternary output signal C in Fig. 2, can be obtained.

The 'Polarity' Sensitive Ternary Circuit

Details of the circuit are shown in Fig. 3. Two Schmitt trigger circuits using transistors VT_1VT_2, and VT_3VT_4 (if necessary) square up the input pulse trains A and B to give respectively antiphase binary outputs $A+$, $A-$, $B+$ and $B-$ at their collectors.

The operation of the circuit is explained with the aid of the waveforms of Fig. 4 where the signals are characterized by a '0' when a pnp transistor is 'off' (i.e. at approximately $-$h.t.) and by a '1' when the transistor is 'on' (i.e. at approximately zero voltage). The symbol '-1' indicates a 'reversed 1' state in common with ternary logic notation.

The base-triggered bistable (VT_5 and VT_6) is used here in an auxiliary mode to generate the pulse train D. Here the bistable is 'set' with the positive-going edge of B (actually the positive-going edge of $B+$) and 'reset' with the negative-going edge of A (actually the positive-going edge of $A-$) in readiness for the next pulse cycle. The negative-going edge of A can thus be considered as the time reference for each cycle, the maximum theoretical operating range for τ being equal to half the period i.e. $\pm T/2$.

Finally the output signal from the bistable, D, is subtracted from $A+$ in the long-tail pair differential gate/meter-drive circuit using transistors VT_7 and VT_8. The resultant differential or ternary output signal between the collectors of VT_7 and VT_8 is the desired polarity-sensitive signal being positive-going (+1) when $t_A > t_B$ (i.e. $\tau > 0$)

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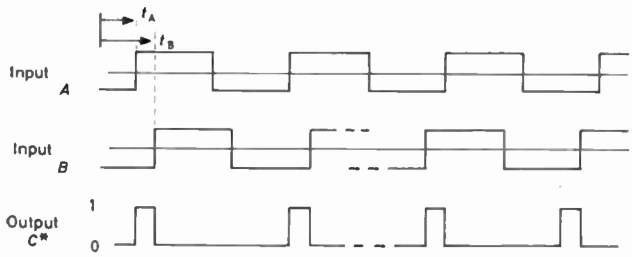

Fig. 1. Pulse trains A and B and bistable binary output C*

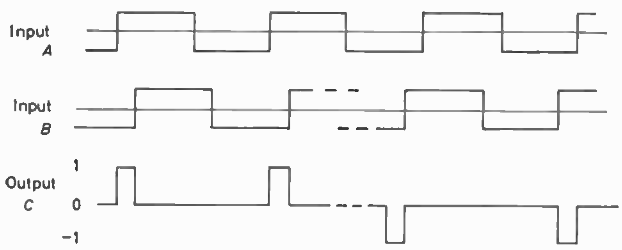

Fig. 2. Pulse trains A and B and desired ternary output C

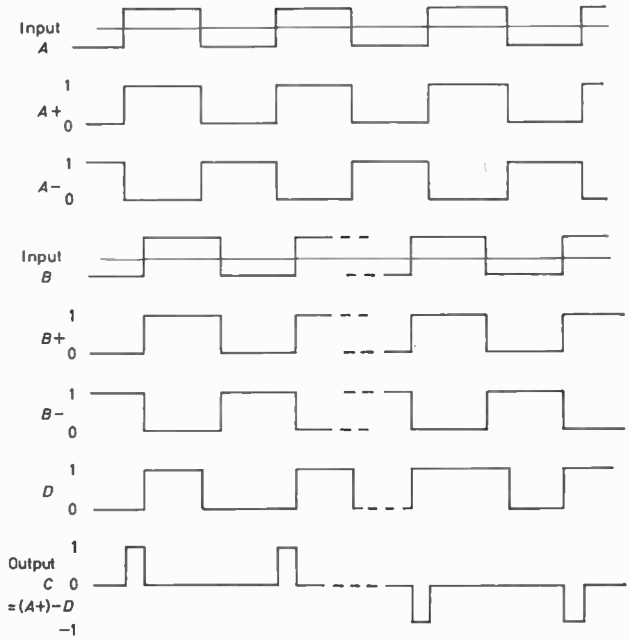

Fig. 4. Waveforms of the ternary circuit

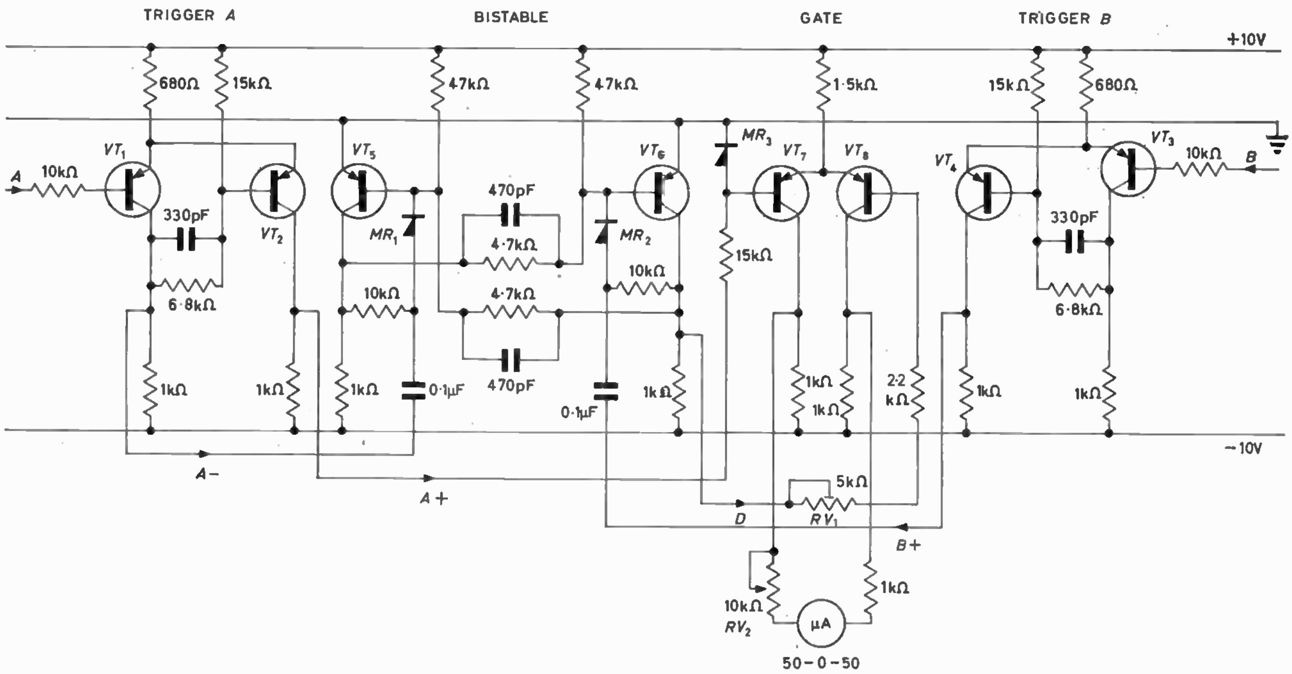

Fig. 3. The 'polarity' sensitive ternary circuit

and negative-going (-1) when $t_B > t_A$ (i.e. $\tau < 0$). The time-average of the output signal C is indicated on a centre-zero moving-coil microammeter whose sensitivity can be varied with RV_2. Diode MR_3 ensures that the base signal on VT_7 does not go positive and potentiometer RV_1 is preset to give zero output deflexion when the pulse trains A and B are coincident.

The transistors used are all 2S302's and the diodes 0A10's. The individual elements i.e. Schmitt triggers, bistable and differential gate are all of conventional design.

In cases where only impulsive inputs are available, a monostable circuit having a delay $T/2$, may be used to 'reset' the bistable.

Phasemeter Conversion

By preceding each Schmitt trigger by a high gain pre-amplifier the circuit described can be readily converted into a phasemeter for input signals of audio and sub-audio frequencies. Important requirements for such an application are that the pre-amplifiers and Schmitt trigger circuits should have no phase shift over the operating range of frequencies, that the pre-amplifier output should at all times be symmetrical even when saturating and that the Schmitt trigger circuits should have negligible backlash.

Some Transformations of the Nodal Admittance Matrix of a Network with Application to a Difference Amplifier

By K. G. Nichols*, M.Sc.

A technique is presented for establishing the admittance matrix of a k-port network, the ports being completely distinct. The result is then specialized to give the admittance parameters of a four-pole network. By a similar technique, but using a different transformation of the admittance matrix, a long tailed pair, or difference amplifier, is analysed for its out-phase and in-phase responses.

(Voir page 651 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 658)

THE nodal analysis method of solving network problems, with its attendant nodal admittance matrix, is well established[1]. Further, node suppression techniques have been presented[2,4,5] for the determination of the admittance parameters of a linear three-pole from the nodal admittance matrix of the network once the input, output and common nodes have been designated. These techniques provide powerful systematic methods of solving network problems and this article is concerned with extensions of these methods. In particular a method for determining the admittance matrix of a linear four-pole, having completely distinct input and output ports, will be presented.

In the first instance, the conversion of the nodal admittance matrix of an n-node network to give an admittance matrix for the network considered as a k-port, all ports being completely distinct, will be considered. This result will then be specialized to the case for a two-port, or four-pole network and applied to a Wheatstone net as a simple example of the technique.

Finally, using the same technique, but a different transformation of the admittance matrix some analysis of a long-tailed pair, or difference amplifier, will be carried out to obtain the ratio of the out-phase to in-phase gains of the amplifier.

The Admittance Matrix of a k-Port Network (*all ports completely distinct*)

The column vector of node currents of an n-node linear network is given in terms of that of the node potentials by the equation:

$$\mathbf{i} = \mathbf{Y}\mathbf{v} \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (1)$$

where $\mathbf{Y}$ is the nodal admittance matrix of the network.

The matrix may be either the indefinite or definite matrix relative to some particular node. In the latter case the admittance matrix and column matrices of equation (1) will be of order $(n - 1)$. Without loss of generality and for simplicity of notation, all matrices will be taken as of order n irrespective of whether they are the indefinite or definite admittance matrix. The implication in the latter case is that the network has $(n + 1)$ nodes.

Methods of determining the nodal admittance matrix for a given network have been considered elsewhere[4,5] and no further explanation will be given here.

In the given network, the nodes should be labelled such that the first port is between the 1st and $(k + 1)$th nodes, the second port between the 2nd and $(k + 2)$th nodes, and so on, there being k-ports in all. This is illustrated in Fig. 1. Such an arrangement is possible, since by hypothesis all the k-ports are distinct.

The voltage across the rth port is denoted by:

$$u_r = v_r - v_{r+k}\,; \quad r = 1, 2, \ldots, k \quad \ldots\ldots\ldots \quad (2)$$

Further the current flowing out of the lower terminal

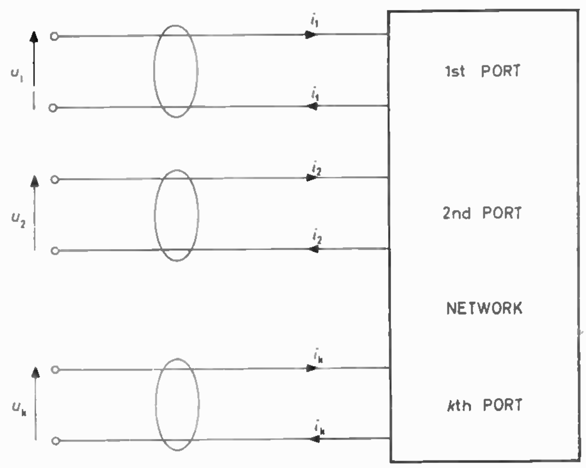

Fig. 1. The n-node network with k-ports selected. The remaining $(n - 2k)$ nodes are not identified in this figure

of the rth port is equal to the current flowing into the upper terminal of the port, i.e.:

$$i_{r+k} = -i_r \quad ; \quad r = 1, 2, \ldots, k \quad \ldots\ldots\ldots \quad (3)$$

Thus the matrix equation (1) may be written as:

$$\left[\begin{array}{c} \mathbf{i}_1 \\ \hline -\mathbf{i}_1 \\ \hline \mathbf{o} \end{array}\right] = \left[\begin{array}{c|c|c} \mathbf{Y}_{11} & \mathbf{Y}_{12} & \mathbf{Y}_{13} \\ \hline \mathbf{Y}_{21} & \mathbf{Y}_{22} & \mathbf{Y}_{23} \\ \hline \mathbf{Y}_{31} & \mathbf{Y}_{32} & \mathbf{Y}_{33} \end{array}\right] \left[\begin{array}{c} \mathbf{v}_1 \\ \hline \mathbf{v}_2 \\ \hline \mathbf{v}_3 \end{array}\right] \quad \ldots\ldots\ldots \quad (4)$$

where the matrices have been partitioned after the first k and $2k$ rows and, where applicable, columns. Here $\mathbf{i}_1$ is the column vector of currents flowing into the first k nodes, $\mathbf{v}_1$ is the column vector of potentials of the first k nodes, $\mathbf{v}_2$ that for the potentials of the next k nodes, and $\mathbf{v}_3$ that of the remaining nodes.

If the current column vector of equation (4) is pre-multiplied by the transformation matrix:

$$\mathbf{W} = \left[\begin{array}{c|c|c} \mathbf{I}_k & \mathbf{0} & \mathbf{0} \\ \hline \mathbf{I}_k & \mathbf{I}_k & \mathbf{0} \\ \hline \mathbf{0} & \mathbf{0} & \mathbf{I}_{n-2k} \end{array}\right] \quad \ldots\ldots\ldots \quad (5)$$

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where $\mathbf{0}$ is the zero matrix of appropriate order and $\mathbf{I}_r$ is the unit matrix of order r, the resultant column vector has $\mathbf{i}_1$ for its first k entries and zero for all other entries.

The transposed inverse of $\mathbf{W}$ is readily found to be:

$$(\mathbf{W}^T)^{-1} = \left[\begin{array}{c|c|c} \mathbf{I}_k & -\mathbf{I}_k & \mathbf{0} \\ \hline \mathbf{0} & \mathbf{I}_k & \mathbf{0} \\ \hline \mathbf{0} & \mathbf{0} & \mathbf{I}_{n-2k} \end{array}\right] \quad \ldots\ldots\ldots\ldots (6)$$

and if the voltage column vector of equation (4) is pre-multiplied by this matrix, $\mathbf{v}_1$ in this vector is replaced by

$$\mathbf{u} = \mathbf{v}_1 - \mathbf{v}_2.$$

Using these transformations equation (4) may be written:

$$\left[\begin{array}{c} \mathbf{i}_1 \\ \hline \mathbf{0} \\ \hline \mathbf{0} \end{array}\right] = \mathbf{WYW}^T \left[\begin{array}{c} \mathbf{u} \\ \hline \mathbf{v}_2 \\ \hline \mathbf{v}_3 \end{array}\right] \quad \ldots\ldots\ldots\ldots (7)$$

The partitioned form of the matrices $\mathbf{W}$, $\mathbf{Y}$ and $\mathbf{W}^T$ enable the product $\mathbf{WYW}^T$ to be readily evaluated, the result being given in equation (8).

$$\mathbf{WYW}^T = \left[\begin{array}{c|c|c} \mathbf{Y}_{11} & \mathbf{Y}_{11} + \mathbf{Y}_{12} & \mathbf{Y}_{13} \\ \hline \mathbf{Y}_{11} + \mathbf{Y}_{21} & \mathbf{Y}_{11} + \mathbf{Y}_{12} + \mathbf{Y}_{21} + \mathbf{Y}_{22} & \mathbf{Y}_{13} + \mathbf{Y}_{23} \\ \hline \mathbf{Y}_{31} & \mathbf{Y}_{31} + \mathbf{Y}_{32} & \mathbf{Y}_{33} \end{array}\right] \quad \ldots\ldots (8)$$

This matrix is thus formed from the original matrix of equation (1) by adding the first column to the $(k+1)^{th}$ column, the 2^{nd} column to the $(k+2)^{th}$ column and so on finally adding the k^{th} column to the $2k^{th}$ column; and then repeating the procedure with the rows.

The next step in the reduction process is to express $\mathbf{i}_1$ linearly in terms of $\mathbf{u}$. A method of achieving this reduction, known as node suppression, has been described elsewhere[3,4,5] and will be utilized here. The partitions after $2k$ rows and columns in the matrix of equation (8) are dropped, but those after k rows and columns retained. The partitioned matrix now has the form:

$$\mathbf{WYW}^T = \left[\begin{array}{c|c} \mathbf{A} & \mathbf{B} \\ \hline \mathbf{D} & \mathbf{C} \end{array}\right]$$

The matrix equation (7) is then multiplied out and breaks down into equations (9) and (10) as follows:

$$\mathbf{i}_1 = \mathbf{Au} + \mathbf{B}\left[\begin{array}{c} \mathbf{v}_2 \\ \hline \mathbf{v}_3 \end{array}\right] \quad \ldots\ldots\ldots\ldots (9)$$

and

$$\mathbf{0} = \mathbf{Du} + \mathbf{C}\left[\begin{array}{c} \mathbf{v}_2 \\ \hline \mathbf{v}_3 \end{array}\right] \quad \ldots\ldots\ldots\ldots (10)$$

If $|\mathbf{C}| \neq 0$, then $\mathbf{C}^{-1}$ exists and equation (10) gives:

$$\left[\begin{array}{c} \mathbf{v}_2 \\ \hline \mathbf{v}_3 \end{array}\right] = \mathbf{C}^{-1}\mathbf{D}\,\mathbf{u}$$

Substituting for $\left[\begin{array}{c} \mathbf{v}_2 \\ \hline \mathbf{v}_3 \end{array}\right]$ into equation (9) gives:

$$\mathbf{i}_1 = [\mathbf{A} - \mathbf{BC}^{-1}\mathbf{D}]\,\mathbf{u}$$

The matrix $[\mathbf{A} - \mathbf{BC}^{-1}\mathbf{D}]$ is the admittance matrix of the network, considered as a k-port, and relates the currents flowing into the ports to the potentials across the ports.

The Admittance Matrix of a Four-Pole (*with completely distinct ports*)

In the special case of a two-port, or four-pole, network, one node may be taken as the reference node. The arrangement is shown in Fig. 2.

In this case $\mathbf{i}_1 = \begin{bmatrix} i_1 \\ i_2 \end{bmatrix}$ and $\mathbf{u} = \begin{bmatrix} v_1 - v_3 \\ v_2 \end{bmatrix} = \begin{bmatrix} v_{in} \\ v_{out} \end{bmatrix}$

The matrix $[\mathbf{A} - \mathbf{BC}^{-1}\mathbf{D}]$ now gives the two by two matrix of the admittance parameters of the four-pole.

As a simple illustration the Wheatstone net of Fig. 3(a) will be considered. This is a two-port network, and if desired, may be redrawn as the unbolted lattice network shown in Fig. 3(b).

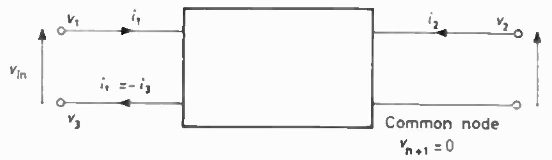

Fig. 2. Two-port network, terminal $(n+1)$ as reference node

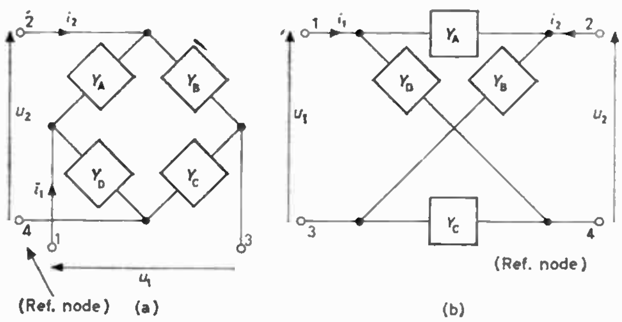

Fig. 3(a). Wheatstone net
(b). Unbolted lattice

By inspection, the definite admittance matrix, relative to node 4, of the network is:

$$\mathbf{Y} = \begin{bmatrix} Y_A + Y_D, & -Y_A, & 0 \\ -Y_A, & Y_A + Y_B, & -Y_B \\ 0, & -Y_B, & Y_B + Y_C \end{bmatrix} \quad \ldots\ldots\ldots\ldots (12)$$

Adding the first row to the third row and then adding the first column to the third column, the following matrix is obtained:

$$\left[\begin{array}{cc|c} Y_A + Y_D, & -Y_A & Y_A + Y_D \\ -Y_A, & Y_A + Y_B & -(Y_A + Y_B) \\ \hline Y_A + Y_D, & -(Y_A + Y_B) & Y_A + Y_B + Y_C + Y_D \end{array}\right]$$

This latter matrix is shown partitioned ready for reduction. Carrying out the reduction, the admittance matrix of the two-port, or four-pole, becomes:

$$\begin{bmatrix} Y_A + Y_D, & -Y_A \\ -Y_A, & Y_A + Y_B \end{bmatrix} - \frac{1}{Y_A + Y_B + Y_C + Y_D}\begin{bmatrix} Y_A + Y_B \\ -(Y_A + Y_B) \end{bmatrix}[Y_A + Y_D, \; -(Y_A + Y_B)]$$

$$= \frac{1}{Y_A + Y_B + Y_C + Y_D}\begin{bmatrix} (Y_A + Y_D)(Y_B + Y_C), & Y_B Y_D - Y_A Y_C \\ Y_B Y_D - Y_A Y_C, & (Y_A + Y_B)(Y_C + Y_D) \end{bmatrix} \quad \ldots\ldots\ldots (13)$$

In particular, from equation (13), it is apparent that zero transmission occurs if:

$$Y_A Y_C = Y_B Y_D \qquad (14)$$

the well known balance condition.

The Long-Tailed Pair or Difference Amplifier

As an illustration of a slightly different transformation of the admittance matrix, a single stage difference amplifier will be considered. The basic circuit of the amplifier is shown in Fig. 4, in which the networks N and N' are supposed to represent similar but not necessarily identical, active devices and which have admittance parameters, relative to node 5, given by the matrices:

$$\begin{bmatrix} \gamma_i , \gamma_r \\ \gamma_f , \gamma_o \end{bmatrix} \text{ and } \begin{bmatrix} \gamma_i' , \gamma_r' \\ \gamma_f' , \gamma_o' \end{bmatrix} \text{ respectively.}$$

The active devices may be thermionic valves, transistors or any other active linear circuit elements.

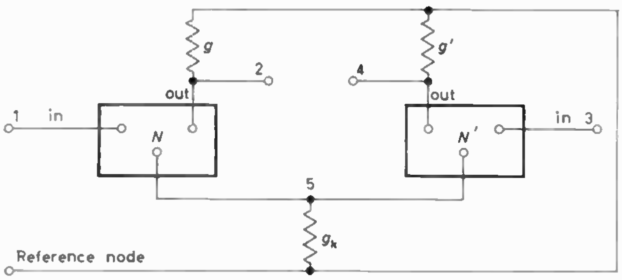

Fig. 4. The equivalent circuit of a difference amplifier

In Fig. 4 the loads g, g' and g_k are assumed to be purely resistive although this restriction is not essential. The definite admittance matrix, relative to the reference node, of the network may be written down by inspection of the figure and is given by:

$$\mathbf{Y} = \begin{bmatrix} \gamma_i , & \gamma_r , & 0 , & 0 , & -(\gamma_i+\gamma_r) \\ \gamma_f , & \gamma_o+g , & 0 , & 0 , & -(\gamma_f+\gamma_o) \\ 0 , & 0 , & \gamma_i' , & \gamma_r' , & -(\gamma_i'+\gamma_r') \\ 0 , & 0 , & \gamma_f' , & \gamma_o'+g' , & -(\gamma_f'+\gamma_o') \\ -(\gamma_i+\gamma_f), & -(\gamma_r+\gamma_o), & -(\gamma_i'+\gamma_f'), & -(\gamma_r'+\gamma_o'), & g_k+2\sigma'' \end{bmatrix} \qquad (15)$$

and so:

$$\mathbf{i} = \mathbf{Y}\mathbf{v} \qquad (16)$$

with i_5 equal to zero.

In equation (15):

$$\sigma = \gamma_i+\gamma_r+\gamma_f+\gamma_o$$

$$\sigma' = \gamma_i'+\gamma_r'+\gamma_f'+\gamma_o'$$

and

$$\sigma'' = \tfrac{1}{2}(\sigma + \sigma') \qquad (17)$$

In all subsequent equations a double primed symbol is to imply averaging of the quantity over unprimed and single primed values of the quantity. Similarly a symbol prefixed by a δ is to imply half the difference between the unprimed and primed values of the quantity, for example:

$$\delta g = \tfrac{1}{2}(g - g') \qquad (18)$$

In the mode of operation to be considered, input signals are applied to nodes 1 and 3 and the output signal is taken between nodes 2 and 4. Thus the circuit of Fig. 4 is strictly a three-port network, the ports being: (1) from the reference node to node 1, (2) from the reference node to node 3, and (3) from node 4 to node 2. The admittance matrix of the three-port can be obtained by using the results of the previous sections of this article. That is the second row of the matrix of equation (15) is added to the fourth row and the second column is added to the fourth column. Following the partitioning off of the last two rows and columns, the reduction to the third order admittance matrix may be carried out.

Instead of following this procedure, an alternative transformation of the matrix will be employed leading to more interesting results associated with the out-phase and in-phase amplifying properties of the circuit.

The two properties which will be investigated are: firstly, the output for out-phase inputs while the in-phase components of the inputs are held at zero, and secondly, the output for in-phase inputs while the out-phase components of the inputs are held at zero. More explicitly:

Firstly, (v_2-v_4) as a function of (v_1-v_3), with $(v_1+v_3)=0$, and secondly,

(v_2-v_4) as a function of (v_1+v_3), with $(v_1-v_3)=0$.

To this end transformation matrices, **W** and **T**, are now sought such that:

$$\mathbf{W} \begin{bmatrix} i_1 \\ i_2 \\ i_3 \\ -i_2 \\ 0 \end{bmatrix} = \begin{bmatrix} i_1 \\ i_2 \\ i_3 \\ 0 \\ 0 \end{bmatrix} \qquad (19)$$

and:

$$\mathbf{T} \begin{bmatrix} v_1-v_3 \\ v_2-v_4 \\ v_1+v_3 \\ v_4 \\ v_5 \end{bmatrix} = \begin{bmatrix} v_1 \\ v_2 \\ v_3 \\ v_4 \\ v_5 \end{bmatrix} \qquad (20)$$

It is readily seen that the required transformation matrices are:

$$\mathbf{W} = \begin{bmatrix} 1,0,0,0,0 \\ 0,1,0,0,0 \\ 0,0,1,0,0 \\ 0,1,0,1,0 \\ 0,0,0,0,1 \end{bmatrix} \qquad (21)$$

and

$$\mathbf{T} = \begin{bmatrix} \tfrac{1}{2},0,\tfrac{1}{2},0,0 \\ 0,1,0,1,0 \\ -\tfrac{1}{2},0,\tfrac{1}{2},0,0 \\ 0,0,0,1,0 \\ 0,0,0,0,1 \end{bmatrix} \qquad (22)$$

Substitution into equation (16) yields:

$$\begin{bmatrix} i_1 \\ i_2 \\ i_3 \\ 0 \\ 0 \end{bmatrix} = \mathbf{WYT} \begin{bmatrix} v_1-v_3 \\ v_2-v_4 \\ v_1+v_3 \\ v_4 \\ v_5 \end{bmatrix} \qquad (23)$$

where the matrix **WYT** is formed by multiplication of the matrices of equations (15), (21) and (22) giving:

WYT =

$$\begin{bmatrix} \frac{1}{2}\gamma_i & , & \gamma_r & , & \frac{1}{2}\gamma_i & , & \gamma_r & , & -(\gamma_i+\gamma_r) \\ \frac{1}{2}\gamma_f & , & \gamma_o+g & , & \frac{1}{2}\gamma_f & , & \gamma_o+g & , & -(\gamma_f+\gamma_o) \\ -\frac{1}{2}\gamma_i' & , & 0 & , & \frac{1}{2}\gamma_i' & , & \gamma_r' & , & -(\gamma_i'+\gamma_r') \\ \delta\gamma_f & , & \gamma_o+g & , & \gamma_f'' & , & 2(\gamma_o''+g'') & , & -2(\gamma_f''+\gamma_o'') \\ -(\delta\gamma_i+\delta\gamma_f), & & -(\gamma_r+\gamma_o), & & -(\gamma_i''+\gamma_f''), & & -2(\gamma_r''+\gamma_o''), & & g_k+2\sigma'' \end{bmatrix} \qquad (24)$$

When considering the out-phase response of the network, the in-phase input signal is kept at zero; that is: $v_1+v_3=0$.

Thus in multiplying out the matrices on the right-hand side of equation (23) the elements of the third column of the matrix of equation (24) are always multiplied by zero; hence the third column of this matrix may be deleted. Further the current i_3 is of little significance and so the third row of the matrix may also be deleted. Equation (23) then reduces to:

$$\begin{bmatrix} i_1 \\ i_2 \\ \hline 0 \\ 0 \end{bmatrix} = \left[\begin{array}{cc|cc} \frac{1}{2}\gamma_i\ , & \gamma_r & \gamma_r\ , & -(\gamma_i+\gamma_r) \\ \frac{1}{2}\gamma_f\ , & \gamma_o+g & \gamma_o+g\ , & -(\gamma_f+\gamma_o) \\ \hline \delta\gamma_f\ , & \gamma_o+g & 2(\gamma_o''+g''), & -2(\gamma_f''+\gamma_o'') \\ -(\delta\gamma_i+\delta\gamma_f), & -(\gamma_r+\gamma_o) & -2(\gamma_r''+\gamma_o''), & g_k+2\sigma'' \end{array}\right] \begin{bmatrix} v_1-v_3 \\ v_2-v_4 \\ \hline v_4 \\ v_5 \end{bmatrix}$$

with $v_1+v_3=0$. $\qquad$ (25)

This matrix is shown partitioned ready for reduction as described in earlier sections.

When considering the in-phase response of the network, the out-phase input signal is kept at zero; that is: $v_1-v_3=0$. This suggests that the first row and column of the matrix of equation (24) should be deleted, followed by interchange of the first and second rows and first and second columns of the resultant matrix in order to obtain an equation similar in form to equation (25). However if this is done, the current i_3 tops the current column vector on the left-hand side of equation (25) and as a result a certain degree of symmetry is lost. This can be avoided by pre-multiplying both sides of equation (23) by the transformation (permutation) matrix:

$$\mathbf{V} = \begin{bmatrix} 0,0,1,0,0 \\ 0,1,0,0,0 \\ 1,0,0,0,0 \\ 0,0,0,1,0 \\ 0,0,0,0,1 \end{bmatrix} \qquad (26)$$

This transformation interchanges i_1 and i_3 on the left-hand side of equation (23) and interchanges the first and third rows of the matrix of equation (24). Subsequently the first row and third column of this matrix are deleted, giving the equation:

$$\begin{bmatrix} i_1 \\ i_2 \\ \hline 0 \\ 0 \end{bmatrix} = \left[\begin{array}{cc|cc} \frac{1}{2}\gamma_i\ , & \gamma_r & \gamma_r\ , & -(\gamma_i+\gamma_r) \\ \frac{1}{2}\gamma_f\ , & \gamma_o+g & \gamma_o+g\ , & -(\gamma_f+\gamma_o) \\ \hline \gamma_f''\ , & \gamma_o+g & 2(\gamma_o''+g''), & -2(\gamma_f''+\gamma_o'') \\ -(\gamma_i''+\gamma_f''), & -(\gamma_r+\gamma_o) & -2(\gamma_r''+\gamma_o''), & g_k+2\sigma'' \end{array}\right] \begin{bmatrix} v_1+v_3 \\ v_2-v_4 \\ \hline v_4 \\ v_5 \end{bmatrix}$$

with $v_1-v_3=0$ $\qquad$ (27)

Again this matrix is shown partitioned ready for reduction.

It is worth pointing out that the admittance matrices of equations (25) and (27) differ in but two entries.

Although the reduction of the matrices of equations (25) and (27) to order two by two has been carried out in some detail, it is not thought worth while to give full details of the algebraic manipulation here but rather just to give some of the more significant results.

The Symmetrical Case

In the special case when the parameters of the two active devices and their loads have identical numerical values, the resultant equations have a simple form. In this case, primed and unprimed quantities become equal and all the incremental quantities are zero.

After reduction equation (25) becomes:

$$\begin{bmatrix} i_1 \\ i_2 \end{bmatrix} = \frac{1}{2}\begin{bmatrix} \gamma_i\ , & \gamma_r \\ \gamma_f\ , & \gamma_o+g \end{bmatrix}\begin{bmatrix} v_1-v_3 \\ v_2-v_4 \end{bmatrix} \qquad (28)$$

with $v_1+v_3=0$.

That is the effective admittance matrix of the amplifier for out-phase input signals is that of either active alone and is independent of g_k.

On the other hand, equation (27) reduces to:

$$\begin{bmatrix} i_1 \\ i_2 \end{bmatrix} = \frac{1}{2}\begin{bmatrix} \Gamma_i, & \gamma_r \\ 0\ , & \gamma_o+g \end{bmatrix}\begin{bmatrix} v_1+v_3 \\ v_2-v_4 \end{bmatrix} \qquad (29)$$

with $v_1-v_3=0$

and where Γ_i is a rather complicated expression in the circuit parameters. The significant point is the zero entry in the admittance matrix of this last equation indicating zero transmission of the in-phase signal.

The Asymmetric Case

When the parameters of the two active devices and their loads do not have identical values, but provided the degree of asymmetry is not too great, equation (25) giving the out-phase response reduces to:

$$\begin{bmatrix} i_1 \\ i_2 \end{bmatrix} = \frac{1}{2}\begin{bmatrix} \gamma_i''\ , & \gamma_r'' \\ \gamma_f''\ , & \gamma_o''+g'' \end{bmatrix}\begin{bmatrix} v_1-v_3 \\ v_2-v_4 \end{bmatrix} \qquad (30)$$

with $v_1+v_3=0$

a result which might be expected from the form of equation (28).

Equation (27) giving the in-phase response reduces, under the same assumptions, to:

$$\begin{bmatrix} i_1 \\ i_2 \end{bmatrix} = \frac{1}{2}\begin{bmatrix} \Gamma_i''\ , & \gamma_r'' \\ \Delta\Gamma_f'', & \gamma_o''+g'' \end{bmatrix}\begin{bmatrix} v_1+v_3 \\ v_2-v_4 \end{bmatrix} \qquad (31)$$

with $v_1-v_3=0$

where for small asymmetries Γ_i'' is approximately equal to Γ_i of equation (28), and where the somewhat more significant

$\Delta\Gamma_f''$ is given by:

$$\Delta\Gamma_f'' = \begin{Bmatrix} 2\,[2g''(\gamma_r''+\gamma_o'')+g_k\,(\gamma_o''+g'')]\,\delta\gamma_f/\Delta \\ -\,[2g''\,(\gamma_i''+\gamma_f'')+g_k\gamma_f'']\,\delta\gamma_o \\ +\,[2\gamma''-g_k\gamma_i'']\,\delta g \end{Bmatrix} \quad \ldots\ldots\ldots\ldots (32)$$

with:

$$\Delta = 2\,\{(\gamma_o''+g'')\,g_k + 2\sigma''g''+2\gamma''\} \quad \ldots\ldots (33)$$

and:

$$\gamma'' = \gamma_i''\gamma_o'' - \gamma_f''\gamma_r'' \quad \ldots\ldots\ldots\ldots\ldots (34)$$

If the input voltages to the difference amplifier are derived from zero impedance sources, equation (30) gives the out-phase voltage gain as:

$$\left(\frac{v_2 - v_4}{v_1 - v_3}\right) = -\,\frac{\gamma_f''}{\gamma_o'' + g''} \quad \ldots\ldots\ldots (35)$$

with $(v_1 + v_3) = 0$

and equation (31) gives the in-phase voltage gain as:

$$\left(\frac{v_2 - v_4}{v_1 + v_3}\right) = -\,\frac{\Delta\Gamma_f''}{\gamma_o'' + g''} \quad \ldots\ldots\ldots (36)$$

with $(v_1 - v_3) = 0$

In equations (35) and (36) any load admittance across terminals 2 and 4 must be added to $(\gamma_o'' + g'')$.

The rejection ratio of the out-phase to in-phase voltage gains is then:

$$r = (\gamma_f''/\Delta\Gamma_f'') \quad \ldots\ldots\ldots\ldots\ldots (37)$$

As an example, suppose the active devices in the network are transistors in common emitter configuration and that the circuit is symmetric except that $\delta\gamma_f \neq 0$. Equations (32), (33) and (37) then give:

$$r = \left\{\frac{(\gamma_o''+g'')\,g_k+2\sigma''g''+2\gamma''}{(\gamma_o''+g'')g_k+2(\gamma_r''+\gamma_o'')g''}\right\}(\gamma_f''/\delta\gamma_f) \quad \ldots\ldots (38)$$

Considering two extremes:

(1) g_k very small, that is a large common emitter resistor,

$$r \simeq \left\{\frac{\sigma''g'' + \gamma''}{(\gamma_r'' + \gamma_o'')\,g''}\right\}(\gamma_f''/\delta\gamma_f)$$

and assuming

$$|\gamma''| \ll \sigma''g'':\; r \simeq \left(\frac{\gamma_i'' + \gamma_f''}{\gamma_r'' + \gamma_o''} + 1\right)(\gamma_f''/\delta\gamma_f)$$

Since

$\left|\dfrac{\gamma_i''+\gamma_f''}{\gamma_r''+\gamma_o''}\right| \gg 1$, the rejection ratio is very large; that is there is virtually no response to in-phase input signals.

(2) g_k very large, that is a small common emitter resistor,

$$r \simeq (\gamma_f''/\delta\gamma_f)$$

In this case the response to in-phase input signals is as if a single transistor with forward transfer admittance parameter $\delta\gamma_f$ were used to amplify the in-phase signal.

The Difference Amplifier with Non-Zero Input Source Impedances

The rather special example, considered in the last section, of zero input source impedances may prove too restrictive for the analysis of certain transistor difference amplifiers. When these source impedances are not zero, one method of analysing the network is to introduce additional nodes at the junction of each source impedance with an input to the amplifier. The admittance matrix of the network, given by equation (15), is therefore increased in order to a 7 by 7 matrix but, apart from additional complexity of expression, the method of analysis is identical to that already presented; the two extra nodes being considered externally unconnected, similar to node 5.

An alternative technique is to replace the input voltage sources and their internal impedances by current generators shunted by internal admittances. The source admittances may then be included in the 5 by 5 admittance matrix of the network. The currents i_1 and i_2 now take the form $v_{s1}Y_{s1}$ and $v_{s2}Y_{s2}$ respectively, where v_{s1} and v_{s2} are the source voltages and Y_{s1} and Y_{s2} are the internal series admittances of the voltage sources.

Subsequently transformations are sought which convert the entries $v_{s1}Y_{s1}$ and $v_{s2}Y_{s2}$ in the current column vector into $\delta v_s = \frac{1}{2}(v_{s1} - v_{s2})$ and $v_s'' = \frac{1}{2}(v_{s1} + v_{s2})$ respectively as required for out-phase and in-phase analyses.

Acknowledgment

The author is indebted to Mr. G. G. Bloodworth for a number of useful discussions concerning aspects of this work.

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Tidal Flow Traffic Signalling

The second tidal-flow scheme to help relieve London's rush-hour congestion came into operation on 30 July 1964. Under the new plan, both lanes of Hammersmith Bridge are allocated to southbound traffic at about three minute intervals during the evening rush period. During the intervening periods of two-way working, when only one southbound lane is available, each of the two south going lanes in the approach from Hammersmith Broadway is given alternate access to the bridge.

The signalling system for traffic control has been designed and installed by Automatic Telephone & Electric Co. Ltd, a principal operating Company in the Plessey Group, in collaboration with the Ministry of Transport.

At the start of the rush-hour, additional traffic signals on electrically-powered trolleys are moved into the middle of the road in the north and south approaches. On the north side of the bridge, southbound traffic is formed into two lanes by two of these movable signals. Both lanes receive a green signal when the bridge is being used entirely for southgoing traffic. During two-way working, these supplementary signals are also used to give alternate access to the bridge to each southbound lane of traffic from Hammersmith Broadway.

Lane control signals for the bridge are displayed on an overhead gantry mounted over the northern approach. The normal southbound lane is indicated by a permanently illuminated white arrow. Another white arrow over the 'offside' lane is switched on when this is available for southbound traffic: when northbound traffic is using the bridge a red 'X' signal appears to warn southbound drivers not to use this lane.

First stage in switching over to full southbound flow after the movable signals are in position is to halt northbound traffic. To do this, the traffic signals on the south side are set to red by a police constable who uses a closed circuit television monitor to check when northbound traffic has cleared the bridge, and both lanes of southbound traffic then receive a green signal.

After a fixed period, which at first will be about one minute, two-way working is then initiated for about two minutes to allow the queue of northbound traffic to disperse. The closed circuit television monitor is again used to check that one lane is clear for northbound traffic which is then released. At intervals during the two-way period, the other southbound lane has sole use of the single lane on the bridge.

A novel feature of the system is the use of buried inductive loop detectors for indicating stationary traffic in the northbound approach.

Two Transistor-Operated Frequency-Selective Amplifiers

By S. Harkness*

This article describes two self-contained transistor-operated frequency-selective amplifiers. Both use a variation of the Wien bridge described by Wigan which gives control of frequency by means of a single variable resistor, so avoiding the difficulties of accurate tracking when a number of resistors have to be varied simultaneously. The amplifiers are particularly suitable for use as detectors in a.c. bridges. One is intended for use in iron-loss testing, where high sensitivity is not required but where, because very large odd-order harmonics may be present, very high selectivity is essential. The frequency coverage is 20 to 600c/s in three ranges. The output is displayed on a moving-coil meter, on which 2 per cent of full-scale deflexion corresponds to 5μV. The other design is a general-purpose detector for linear audio frequency bridges of the highest precision. The emphasis here is on high sensitivity, full-scale deflexion of the output meter corresponding to a signal of about ½μV. The frequency coverage is 20c/s to 20kc/s in 6 ranges.

(Voir page 651 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 658)

VALVE-OPERATED amplifiers, using frequency-selective networks in the feedback path, have been used for many years at the NPL[1] and elsewhere in a number of branches of electrical measurements. An increased versatility is evidently afforded by compact battery-operated amplifiers using transistors, not merely on grounds of portability but particularly as a result of their reduced earth admittances and freedom from hum.

A major difficulty in the use of frequency-dependent null networks—such as the parallel-*T*, in which the frequency is varied by simultaneously varying a number of resistors—is that of obtaining sufficiently good tracking; a difficulty which increases as the design selectivity is increased.

The amplifiers to be described use a variation of the Wien bridge described by Wigan[2] which gives control of frequency by means of a single element. Since this work was begun a number of other circuits giving single-element control have also become known[3].

General Requirements

The desirable properties of a general-purpose selective amplifier include high selectivity, low noise, low intermodulation-distortion and a wide range of frequencies with uniform sensitivity.

It is impossible to design for the best performance in all these respects simultaneously, and different designs are required for specific purposes. Two amplifiers are described in this article, the first specifically for use as detector in an a.c. bridge used for iron testing, and the second as a general-purpose instrument which would be suitable as a detector in linear a.f. bridge circuits of the highest precision.

A possible source of error in bridge measurements is that caused by modulation of unwanted harmonics at the detector terminals which produces a component at fundamental frequency and may give rise to an erroneous balance[1,4]. Such errors are most likely to occur to a serious extent in bridges which are either non-linear or where frequency (usually as f^2) is involved in the balance equation. For measurement purposes the latter are usually avoided. Providing the bridge is linear or produces only odd-order components, such errors are minimized by using bridge supplies with negligible even-order distortion. If harmonics are present and sufficiently large they may be observable as a residual at the bridge balance. It is possible to arrange that if harmonics cannot be observed then no significant intermodulation is occurring, but this becomes more difficult as the selectivity of the amplifier is increased.

For minimum intermodulation distortion a frequency-selective stage should occur early in the amplifier so that unwanted frequencies are removed at the lowest possible level.

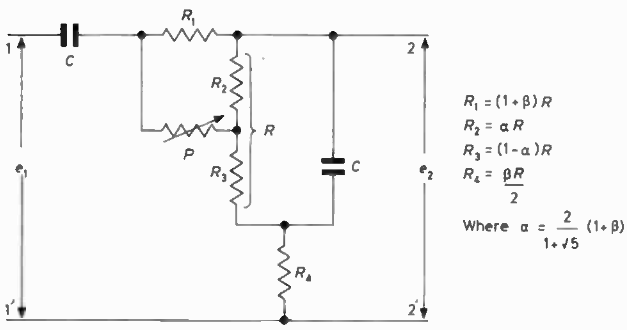

Fig. 1. Modified Wien bridge arms

Unfortunately it is not possible to design a selective stage of the *RC* feedback type having minimal noise. This is because noise associated with the feedback circuit will be present however the feedback and signal are mixed. It follows that for lowest noise it is necessary to precede the selective stage with a low-noise aperiodic stage, and sacrifice performance with respect to intermodulation.

Frequency Selective Network

The frequency-dependent part of the modified Wien bridge network is shown in Fig. 1. The bridge is completed by two resistive arms placed across the terminals 1 and 1′ producing a voltage between their junction and terminal 1′ of $e_1/3$. The output from the bridge is then between this junction and terminal 2. The null frequency of the bridge is varied by the resistor *P*.

The conditions for balance shown below are obtained from the equations given by Wigan[2].

If f_∞ and f' are the frequencies at which the bridge is

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TABLE 1

Component Values and Specifications

RESISTORS	$R_1 = 11\text{k}\Omega$, $R_2 = 6{\cdot}8\text{k}\Omega$, $R_3 = 3{\cdot}2\text{k}\Omega$, $R_4 = 500\Omega$. All above $\pm$ 0·2 per cent. Plessey Metallux. $P = 50\text{k}\Omega \pm 2$ per cent. Inverse log wire-wound.
CAPACITORS	$C = 1\mu\text{F}$ for $f = 18$ to 62c/s and then in series 0·31, 0·1, 0·031 . . . for higher frequency ranges. Values 1, 0·31 and 0·1 μF $\pm \frac{1}{2}$ per cent Polystyrene. All lower values $\pm \frac{1}{2}$ per cent silver mica.

balanced for $P = \infty$ and $P =$ any finite value respectively then:

$$f' = f_\infty \left[1 - \frac{\sqrt{5}\,(1+\beta)}{1 + (1 + (P/R))\dfrac{1+\sqrt{5}}{2}} \right]^{-\frac{1}{2}}$$

where $f_\infty = 1/2\pi CR$

It is seen that the frequency variation which can be effected by variation of P is a function of β and can be made infinite. The selectivity of the circuit, however, decreases as the frequency increases from f_∞. A frequency range of about 3·5:1, corresponding to $\beta = 0{\cdot}1$, was chosen to give two frequency ranges per decade with sufficient overlap. This also gave sufficient discrimination of tuning and a reasonable tuning law using a single turn inverse log potentiometer.

Table 1 shows the values and specification of the components used in the network in the amplifiers to be described.

The resistive bridge arms are formed by the collector and emitter loads of a transistor. As the base current flows only in the emitter circuit the ratio of the loads to produce a 2:1 voltage ratio is not exactly 2:1 but will approach

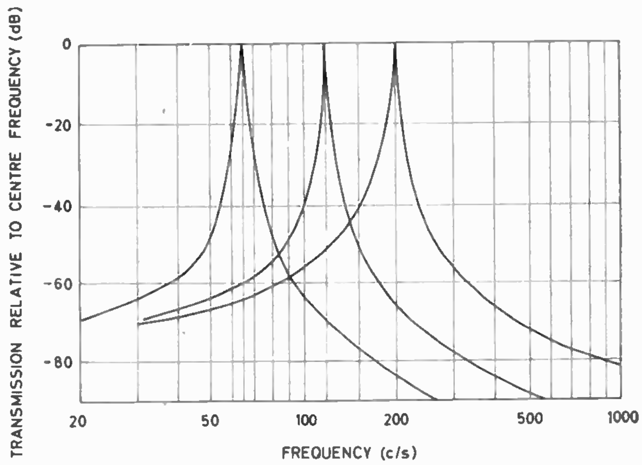


Fig. 3. Frequency response

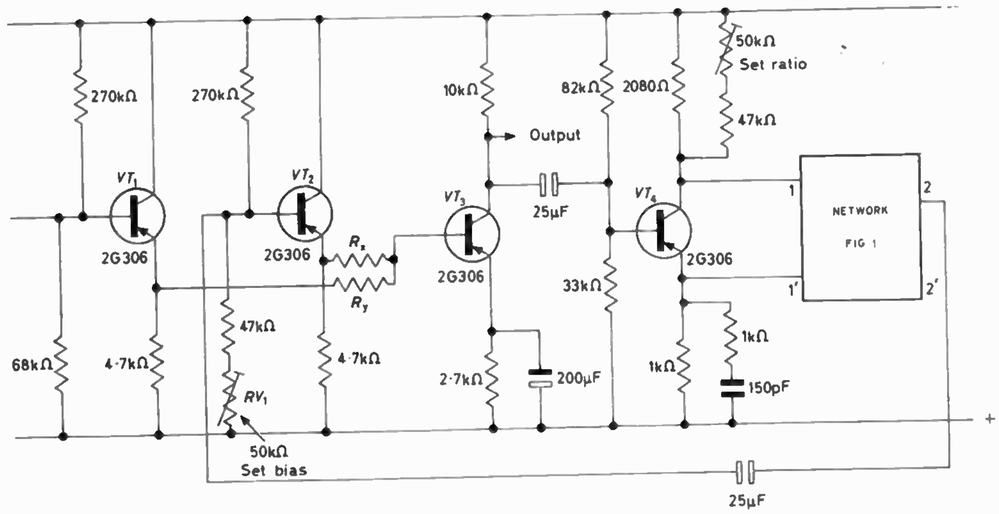


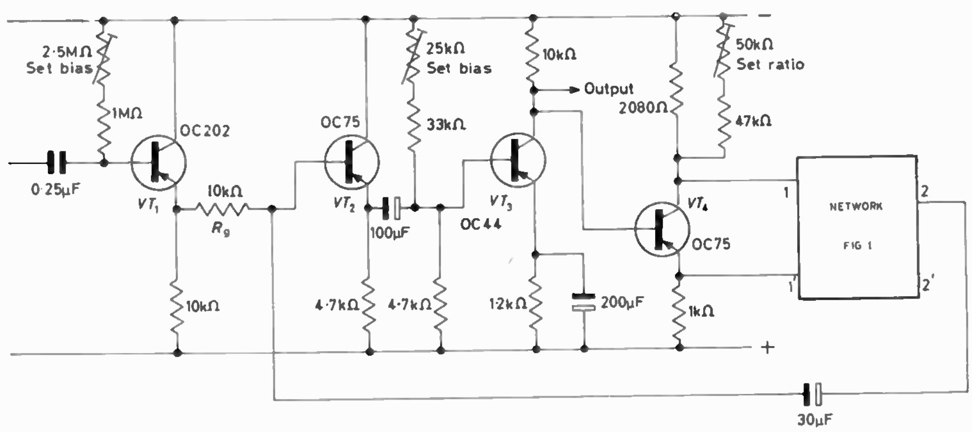
Fig. 4. Frequency selective stage

this value as the transistor current gain is increased. A transistor of high current gain is therefore chosen so that the ratio remains sufficiently constant with changes in external conditions. A small part of the collector load is made variable to facilitate setting up the correct ratio.

Amplifier for Magnetic Loss Measurement

As the magnetic circuit incorporating the test material is non-linear, very large odd order harmonics may be present in the measuring bridge, and a very high selectivity is required to avoid masking the bridge balance. Further the effect of intermodulation between these harmonics and even harmonics in the source must be considered. A sensitivity of no greater than 5μV was required in this application and noise was therefore a minor problem. The frequency coverage was 20c/s to 600c/s.

Fig. 2. Frequency selective stage



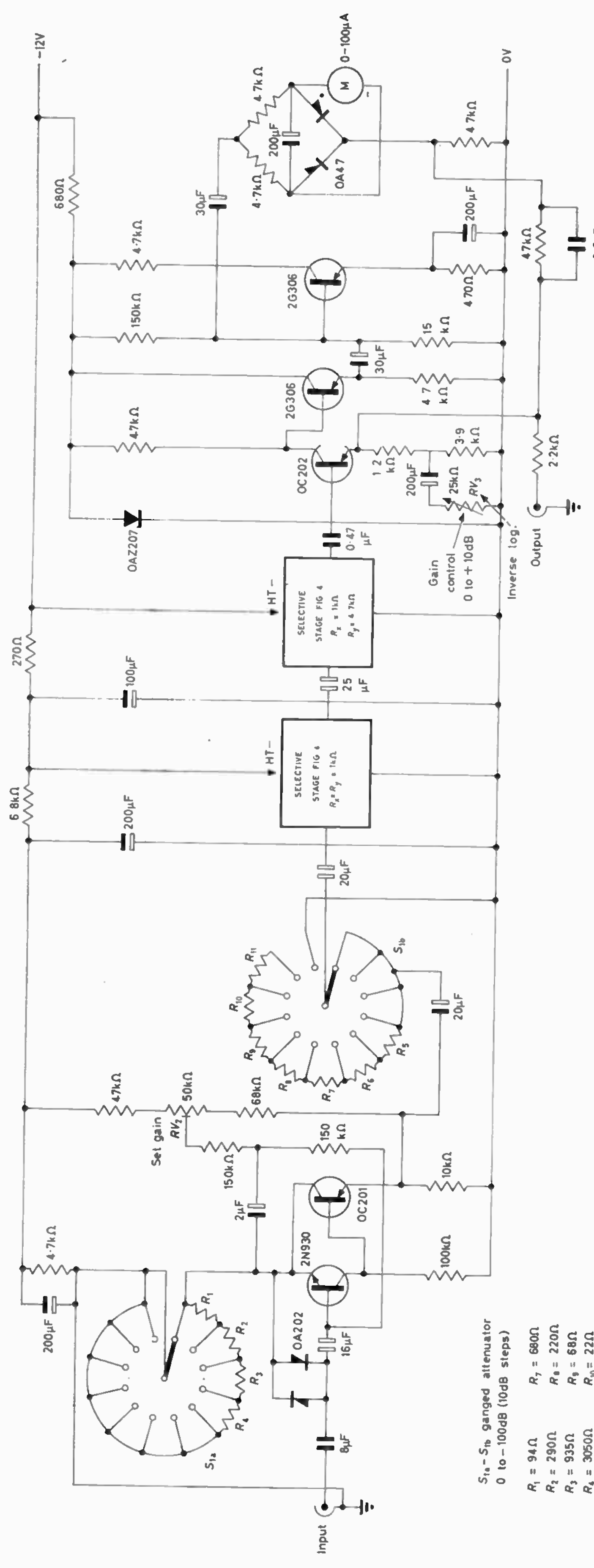


Fig. 5. Frequency selective amplifier

The amplifier consists of a resistive attenuator, two similar cascaded frequency selective stages, each as shown in Fig. 2, and a metering circuit. The selective feedback is returned in shunt with the signal at the base of VT_2: this allows the maximum use of the available feedback. VT_2 is an impedance changer and avoids loading the bridge to the extent that would result from returning the feedback to the base of VT_3. A result of returning the feedback in shunt with the input is that at frequencies above resonance the gain tends to become Z_{OB}/R_g where Z_{OB} is the output impedance of the Wien bridge. This causes a favourable change in the response above resonance as compared to series addition of feedback.

Also, as the value of Z_{OB} at resonance varies with the setting of the frequency dial, there is on any frequency range a fall in the gain at resonance of about 5dB for each tuned stage as the frequency is varied from minimum to maximum.

As the presence of R_g produces a substantial decrease in signal at the base of VT_2 the circuit has poor noise characteristics but improved intermodulation characteristics.

There are three frequency ranges covering 20c/s to 600c/s and the frequency response at three points on the centre range is shown in Fig. 3.

Facility is made for increasing the selectivity of the first selective stage by incorporating a variable resistor of 100Ω in the collector of VT_4 which unbalances the bridge to provide positive feedback at the tuned frequency. An increase of at least 10dB in 3rd harmonic rejection relative to the gain at tuned frequency is obtainable while maintaining adequate stability.

A simple half-wave rectifier circuit is used to display the output on a moving-coil meter, giving a maximum sensitivity for the amplifier-detector of about 2 per cent of full-scale deflexion for 5μV at the input.

General Purpose Amplifier

The overriding consideration in this case was that the amplifier should have low-noise characteristics and a convenient optimum source resistance. It was also desirable to cover the frequency range of 20c/s to 20kc/s and to have a level amplitude response.

The general arrangement is similar to that of the previous amplifier with the addition of a low-noise pre-amplifier. This amplifier should have sufficient gain to control the overall amplifier-noise figure, but not so large or non-linear as to increase intermodulation distortion unnecessarily.

The circuit adopted for the selective stages is shown in Fig. 4. The signal and feedback voltages are buffered by VT_1 and VT_2 and summed at VT_3 base. This produces a gain at the tuned frequency which on any frequency range is practically independent of the setting.

There is, however, a loss both in the feedback loop and in the stage gains. The values of R_x and R_y may be chosen so as to reduce one loss at the expense of the other.

In the first selective stage, equal values of R_x and R_y are chosen as a compromise between low-noise and high-selectivity and in the second stage R_x is made $5R_y$ to produce a higher selectivity at the expense of gain.

The bias of VT_4 is well stabilized in order to minimize changes in the a.c. collector-emitter voltage ratio. The CR network across the emitter load provides high frequency correction. A variable resistor RV_1 is provided to equalize the emitter currents of VT_1 and VT_2.

It has been found that a small percentage of the transistors tried for VT_2 were excessively noisy under the operating conditions. This is only of importance in the first selective stage and is readily detected in the complete equipment as a failure of the noise at the amplifier output to be significantly reduced as the amplifier gain is reduced.

The complete amplifier circuit is shown in Fig. 5. The input stage comprises a silicon npn planar type 2N930 and a silicon pnp alloy type OC201. The input transistor

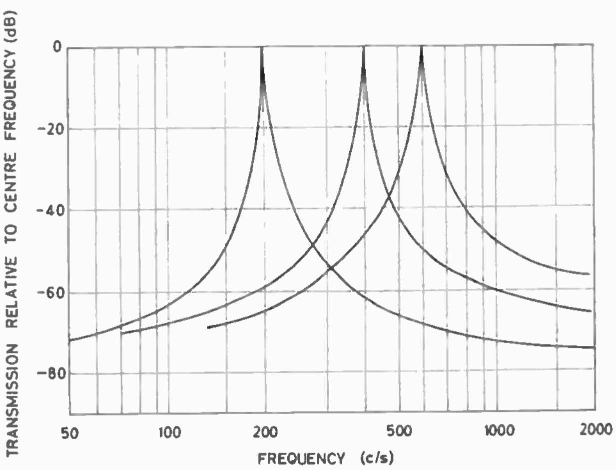


Fig. 6. Frequency response

is operated at about 30μA collector current. The d.c. conditions of this pair of transistors are stabilized by heavy feedback which simultaneously stabilizes the a.c. gain. The a.c. gain is set to 100 with switch S_1 set at 0dB, by adjustment of the 50kΩ potentiometer RV_2 in the d.c. biasing circuit.

The attenuator between this and the following stage gives a total of 100dB in 10dB steps. The first four steps are obtained by negative feedback, thus increasing the linearity of the stage as the input signal is increased. The remaining steps are obtained by a tapped divider across the output of this stage.

The design of the attenuator takes into account the input impedance of the following stage, which is largely that due to the biasing network of the input transistor.

The input resistance of the input stage varies according to the attenuator setting and is approximately 70kΩ at 0dB and 2MΩ from -40dB onwards.

The metering stage which incorporates a negative feedback path including the rectifier circuit has a linearity as high as that of the meter, and does not suffer from lack of sensitivity at low voltage levels.

A variable resistor RV_3 in the meter circuit allows a continuous 0 to $+10$dB variation in the sensitivity.

At minimum setting of this control full-scale deflexion of the meter corresponds to a signal of $\frac{1}{2}\mu$V $\pm$3dB depending on frequency.

There are six frequency ranges covering the frequency range 20c/s to 20kc/s. The frequency response is shown at three settings on the 200 to 600c/s range in Fig. 6.

The equivalent short-circuit noise voltage e_n and open-circuit noise current[5] i_n, are shown as a function of frequency in Fig. 7.

In order to obtain the lowest possible value for e_n there should be no appreciable impedance between the input terminals and the input transistor. Thus at 20c/s the

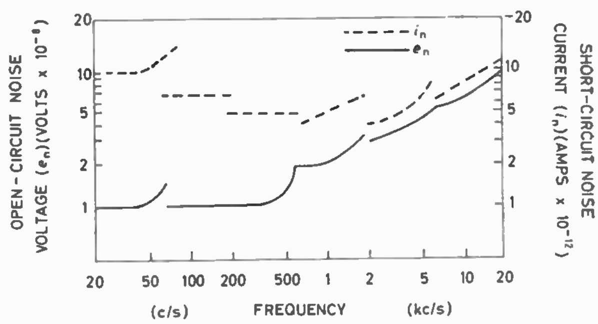


Fig. 7. Short-circuit and open-circuit amplifier noise

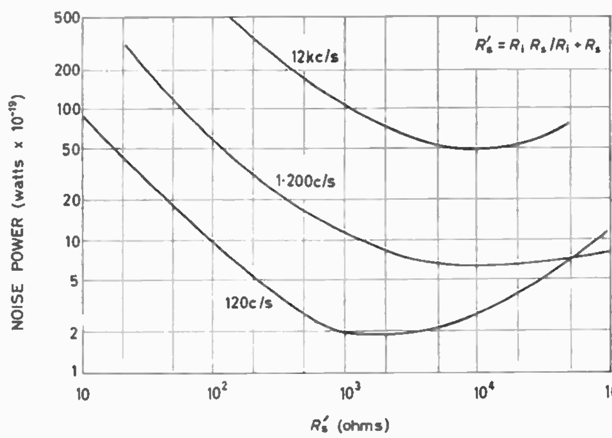


Fig. 8. Amplifier noise power

smallest tolerable value of input capacitor is 8μF, although a very much smaller value would produce negligible loss in gain.

The noise power at the input is plotted as a function of the equivalent source resistance R_s' for three frequencies in Fig. 8. R_s' is given by the parallel resistance of the source resistance R_s and the amplifier input resistance R_i. The minimum on these curves agrees closely with the optimum predicted from $R_o = e_n / i_n$.

In order to test for intermodulation equal signals of 2000 to 3000c/s are added at the amplifier input. With both the attenuator and the variable gain control set at 0dB signals of 100μV each produce a full-scale deflexion (equivalent to 0·7μV at the input) at the difference frequency. With the attenuator changed to -40dB (i.e. maximum feedback in initial stage) signals of 10mV each produce about half full-scale deflexion. If either of the above is repeated with the same overall gain, but obtained by increasing the variable to $+10$dB and setting the attenuator 10dB down then the deflexion at difference frequency is very little greater than the breakthrough of the 2000c/s signal.

A further modification, not shown, is the facility to make either or both selective circuits aperiodic, simultaneously incorporating about 10dB of negative feedback. This is done by breaking all connexions to the frequency selective network shown in Fig. 4 and connecting a feedback resistor between VT_4 emitter and the capacitor in series with VT_2 base.

Battery Supply

Both amplifiers described are powered by two PP9 batteries. These feed a stabilizer which uses a compound emitter-follower and a Zener diode. This provides a fairly constant supply at about 12V for battery voltages down to about 13·5V.

Acknowledgments

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A Numerically Controlled Punching Press

Brown, Boveri & Cie AG, Mannheim, Germany, recently exhibited a numerically controlled punching press with turret tool holder. The turret is equipped with 24 tools for the punching of plate up to 0·118in thick. The sheet is clamped to a table 94·5in long and 70·85in wide.

The table is positioned electrically and simultaneously along both co-ordinates at a speed of approximately 33ft/min with a positioning error of better than 0·004in ensured by an analogue basis of the measuring system. At the same time the turret tool holder is indexed to the programmed tool.

The data input is numerical throughout in the form of eight hole punched tape, using the programming code 8B according to the German specification VDI 3259. Where the preparation of punched tape appears to be uneconomical, i.e., for one-off or small batches, the information can be keyed in manually at the control console.

The instructions stored in the 15 decade memory comprise the following information:

Instruction number

Tool number

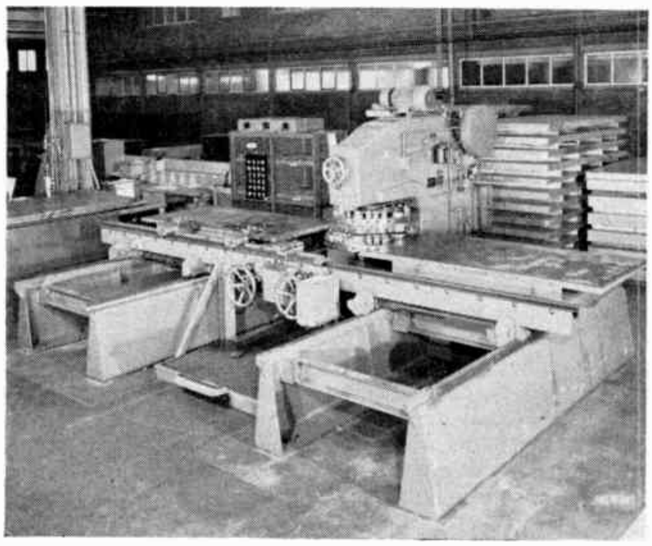
Travel to position along X and Y co-ordinates.

The order number for the part is given at the beginning of the programme, which may be interrupted at any position to indicate 'change of tools.'

Where the programme interruption has not been intended, viz. power failure or pressing of the emergency button, all measured values are retained so that the programme will continue from the same position when the machine is restarted.

In addition, a facility for nibbling has been incorporated in which only the distance is controlled. The starting and finishing points of the nibbling distance are programmed and between these points the machine punches automatically at a speed of about 60 strokes/min.

The Brown Boveri punching press



A Transistor Instrumented Nuclear Reactor

A 100kW nuclear reactor has been built by Pye Ltd for Liverpool and Manchester Universities.

The reactor, which has a fully transistorized instrumentation system, is 20ft high and covers an area of 400ft². It is situated at Risley, Lancashire and will be used for postgraduate research and to teach the principles, design and operation of nuclear equipment. The new Pye reactor will also be used as a research facility for a wide range of investigations. The Pye reactor is a light-water-and-graphite-moderated assembly with a thermal output of 100kW. The basic element of the moderator is a rectangular prism constructed of graphite bricks. The graphite surrounds the core and acts as the principal moderator and reflector. It also forms the whole of the primary thermal column and part of the secondary thermal column, two of the main experimental facilities of the reactor. The core assembly is capable of taking a maximum of 24 fuel elements. The fuel charge contains 3·5kg of uranium 235.

The coolant and part-moderator is light-water pumped up through the elements with a maximum flow rate of 90gal/min. The heat is finally dissipated by an air-cooled radiator outside the reactor building. The reactor is controlled by moving four control blades in the spaces between the fuel boxes. An additional feature is the facility for dumping water from all the fuel boxes through an electromagnetically controlled dump valve. 350 tons of barytes concrete is used for shielding.

The nuclear detectors used to measure the reactor power are situated in stringers in the primary thermal column. The instrumentation comprises five nuclear channels, a start-up channel using a fission chamber, a linear and a logarithmic power channel both using compensated ionization chambers and two shut-down channels using uncompensated chambers.

The safety circuits of the reactor consist of four guard lines; primary and secondary lines associated with SCRAM conditions and primary and secondary lines associated with TRIP conditions. Any break in the SCRAM guard lines causes all rods to drop in and water to be dumped; any break in the TRIP guard lines causes the two shim control rods and the regulating rod to be dropped. A system of alarm signals is incorporated with the safety circuits.

Experiments on an X-Band Garnet Delay Line

By J. H. Collins*, M.Sc., B. Yazgan*, M.S., and J. Cochrane*

Pulse experiments are described on the excitation of X-band spin-acoustic waves in a ⅛in diameter axially magnetized single crystal disk of yttrium iron garnet, inserted in a room temperature TM_{010} resonant cavity. The minimum insertion loss of the device was 45dB, over the input power range −36 to +7dBm.

(Voir page 651 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 658)

RECENTLY Eshbach[1,2] has reported the generation, at X-band frequencies and liquid helium temperatures, of radially propagating spin and shear waves in thick disks of single crystal yttrium-iron garnet (YIG) subject to axial magnetic field biasing. The technological significance of these results lies in the future possibility of fabricating physically compact transmission lines, for the 3 000 to 20 000Mc/s frequency range, yielding microsecond order variable time delays with insertion losses comparable to

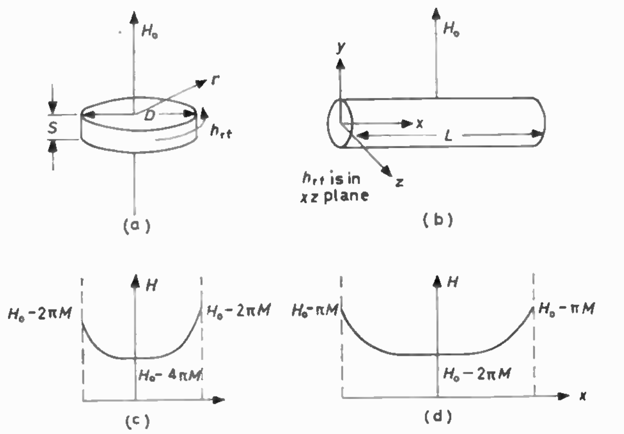


Fig. 1. Suitable geometries for spin-wave excitation in yttrium-iron garnet
(a) Thick disk geometry
(b) Finite length rod geometry
(c) Magnetic biasing field distribution within disk
(d) Magnetic biasing field distribution within finite rod

their bulky coaxial line and waveguide counterparts. This article describes the authors' experiments on a room temperature X-band garnet delay line aimed at determining its insertion loss and echo time as a function of microwave drive power and magnetic biasing field.

Principle of Operation

In order to excite a spinwave (a magnetic dipole wave of extremely rapid spatial variation) in a saturated ferrite from a spatially uniform r.f. drive field, without recourse to parametric effects, it is necessary that the instantaneous transverse magnetic moment does not sum to zero over the specimen[3]. This requirement can be realized by both the thick disk and finite length rod geometries shown in Fig. 1. Each is biased by a uniform d.c. magnetic field, H_0, at right-angles to which is applied the microwave magnetic field, h_{rf}. The appropriate distributions of d.c. magnetic field are such that a radial variation of one-half and an axial variation of one-quarter the saturation magnetization, $4\pi M$, occur in the disk and rod cases respectively. Both arrangements excite plane waves, favoured to travel along the direction of the stated field variation, at the drive angular frequency, ω, for a range of H_0 determined by $4\pi M$, ω and the gyromagnetic ratio, γ. Focusing attention on the disk this range is defined approximately by:

$$\sqrt{[((\omega/\gamma))^2 + (2\pi M)^2]} < H_0 < 2\pi M + \sqrt{[(\omega/\gamma)^2 + (2\pi M)^2]} \quad \ldots\ldots (1)$$

The lower and upper bounds correspond to excitation near the edge and centre of the disk respectively, the latter being the more efficient: these bounds lie below the normal ferromagnetic resonance (uniform precession mode) and magnetostatic mode spectrum.

It is possible in a cubic crystal for the dispersive spin-waves to be degenerate in frequency, phase velocity and propagation direction with acoustic shear waves polarized parallel to H_0, particularly if the latter is applied in a [100] direction. For an infinite uniformly magnetized medium continuous interaction then occurs between the magnetization and the mechanical strain of the lattice giving rise to a coupled mode system. However, in the case of non-uniform magnetization, the coupled region is very sharp resulting usually in conversion of spin to acoustic energy and virtually pure mode propagation throughout. Thus in both the thick disk and ferrite rod geometries the non-uniform internal d.c. magnetic field can have two effects, namely, to convert electromagnetic to spin energy followed by spin to acoustic energy. Since the inverse processes occur, by the symmetry of the arrangements, the net effect is that a portion of the incident electromagnetic energy is stored and decays in the ferrite for a finite time before being partially released in electromagnetic form. A further property is that the delay is variable by changing the d.c. field level and hence moving the coupling plane in the sample.

When the d.c. field, H_0, is in proximity to the upper bound defined in equation (1), the maximum radial spin wave propagation number is less than that necessary to provide magnetoelastic interaction and hence acoustic mode propagation. Propagation throughout the disk then takes place in the form of spin waves.

Description of Apparatus

The YIG disk, of 0·125in diameter and 0·012in thick with its axis oriented along [100], was cut from a 5 gram single crystal supplied by J. Nielson, Airtron Co. Morris Plains, N.J. The sample was supported centrally in the gap of a re-entrant TM_{010} brass cavity, resonant to 9 110Mc/s, by a 0·100in p.t.f.e. post bonded to the cavity metal post (see Fig. 2). The cavity was coupled to the WG16 system by a circular iris, thin with respect to the size of the hole in it, chosen such that the cavity system was a reasonable match at the nominal magnetic biasing field, which was applied along the cavity axis from a current stabilized electromagnet of simple construction. The measured loaded Q of the cavity was about 200.

Fig. 3 shows a simplified block diagram of the equip-

* *The University, Glasgow.*

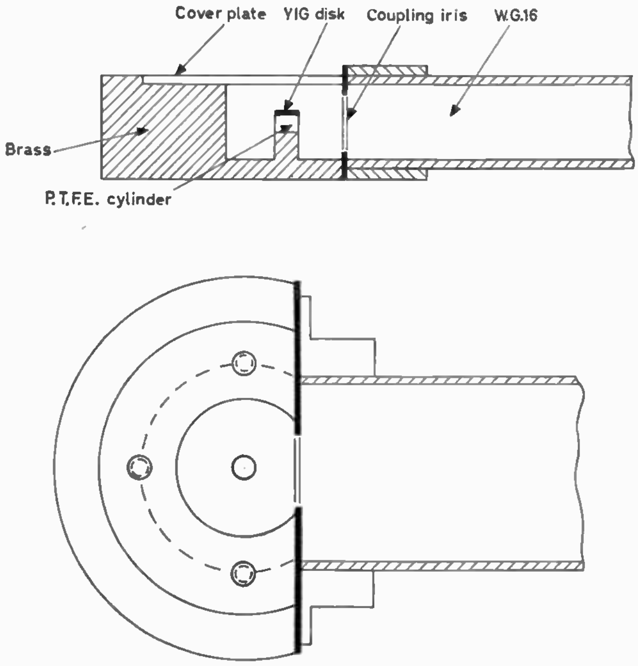


Fig. 2. Re-entrant cavity construction

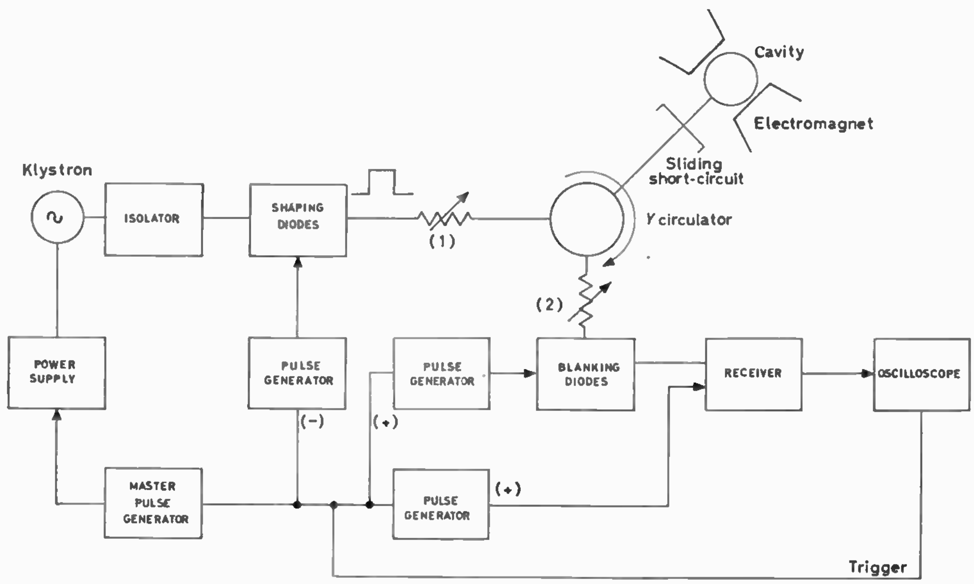


Fig. 3. Simplified arrangement of apparatus

ment. The klystron was cathode modulated from a buffered pulse generator which also supplied pulses to the oscilloscope, receiver blanking and varactor pulse shaping circuits. Attempts were made to shape the klystron output pulse using three cascaded varactor diodes (Ferranti Type 100XSS) operated as gates. However, with a common drive pulse (7V amplitude) and little interstage isolation the expected rejection ratio of 25dB per stage was not achieved. Measurements were, therefore, generally conducted using an incident pulse of between 0·35 and 0·50μsec duration, 0·15μsec rise and fall time, obtained directly from the klystron. The microwave pulse was passed through attenuators (1) and a three port *Y* circulator to the test cavity. Power reflected instantaneously from the cavity, that is, without absorption and later re-emission from the disk, is attenuated by the blanking diodes and by blanking in the i.f. strip of the balanced mixer superhet receiver. The power reflected by the blanking diodes is absorbed in an isolator. For the i.f. strip a 3V positive pulse was applied to the cathode of the first valve. By the time the energy is re-radiated from the sample the diodes are in all pass condition and the i.f. blanking is removed. The energy of spin-acoustic origin thus enters the receiver, there to be amplified and subsequently displayed on the oscilloscope. The receiver system contained a 30Mc/s i.f. strip of 10Mc/s bandwidth and had an overall noise figure of 12dB, corresponding to a matched noise power level of −92dBm.

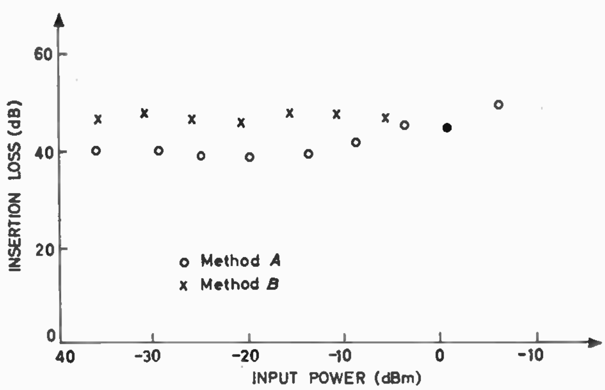
Measurement Procedure and Results

Essentially insertion loss and echo time measurements were made on the YIG system, at room temperature, using the following two procedures:

METHOD A

Having adjusted the system to obtain the desired echo with the main pulse blanked, and in the absence of receiver saturation, the height of the echo on the oscilloscope was noted. The blanking was removed, the cavity arm short-circuiting switch closed and attenuators (1) adjusted until the input pulse was brought to the same height as the echo. The estimated accuracy of the procedure is about 2dB.

Fig. 4. Insertion loss of echo pulse versus input power



METHOD B

Attenuators (2) were inserted until the echo pulse coincided with receiver noise level. The estimated accuracy is about 5dB.

In both methods measurements were arranged of incident power using a power meter, where possible, and also with respect to receiver noise level. For the higher incident powers receiver saturation (dynamic range approximately 40dB) was possible from the pulse of spin-acoustic origin. Small variations in the results observed at different times occurred due to changes in shape of the main pulse and the settings of components in the cavity arm.

The insertion loss, between the main and echo pulses, at a biasing field of about 4 100 gauss is shown in Fig. 4 over the range of input power -36 to $+7$dBm. The lower limit was set by the requirement that the echo signal be a reasonable height above noise and the upper limit by the pulsed power that the varactor diodes could handle. The observed loss is typically 45dB although some signs of non-linearity at 6mW, resulting in increased insertion loss, are evident using method A. Fig. 5 shows a typical photograph of the receiver output as a function of time: the incident and first reflected pulse are 81 and 33dB above noise respectively, the oscilloscope time scale is 0·5μsec/cm. To the left of the origin, the small random pulses are due to inefficient blanking. The pulse just to the right of the origin appears delayed by approximately 0·5μsec with respect to the signal incident on the cavity. Further output pulses occur but these are thought to be spurious and due to receiver ringing. One of the experimental difficulties, arising from the small time delay available in the size of disk employed, is that the blanking pulses cannot be removed quickly enough to observe the leading edge of the spin-acoustic pulse.

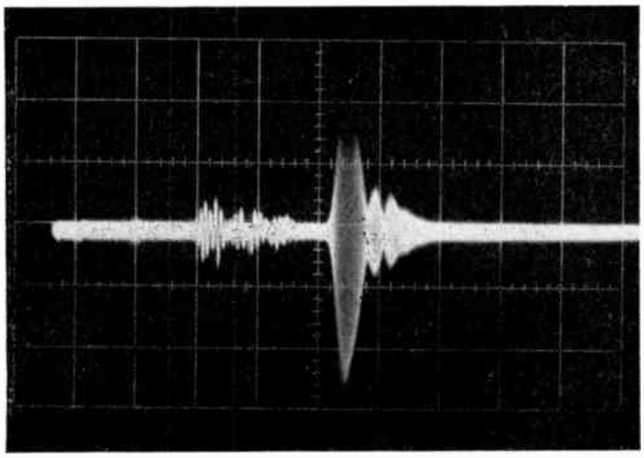
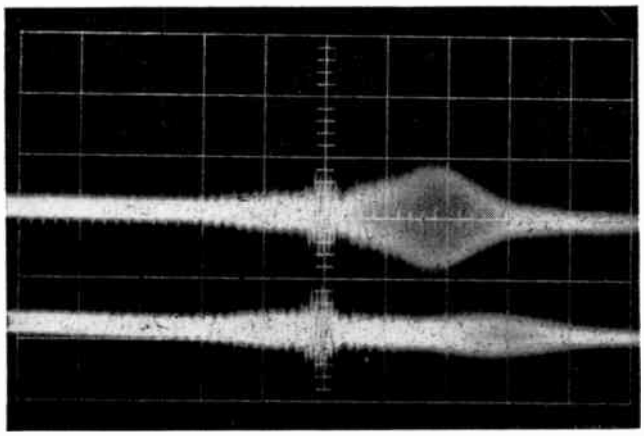


Fig. 5 (above). Typical echo pulse from YIG disk
Input pulse length = 0·3μsec
Oscilloscope scale = 0·5μsec/cm

Fig. 6 (below). Echo pulses from YIG disk showing effect of varying magnetic bias field
Input pulse length = 0·5μsec
Oscilloscope scale = 0·2μsec/cm



The photographs in Fig. 6, oscilloscope time scale 0·2μsec/cm, were taken using diode shaping circuits with fairly slow removal of the blanking pulses. The main and reflected pulses (upper trace) were 63 and 18dB above noise respectively. In the lower trace the magnetic field has been decreased by about 50 gauss resulting in an echo reduction of 15dB for a delay increase of 0·2μsec. The maximum delay observed between the trailing edges of the main and echo pulses in our experiments was 1·5μsec at an incident power level of 6mW. It was found that both the magnitude and orientation of the magnetic biasing field were crucial to about 100 gauss and 1° respectively for the spin-acoustic pulse generation. The normal cavity absorptions by ferromagnetic resonance and magnetostatic modes were observed at biasing field in accordance with conventional theory.

Discussion

It was observed that echoes only occurred over a small range of magnetizing field close to the upper limit of equation (1). Hence it is unlikely that spin-acoustic energy conversion occurred and that rather propagation was in the form of pure spinwaves. Theoretically this phenomenon is expected over a field interval, δH_o, given by:

$$\delta H_o = \frac{\omega_{ex} l^2}{\gamma}((\omega/v_s))^2 \qquad \ldots\ldots\ldots\ldots\ldots (2)$$

where $\omega_{ex} l^2$ is the exchange constant, namely 0·88 $\times$ 10^{-1} (cm^2 sec^{-1}) for YIG.

v_s is the shear wave velocity, namely 3·87 $\times$ 10^5 (cm sec^{-1}) for YIG. The computed value for δH_o of 110 gauss compares favourably with the observed range of about 100 gauss.

The 15dB increase in insertion loss for a 0·2μsec increase in delay indicates a very high attenuation factor for room temperature operation at X-band frequencies. Reductions in frequency and temperature would be expected to decrease the propagation losses very significantly.

The absence of echoes at lower fields could be due to the reduced efficiency of spinwave excitation at these fields. A more tenable reason is that as the YIG disk was unpolished and contained surface pits of the order of 5 microns and, further, was bonded to a p.t.f.e. post, high external acoustic damping occurred, suppressing any acoustic propagation. Experiments are in hand with polished samples, loosely clamped, to check these assertions. Improvements in cavity design to reduce coupling losses are also contemplated.

Acknowledgment

Acknowledgments are due to the Ministry of Aviation for supporting this work, to P. Hlawiczka for helpful discussions, and to Professor J. Lamb for facilities made available to the authors in the Electrical Engineering Department, University of Glasgow.

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An Electronic Scanner for Transistor Pairs used in Balanced Pre-Amplifiers

By R. R. Vierhout† and A. J. H. Vendrik†

It is shown that the transistor parameter which determines mainly the rejection ratio of balanced voltage pre-amplifiers is $r_b^* = h_{11}/(h_{21}+1)$.

An electronic scanner which displays this quantity against emitter-current for two transistors simultaneously has been developed. A possibility for calibration is included.

(Voir page 651 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 658)

In balanced amplifiers the rejection ratio of out-of-phase to in-phase input voltages is theoretically infinite if the transistors and connected impedances at both sides are exactly equal. If the parameters of both transistors are not equal it is possible to obtain an infinite rejection ratio by using collector impedances with appropriate different values.

However, the last procedure has practical difficulties if the parameters of both transistors differ too much from each other. E.g. the collector emitter bias voltage of one of them may become too small causing distortion of moderate input signals. Moreover, a series product becomes cheaper using transistor pairs.

Therefore an electronic scanner has been developed which displays on an oscilloscope screen simultaneously the quantity $h_{11}/(h_{21}+1)$ of each of two transistors at various emitter currents. This quotient is the relevant quantity. The voltage rejection ratio in balanced pre-amplifiers is mainly dependent on this quantity as will be shown.

Theory

Fig. 1 shows a balanced amplifier with two transistors. The resistances r_{b1} and r_{b2} are the base-emitter resistances at the working condition transferred to the input. This transferred resistance r_b is the parameter h_{11}. The parameter h_{21} is the current amplification factor β.

If the voltage at the connected emitters is called V_e and the internal collector emitter conductances of the transistors are neglected, then:

$$\frac{V_1 - V_e}{r_{b1}}(\beta_1 + 1) + \frac{V_2 - V_e}{r_{b2}}(\beta_2 + 1) = (V_e/R_e) \quad \ldots\ldots\ldots\ldots (1)$$

$$-V_{c1} = R_{c1} \,.\, i_{c1} = R_{c1} \,.\, \frac{V_1 - V_e}{r_{b1}} \,.\, \beta_1 =$$

$$R_{c1}\frac{\beta_1}{\beta_1 + 1}\,\frac{V_1 - V_e}{r_{b1}}(\beta_1 + 1) \quad \ldots\ldots (2a)$$

$$-V_{c2} = R_{c2}\frac{\beta_2}{\beta_2 + 1}\,\frac{V_2 - V_e}{r_{b2}}(\beta_2 + 1) \quad \ldots\ldots\ldots (2b)$$

Let

$$\frac{r_b}{\beta + 1} = \frac{h_{11}}{h_{21} + 1} = r_b^* \text{ and } R_c\,\frac{\beta}{\beta + 1} = R_c^*$$

The quantity r_b^* is h_{11} transferred to the emitter side.

From equations (1), (2a) and (2b) it follows that:

$$-V_{c1} + V_{c2} =$$

$$\frac{(V_1 - V_2)(R_{c1}^* + R_{c2}^*) + V_1 r_{b2}^*\,(R_{c1}^*/R_e) - V_2 r_{b1}^*\,(R_{c2}^*/R_e)}{r_{b1}^* + r_{b2}^* + \dfrac{r_{b1}^* \,.\, r_{b2}^*}{R_e}} \quad \ldots\ldots\ldots\ldots (3)$$

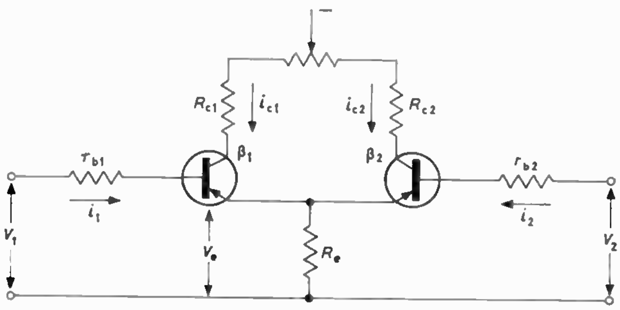

Fig. 1. The balanced pre-amplifier

Calling the rejection ratio D then:

$$D = \frac{2R_e((1/R_{c1}^*) + (1/R_{c2}^*)) + (r_{b2}^*/R_{c2}^*) + (r_{b1}^*/R_{c1}^*)}{(r_{b2}^*/R_{c2}^*) - (r_{b1}^*/R_{c1}^*)} \quad \ldots\ldots\ldots\ldots (4)$$

If $r_{b1}^*/R_{c1}^* = r_{b2}^*/R_{c2}^*$ the rejection ratio D will be infinite.

As $\beta \gg 1$ $R_c^* = R_c$, therefore practically independent of transistor parameters. The transistor quantity which determines the rejection ratio is r_b^*.

Small signal transistors have an r_b^* of about 20Ω, R_c is several hundreds to thousands of ohms, R_e is of the order of the value of R_c. Hence the last two terms in the numerator of equation (4) can be neglected. Then if $R_{c1}^* = R_{c2}^*$ and r_{b1}^* differs less than 10 per cent from r_{b2}^*

$$D > (4R_e/0{\cdot}1\, r_b^*) \simeq (4000/2) = 2000 \text{ if } R_e = 1000\Omega$$

In this calculation the collector-emitter resistance is considered to be infinite. However, if this resistance is assumed to have a value of about 20kΩ, a more complicated calculation‡ shows that 100 per cent difference of the collector-emitter resistances of the two transistors results in a rejection ratio of about 2800 if $r_{b1}^* = r_{b2}^*$.

As the collector-emitter resistances of transistors of the same type differ less than 100 per cent, the influence of this difference can be neglected if one is only interested in

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‡ *Taking into account finite values of the collector emitter resistances* r_{d1} *and* r_{d2} *the rejection ratio* $D = 2/3 \,.\, R_e/r_b^* \,\dfrac{r_{d1} + r_{d2} + 2r_{d1} \,.\, r_{d2}/R_e}{r_{d2} - r_{d1}}$ *if* $r_{b1}^* = r_{b2}^*$

rejection ratios of less than 2000, involving a difference in r_b^* of less than 10 per cent.

The Scanning Circuit

The value of r_b^* as a function of the emitter current is measured with the circuit shown in Fig. 2.

A ramp voltage is put between the base and point A. If the duration of a cycle is T and the top voltage E, then the equation of the curve during a cycle is equal to $V = (E/T).t$, taking $t = 0$ at the start of the cycle. The diode restores the zero voltage at $t = 0$. The voltage at B is equal to:

$$\frac{r_b}{(\beta+1)R_e+r_b}.(E/T).t = \frac{r_b^*}{R_e+r_b^*}.(E/T).t \simeq$$
$$r_b^*(E/R_e).(t/T) \text{ if } R_e \gg r_b^*$$
$$[R_e = 4{\cdot}2\text{k}\Omega;\ r_b^* = 15\Omega]$$

A differentiating circuit between B and D transfers this voltage to $r_b^*.(E/R_e).(RC/T)$ at point D. This voltage is proportional to r_b^* if E, R_e, RC and T are held constant. Then if point D is connected to the y-input of an oscilloscope, the vertical deflexion is directly proportional to r_b^*.

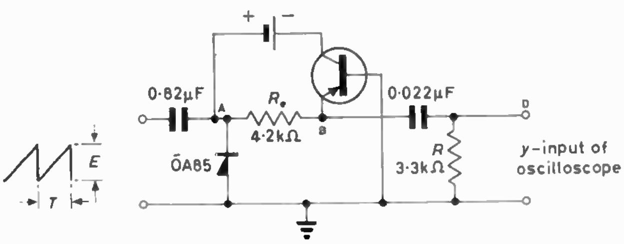


Fig. 2. The basic scanning circuit for $r_b/(\beta+1)$ of one transistor

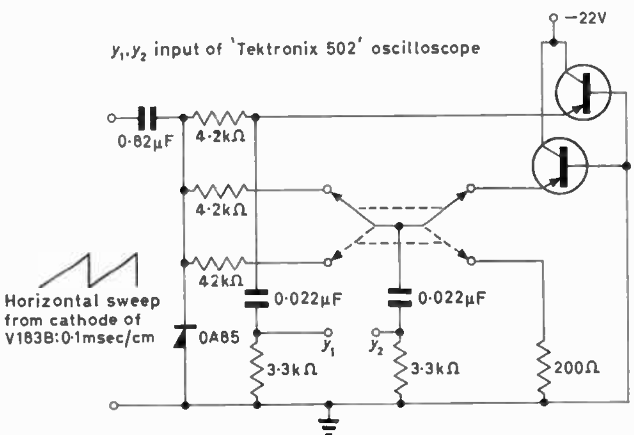
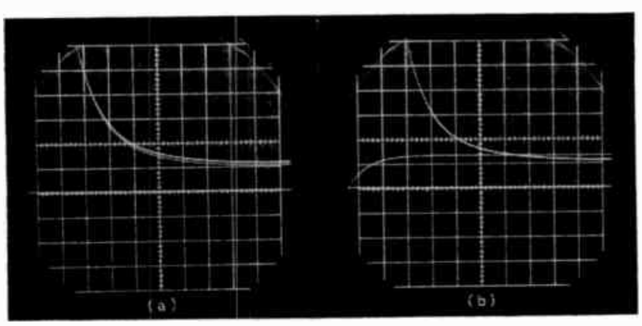


Fig. 3 (above). Circuit for comparison of the values of $r_b/(\beta+1)$ of two transistors and of one transistor and a fixed resistance

Fig. 4(a) (below left). $r_b/(\beta+1)$ of two transistors simultaneously displayed (vertically) against emitter-current (horizontally)

Fig. 4(b) (below right). $r_b(\beta+1)$ of one transistor compared with a fixed resistance displayed vertically against current flowing through the emitter or the resistance respectively



A Tektronix double beam oscilloscope type 502 was used and the ramp voltage taken from the cathode of V_{183B} of the horizontal amplifier. This voltage actually sweeps from about −19V to +2V, so that the maximum current $E/R_e = 5$mA. A time-base velocity of 0·1msec/cm was chosen, so that $T \simeq 1$msec.

$RC = 3{\cdot}3 \times 10^3 \times 22 \times 10^{-9}\text{sec} = 0{\cdot}07$msec, which is sufficiently small in comparison with $T \simeq 1$msec to consider the circuit between B and D as a differentiating circuit.

The voltage at point D is now:

$$r_b^*.5.(0{\cdot}07/1) = 0{\cdot}35\,r_b^*\text{mV} \quad \ldots\ldots\ldots (5)$$

This voltage is fed to one of the y-inputs of the same oscilloscope, the sensitivity of which is chosen to be 5mV/cm.

An increase of r_b^* with 10Ω causes a beam deflexion of 7mm. If the transistor has an r_b^* of 15Ω a difference of 10 per cent in this value causes a deflexion difference of 1mm, which can be read easily.

In order to compare two transistors, two of these circuits are used having only the ramp voltage input signal, the d.c. restoring input circuit and the battery in common. The two output voltages are fed to the y_1 and y_2 input of the oscilloscope respectively.

The horizontal deflexion is proportional to the ramp voltage and hence to the emitter current. The maximal deflexion is about 10cm, the maximal emitter current about 5mA. During the ramp voltage cycle the course of r_b^* of both transistors as a function of the emitter current is displayed on the oscilloscope screen. The emitter current at which the two r_b^* values are equal can immediately be read. In the balanced amplifier this emitter current can be used.

In order to measure the absolute value of r_b^* the resistor R_e is replaced by a fixed resistor of 42kΩ and the transistor by a resistor of 200Ω. The deflexion of the beam obtained with this circuit corresponds to an r_b^* of 20Ω.

It is advantageous to use the horizontal sweep for the input of the circuit as the back sweep is blanked on the oscilloscope and no triggering is needed.

The actual circuit including the calibration part is shown in Fig. 3.

Fig. 4(a) shows the display of the course of r_b^* of two transistors and Fig. 4(b) of one transistor and the calibration resistance.

An Aid to Magnetic Tape Editing*

'Word spotting', i.e. locating a particular part of a recording, is a technique frequently used in the editing of magnetic tape recordings. One well known method uses a drum with a replay head mounted in its circumference. The magnetic tape is held stationary over part of the circumference of the drum, and as the drum rotatoes, that part of the tape which is held against the drum is replayed once per revolution. By adjusting the position of the tape, the required part of the recording can be found. The tape is then cut at the appropriate point and spliced to a second tape which has been treated in a similar manner. However, the final result cannot be properly assessed until the splicing has been carried out and the spliced tape replayed, and it would be much more convenient if the editor could hear the result before committing himself to cutting the tapes.

An improved method has therefore been devised using two similar drums rotating at the same speed and so arranged that when the replay heads reach a certain marked point on the circumference, the output is switched from one replay head to the other. When two tapes are held against the drums, the output sounds as if the two tapes had been cut at the marked point and spliced together. The simulated cutting point on either tape is adjusted by moving either or both tapes to and fro, and so the precise point at which the tapes should be cut in order to give the required result can be found and the result heard before the cutting and splicing are carried out.

* *A communication from E.M.I. Ltd.*

LETTERS TO THE EDITOR

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

Remarks on the Error-Function Limiter

DEAR SIR,—Consider a device whose transfer characteristic is defined by:

$$y = \frac{1}{K\sqrt{(2\pi\sigma_d^2)}} \int_0^x \exp(-t^2/2\sigma_d^2)\,dt$$

for $x \geqslant 0$, and $y = 0$ for $x < 0$.

Since the error function is given by

$$\operatorname{erf} x = 2/\sqrt{\pi} \int_0^x \exp(-t^2)\,dt$$

the above characteristic becomes

$$y = 1/2K \operatorname{erf}(x/\sigma_d\sqrt{2})$$

(The limiting value of y is $1/2K$). This device can be referred to as a half-wave error function limiter.

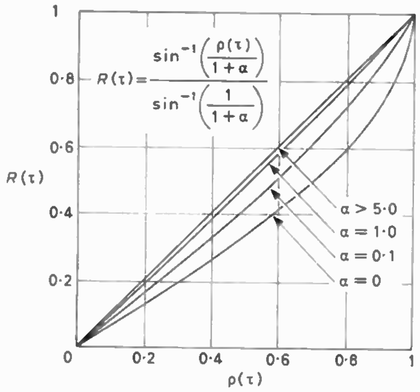


Fig. 1. Relation between R(τ) and ρ(τ)

In a study of the symmetrical (odd) limiter with the above characteristic, Baum[1] has shown that if the input to the device is noise with a Gaussian probability distribution

$$f(x) = \frac{1}{\sqrt{(2\pi\sigma_i^2)}} \exp(-x^2/2\sigma_i^2)$$

then the normalized correlation function at the output of the device is given by

$$R(\tau) = \frac{\sin^{-1}[\rho(\tau)/(1+\alpha)]}{\sin^{-1}[1/(1+\alpha)]}$$

where $\rho(\tau)$ is the normalized input correlation function and $\alpha = \sigma_d^2/\sigma_i^2$. (Our notation differs slightly from Baum's).

As $\alpha \to 0$ on obtains in effect the ideal hard limiter and as $\alpha \to \infty$ the linear amplifier.

The purpose of this note is to show how $R(\tau)$ varies with $\rho(\tau)$ for various values of α. The curves are shown in Fig. 1. The striking feature is the small departure from linearity even for values of α not too far from zero. Let $\Delta R(\tau)$ denote the deviation of $R(\tau)$ from $\rho(\tau)$:

$$\Delta R(\tau) = \rho(\tau) - \frac{\sin^{-1}[\rho(\tau)/(1+\alpha)]}{\sin^{-1}[1/(1+\alpha)]}$$

Differentiating with respect to $\rho(\tau)$ and setting the resulting derivative equal to zero, one finally obtains for the maximum difference

$(\Delta R)_{max} =$

$$(1+\alpha)\{1-[(1+\alpha)\sin^{-1}(1/(1+\alpha))]^{-2}\}^{\frac{1}{2}} - \frac{\cos^{-1}\{[(1+\alpha)\sin^{-1}(1/(1+\alpha))]^{-1}\}}{\sin^{-1}(1/(1+\alpha))}$$

This maximum deviation occurs when

$$\rho(\tau) = (1+\alpha)\{1-[(1+\alpha)\sin^{-1}(1/(1+\alpha))]^{-2}\}^{\frac{1}{2}}$$

In particular, for $\alpha = 0$, $(\Delta R)_{max} = 0{\cdot}211$, occurring when $\rho(\tau) = 0{\cdot}771$, and for $\alpha = 1$, $(\Delta R)_{max} = 0{\cdot}018$ occurring when $\rho(\tau) = 0{\cdot}594$. As α becomes large $(\Delta R)_{max}$ becomes approximately $(9\sqrt{3})^{-1}(1+\alpha)^{-2}$ which occurs when $\rho(\tau) = 1/\sqrt{3}$. Thus for $\alpha = 5$, $(\Delta R)_{max} \simeq 0{\cdot}00178$.

H. KAUFMAN,
McGill University

G. E. ROBERTS,
RCA Victor Co. Ltd.,
Research Labs.,
Montreal.

REFERENCE

1. BAUM, R. F. The Correlation Function of Smoothly Limited Gaussian Noise. *IRE Trans. Information Theory, IT-3.* 193 (Sept., 1957).

Transistorized Frequency Divider Using a Phase Shift Oscillator

DEAR SIR,—The circuit shown in Fig. 1 was developed in an endeavour to improve upon existing frequency dividers from the point of view of simplicity, reliability, and purity of output waveform. It belongs to the class of locked oscillator dividers, but has the advantage of having no inductors like blocking oscillators and, as its output waveform is nearly sinusoidal, it presents no filtering difficulties which often arise because of the distorted output waveform generated by multivibrators, blocking oscillators, etc.

The divider is basically a phase-shift oscillator and has been used to divide by factors of up to five with perfect reliability over an ambient temperature range of 15° to 35°C, with supply voltage variations of ±10 per cent. It is probable that reliable division by factors of up to about ten could easily be achieved. It does, of course, have the defect, common to all locked oscillator dividers, of giving a spurious output if the locking signal

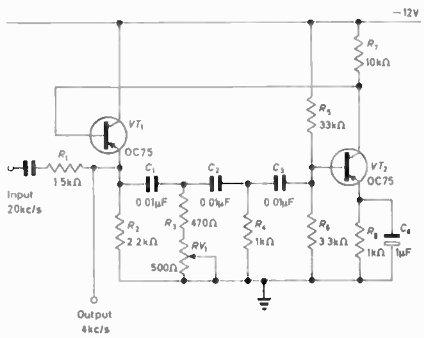


Fig. 1. The frequency divider

fails. Although the regenerative divider[1] overcomes this it is at the expense of greater complexity. The interesting parallel-T divider, described by Butler[2], was found to be unduly sensitive to temperature changes, while the present circuit has fewer components and a purer output waveform.

In the circuit as shown, which divides 20kc/s by five. the input needs to be about 0·2V for reliable locking, but considerable variation is possible before it slips out of lock. The output voltage is about 2V, and is thus ample to drive another divider of a chain without intermediate amplification.

If VT_2 is changed to an AFZ12 and $C_1 = C_2 = C_3 = 0{\cdot}002\mu F$ then the circuit divides 100kc/s by five with no other component changes.

Yours faithfully,

V. B. GERARD,
Physics and Engineering Laboratory,
Department of Scientific and Industrial Research,
New Zealand.

REFERENCES

1. BUTLER, F. A Regenerative Modulator Frequency Divider Using Transistors. *Electronic Engng. 31,* 72 (1959).
2. BUTLER, F. Transistorized Frequency Divider Using a Switched Parallel-T Oscillator. *Electronic Engng. 33,* 664 (1961).

A Method of Driving Uniselectors by Transistors

DEAR SIR,—A circuit was required which was capable of driving up to four 75Ω, 50V uniselectors in parallel. The circuit described can supply current pulses of at least 2·5A at a supply voltage of 50V. The circuit is insensitive to variations of the 50V supply and pulse length is almost independent of the load impedance and the transistor parameters.

A switching type transistor is coupled to a compound pair connected in the Darlington configuration to form an emitter coupled monostable multivibrator of novel properties. The transistor VT_3 (see Fig. 1) is a line output transistor type AU101 which has $I_{c(max)} = 10$A and $C_{ceo} = 120$ V. The emitter coupled circuit has two coupling links in the feedback loop: one of these (the direct connexion of VT_1 to VT_2) is at a high impedance level, the other (the capacitor C between the emitters of VT_3 and VT_1 is at a very low impedance level. Thus an enormous asymmetry in the power handling capacity of the two switches, (i.e. VT_1 and the compound pair VT_2 and VT_3) can be tolerated for monostable operation. Under quiescent conditions the collector of VT_1 is slightly positive; thus VT_2 is cut off and the current in the emitter of VT_3.

$I_e = I_{ceo}\ (VT_3) + (1 + \beta(VT_3))\ I_{ceo}\ (VT_2)$, is of the order of a few milliamperes so $V_e\ (VT_3)$ is approximately zero. A positive pulse at the input will allow the collector of VT_1 to become negative, turning on VT_2 and thence VT_3, and the voltage across the silicon diode MR_1 will rise to about 1·2V. This voltage step at the emitter of VT_3 is passed through the capacitor C and turns VT_1 further off. The circuit is now in its quasi-stable state and the constant current source starts to charge the capacitor C in a linear fashion as shown in Fig. 2. When the emitter of VT_1 becomes positive again VT_1 will start to reduce the base current of VT_2 until finally VT_3 comes out of bottoming and the loop gain becomes greater than unity. The circuit will thus return to its quiescent state, the capacitor discharging with a time-constant equal to R_cC where R_c is the collector load of VT_1.

If the voltage step at the emitter is V volts and the current from the constant current source is I amperes, then the pulse duration will be:

$T = CV/I$sec, where C is in farads.

The non-linearity of the forward characteristic of the silicon diode MR_1 is used to provide high loop gain during the switching interval, but yet to give a sensibly constant voltage step under varying load conditions. Thus the pulse length is almost independent of the number of uniselectors connected to the output terminals at the instant of any particular pulse. The 10Ω resistor shunting the diode is to reduce the voltage due to the leakage current that would otherwise develop across the diode under quiescent conditions.

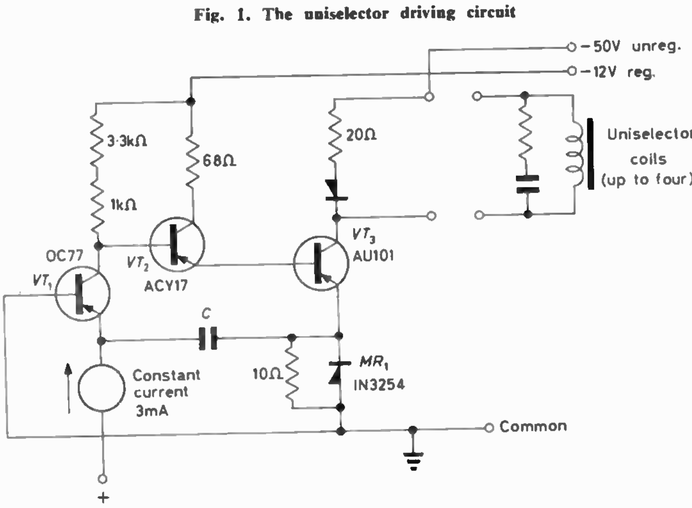
The constant current source in the case of the prototype was a 15kΩ resistor connected from a 48V positive, stabilized line. Where a stabilized voltage is not available the circuit of Fig. 3 can be used and gives similar results.

The maximum voltage allowed on the collector of an AU101 transistor under conditions of slight forward bias ($I_b = I_{ceo}(VT_2)$) will be slightly less than the 120V of $V_{ceo(max)}$, so a damping circuit was designed to limit the back e.m.f. due to the inductance in the circuit under conditions of maximum load to less than 100V. The best compromise between recovery time and allowable voltage swing is usually a combination of resistor capacitor and resistor diode damping. In the prototype a network with a 220Ω resistor and 2μF capacitor is connected across each uniselector coil and a diode and 50Ω resistor is connected to the output terminals of the trigger circuit. Although not optimum this damping circuit gives a recovery time which is much shorter than the recovery time of the trigger circuit which is about $2R_cC$ where R_c is the collector resistor of VT_1.

The trigger sensitivity of the circuit is adjusted by varying R_c and hence the voltage on the collector of VT_1 under quiescent conditions. Care must be exercised to ensure that R_c is low enough to satisfy the condition:

$$I_b(VT_2) \,.\, \beta(VT_2)_{(min)} \,.\, \beta(VT_3)_{(min)} > I_c(VT_3)_{(max)},$$

Fig. 1. The uniselector driving circuit



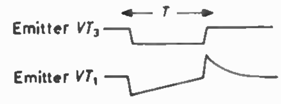


Fig. 2. Circuit waveforms

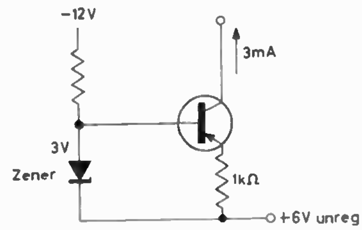


Fig. 3. Constant current source

i.e. there must be enough base current available to VT_2 to ensure bottoming of VT_3 under all load conditions. An alternative way of varying the sensitivity is to leave R_c fixed and vary the current supplied to the emitter of VT_1. The prototype was adjusted for triggering on 3V pulses, and is completely stable under widely varying ambient conditions.

The prototype of this circuit is being used to drive some of the uniselectors in the memory stage of a cosmic ray observatory taking part in the International Quiet Sun Year. The circuit is however obviously capable of application in many other situations requiring large current pulses with inductive loads.

Yours faithfully,

K. J. Gough,

Institute of Nuclear Sciences,
Lower Hutt, New Zealand.

Negative Input Resistance of Transistors

Dear Sir,—An interesting effect, which does not appear to be widely known, has been observed on some junction transistors. These transistors exhibit the property of negative incremental input resistance when used in a common emitter circuit. This effect has been observed on a small proportion of CV7062 when the h.t. is 50V and the collector load is 600Ω. For a base current range of 0 to 10mA the base to emitter voltage has a maximum within this range.

If the mean base current is enough to drive the transistor into the negative slope resistance region and if sufficient capacitance is connected across the base and emitter, the circuit will oscillate.

Yours faithfully,

R. J. Thresher,

and

R. F. Gowlett,

The General Electric Co. Ltd.,
Middlesex.

BOOK REVIEWS

Transistor Applications

By R. F. Shea. 273 pp. Demy 8vo. John Wiley & Sons. 1964. Price 60s.

THIS book is aimed at the electronic engineer wishing to understand the theory of various transistor applications. It covers the two-port network and its external circuit properties, biasing, analysis of single and cascaded amplifier stages, feedback networks. Then the various amplifier types are considered and the book ends with switching circuits, negative resistance devices and a brief discussion of integrated circuits. Many of the chapters are supported by numerical examples.

In the chapter on single amplifier stages it would have been instructive to include an example illustrating the effect of parameter variations on the external performance of a transistor stage. This would have stressed why negative feedback is so widely used in practice to reduce the sensitivity to parameter variations. But then the Author dismisses the whole subject of feedback as an example of matrix manipulations without discussing the basic effects of feedback on sensitivity, harmonic distortion, driving-point impedance, stability etc.

In the chapter on high frequency tuned amplifiers there is no mention of the likelihood of instability due to internal feedback and the associated problem of neutralization. In chapter 17 the discussion on transistor *RC*-oscillators ends without even deriving the frequency of oscillation and the minimum necessary condition for oscillation, and yet both could have been obtained quite readily from Equation 17·27 of the text.

The noise performance of the field effect transistor is discussed and yet the noise performance of the ordinary junction transistor is completely ignored. The treatment of logic circuits is too brief to be of any practical value.

In the chapter on transient response the terminology is confused. Thus, for example, $i_C(s)$ is used to denote the Laplace transform of the collector current $i_C(t)$. This can imply that $i_C(s)$ is obtainable from $i_C(t)$ by simply substituting s for t, which of course is not the case.

S. S. HAKIM.

Electromagnetic Slow-Wave Systems

By R. M. Bevensee. 464 pp. Med. 8vo. John Wiley and Sons. 1964. Price £6 12s.

THIS book gives a theoretical discussion of the various tubes in which beams of electrons interact with slow electromagnetic waves, used as microwave oscillators and amplifiers. The subject matter includes helix structures, coupling through small holes, cavity chains, the interaction of electron beams with waves in helices and cavity chains, and travelling-wave parametric amplifiers and masers. The treatment is well organized and the material is straightforwardly and readably presented.

The author pays particular attention to the meanings of such terms as "wave", "mode", and "harmonic", which, as he rightly observes, "are often bandied about rather carelessly". It is disappointing, therefore, to find in the introduction that he does not make clear, as he purports to do, what the distinction is that he draws between a wave and a mode. This is a small point, however, which does not detract seriously from the value of the book.

The author also refers, in the preface, to his "uncommon notation, which has the advantage of preciseness at the cost of being somewhat cumbersome". In this he does himself less than justice; the notation was found to be clear and not at all cumbersome. It was rather startling to see "phase shift" used as a verb.

The book will be of interest to research and development engineers and designers working in the field, for whom it can be thoroughly recommended as a valuable work of reference; newcomers to the field will find it a useful introduction. The treatment is at a level suitable also for undergraduate students, but these readers would only require a fraction of the material for examination purposes, and so would be better recommended to borrow the book from a library than to buy it.

R. A. WALDRON.

Frequency of Self Oscillation

By Janusz Groszkowski. 530 pp. Med 8vo. Pergamon Press. 1964. Price £5

IT is unusual to see such a detailed and comprehensive treatment of a single topic concentrated in a single volume. With the bibliography, it covers over five hundred pages, and it represents a considerable part of the life work of Professor Groszkowski, who has been publishing papers on this subject since the 1920s. The book gathers together his original work on the non-linear theory of oscillation.

While the latest techniques are treated (for example, tunnel diode oscillators and other semiconductor circuits) the very completeness of the study means that a number of the devices which are described and analysed in detail in the text are historical in nature, and their value is consequently academic rather than practical. In fact the book, while it is an accurate and useful treatise on the theory and practice of oscillation has a slightly nostalgic quality, representing, as it does, a forty year span of study and research by its writer. The author restricts himself to the lower end of the radio frequency spectrum and microwave techniques are scarcely mentioned other than references in the bibliography. The mathematical treatments are clear and readable and there are detailed descriptions of a wide variety of devices, covering both the theoretical analysis of their method of operation, and the practical arrangement of their circuits and construction. The author clearly takes a delight in attention to detail.

A particular feature of the book is its comprehensive bibliography; this is arranged chronologically by years and alphabetically by author within the year. This is irritating at first, but individual subdivision by subject would be almost impossible without cross-referencing and overlapping.

In conclusion, one feels that this book is not likely to be particularly useful to the general worker in the electronics field, but may serve the specialist, and would certainly be a valuable addition to any radio reference library.

J. W. SUTHERLAND.

Digital Magnetic Recording

By A. S. Hoagland. 154 pp. Med. 8vo. John Wiley & Son. 1964. Price 60s.

AS the author states in his preface, this book is not intended to be read as a general background to the subject of magnetic recording. It is, in fact, a comparatively detailed analysis of the techniques of digital magnetic recording and apart from a few introductory passages, concentrates entirely on the digital aspect.

The first chapter deals with the basic characteristics required of the magnetic materials and the read and write heads; and introduces the reader to the idea of storage density. The different modes of recording (i.e. longitudinal, vertical and transverse) are mentioned briefly and the chapter ends by delineating the parameters required for digital data storage.

The next chapter, entitled Mass Storage, explores the requirements for computer storage and includes an historical section which compares the capacity of various computers over the last decade. Comparison is made between sequential access and random access. A short review of magnetic recording technology is included and "state of the art" figures are given. A very brief mention is made of non-magnetic media and the chapter ends with a section on future trends.

The rest of the book deals with the theory of magnetic recording and is divided into four chapters. The first of these is called principles of magnetics and after an historical background goes on to derive magnetic field and boundary value relations. Some knowledge of vector notation is required. A short treatment of the physical basis of ferromagnetic behaviour concludes the chapter.

This is followed by an analysis of magnetic head—magnetic surface interaction phenomena which leads to a section on magnetic head design and magnetic storage surface characterisation.

The last chapter, called Digital Recording Techniques, relates the digital magnetic theory of the previous chapters to bit density and deals with the various forms of pulse coding. There is a section on readback techniques and equalization of the recording channel and the book concludes with a brief word on redundant codes.

For anyone requiring a detailed insight into digital magnetic recording theory, this book together with the chapter references should provide a very good initiation.

R. COUZENS.

Energy Band Theory

By J. Callaway. 357 pp. Med. 8vo. Academic Press. 1964. Price 80s.

THIS book aims to give a comprehensive survey of the procedures employed in band structure calculations, and is of the high standard to be expected from an acknowledged authority on this subject. It is broadly divided into three parts. The first is a discussion of various approximational methods which have been employed to solve the Schrödinger equation for the appropriate crystal potential. The second part describes the results of experimental and theoretical determination of the band structures of particular materials, and the third part deals with point impurities and the effect of external fields.

The book is remarkable for the care which has been taken to give proper weight to the various topics. Each subject, once taken up is treated concisely and fully, and rival techniques are discussed impartially. All the well-known approximational approaches such as the OPW, APW, cellular and variational methods are discussed in appropriate detail, as are the Hartree-Fock equations and the determination of the crystal potential. The chapter describing the results for actual materials is largely an inventory of mathematical papers but is valuable for reference purposes. The book concludes with quite detailed discussions of the problems of point impurities, magnetic field and optical effects. Finally a formidable bibliography, occupying twenty pages, serves to remind the reader of the scholarship needed to write a book of this sort.

It would be wrong to suggest that all this is suitable for general reading. The emphasis is primarily upon the mathematical techniques employed in the calculations, and these assume a familiarity with the language of analysis beyond that acquired in a normal physics or engineering degree course. Furthermore the principles, methods and results of energy band calculations are not needed by the practising electronic engineer. Many physicists will be content with the principles and the results and leave the methods to the mathematician.

Professor Callaway's book can be thoroughly recommended to mathematicians, and to metallurgists and solid state physicists with an interest in band structure.

J. R. ACTON.

The Theory of Electromagnetism

By D. S. Jones. 807 pp. Med 8vo. Pergamon Press. 1964. Price £5 5s.

THIS book is published in an international series of monographs on Pure and Applied Mathematics and approaches the problems of electromagnetism from the viewpoint of a mathematician. Nevertheless, the topics in electromagnetism which are dealt with are those of current concern in Physics and Electrical Engineering and workers in these fields will find the book an invaluable guide to the methods used in present-day mathematical papers on their subjects.

The author, Professor of Mathematics at Keele University, has aimed to produce a text which would take the student from a first acquaintance with Maxwell's equations to within striking distance of the frontier of modern research.

The first hundred pages, starting from Maxwell's equations, develop the various representations of the electromagnetic field suitable for different problems. The scalar, vector and Hertz potentials and integral forms are introduced, and coordinate systems are discussed in unusual detail. After a chapter on Special Relativity including a section on the electrodynamics of moving media, there follow three chapters on radiation, cavity resonators and the theory of waveguides. Among the topics covered are the theory of the cylindrical aerial, the effect of conductivity and boundary perturbation on a resonant cavity and the use of scattering matrices in the analysis of waveguide junctions.

The next five chapters concern wave propagation, and include refraction in the earth's atmosphere, surface waves, scattering by smooth objects and objects with edges, and generalization of harmonic wave solutions for aperiodic disturbances. The final chapter discusses polarization in terms of the electron theory and also deals with fluid motion, shock waves, magneto-hydrodynamics and plasma dynamics.

The coverage of this 800 page monograph can be seen to be wide and to correspond to a remarkably large fraction of the present frontier of electromagnetic theory. (One omission noticed was the modern development of the concept of of the angular spectrum of plane waves.) The author has largely succeeded in his aim of linking the graduate's bridgehead to this frontier, although it should be said that the student of mathematics, accustomed to starting from theorems rather than the physics of a problem, will find the path easier than the physicist or engineer.

The text is exceptionally well annotated with original references, and problems are included at the end of each chapter.

E. D. R. SHEARMAN.

Progress in Semiconductors—Vol. 8

Edited by A. F. Gibson and R. E. Burgess. 244 pp. Med. 8vo. Temple Press Books Ltd. 1964. Price 70s.

The normal practice in this series is to publish in each volume seven or eight critical reviews, written by international authorities.

It was evident, however, when an article was commissioned on transport in semiconductors, that the usual length was not appropriate. Accordingly, Volume 8 consists of one long article giving a critical survey of all aspects of The Electrical Conductivity of Germanium.

Problems in Electronics

By J. C. Higgins. 288 pp. Demy 8vo. Edward Arnold. 1964. Price 40s. Paper Cover 24s.

This book is intended for degree, Dip. Tech., H.N.C. courses, graduateship courses for professional institutions, and the City and Guilds Telecommunications courses. Each chapter comprises a summary of important laws, definitions and formulae, one or more worked examples, and a number of problems for the student to work through either alone or during tutorials. The last chapter contains examination papers taken from the Graduateship examinations of the Institution of Electrical Engineers, the Institution of Radio and Electronic Engineers, and the Institute of Physics, and examinations set for the Diploma in Technology and the Higher National Certificate. Many of the problems in the book are drawn from examinations of the institutions mentioned above, London University B.Sc. and M.Sc. examinations and the City and Guilds of London Institute radio and telecommunications examinations.

Earth Resistances

By G. F. Tagg. 258 pp. Med. 8vo. George Newnes. 1964. Price 55s.

The first part of this book deals with the specific resistance or resistivity of the soil itself, which is of importance in two ways. Measurements of this quantity under specified conditions can give information as to the nature of the soil beneath the surface, and can also detect the presence of underground features, such as caverns, tunnels, faults, etc.

The second part of the book deals with the design of earth electrodes, the various instruments used to make the required measurements and the special problems encountered in the measurements.

Although a fairly advanced grade of mathematics is necessarily employed, the final and practical information is given in the form of curves.

Amateur Radio

By F. G. Rayer. 191 pp. Demy 8vo. Arco Publications. 1964. Price 30s.

This book covers the field of short-wave listening and amateur transmitting, and includes sections on receivers, aerials, amplifiers and modulators, propagation, test equipment and transmitters.

Practical working circuits are provided throughout.

ELECTRONIC EQUIPMENT

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

(Voir page 645 pour la traduction en français; Deutsche Übersetzung Seite 652)

DEWPOINT THERMOMETER

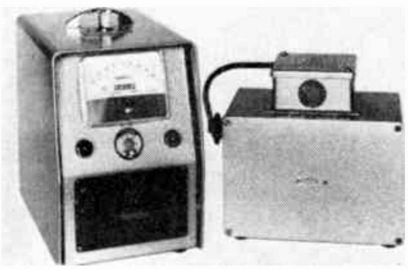
Shaw Moisture Meters, Rawson Road, Westgate, Bradford 1, Yorkshire

(Illustrated below)

This new dewpoint measurement unit gives absolute readings of the ambient and dewpoint temperature of air. It has been specially designed to indicate, on a meter dial, air and dewpoint temperatures for meteorological purposes.

The indicator unit is intended to be positioned indoors and the measuring unit can be up to 40ft (12m) away from the indicator installed in a Stevenson Screen (distance must be stated when ordering).

The accuracy of the dewpoint temperature reading is inherent in the design. The light from the lamp, operated from a stabilized voltage supply is reflected on to a photocell when dew is formed on a mirror surface. The current



to the thermo-module used to cool the mirror then ceases, so that the mirror surface, of which the temperature is measured, continually cycles between the dew and no dew conditions.

This prevents drift in any of the circuits from causing cooling of the mirror below the dew point temperature. An indicator light on the front panel is used to indicate that the dew-no dew, condition is being maintained. This method of operation is considered a significant technical advance over the servo systems which have been used previously in the automatic measurement of dewpoint temperature. Solid state components are used throughout.

EE 73 751 *for further details*

MICROWAVE TERMINATIONS

Morganite Resistors Ltd, Bede Trading Estate, Jarrow, Co. Durham

In a new range of small size high power microwave terminations Morganite Resistors Ltd can now supply aluminium-finned terminations which can be pressurized to 40lb/in^2 working. These type 'A' terminations are available for WG 16 (WR 90) with a mean power rating of 500W at a maximum surface temperature of 300°C. The v.s.w.r. from 8·4 to 10Gc/s is 0·95 (1·05) or better and the overall dimensions are only 6½ × 3½ × 4in. Similar terminations are being developed for the range of waveguide sizes from WG 10 (WR 284) to WG 18 (WR 62).

For lower power applications the type 'M' series can be supplied. These loads have a metal coated finish and in the WG 16 size a power of 350W can be dissipated with an approximate surface temperature of 600°C. The type 'W' range of water cooled terminations is designed for the very high power applications; in the WG 16 size an average power of 1 500W can be dissipated.

EE 73 752 *for further details*

400c/s POWER UNIT

Ashgrove Instruments Ltd, 96 Amyand Park Road, Twickenham, Middlesex

(Illustrated below)

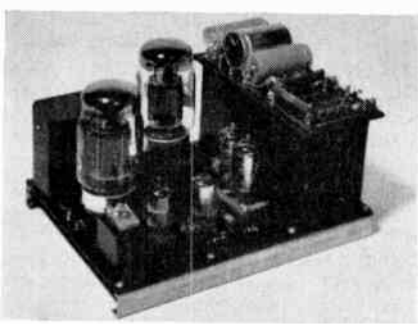
This power unit has been designed to supply synchros, magslips, servomotors, tachogenerators, pick-offs, and all equipment that requires a 400c/s supply.

It is possible to set the output resistance control so that the output voltage rises, remains constant, or drops when the load is applied. The voltage drop in supply lines, fuses, and meters can be offset in this manner if required.

Approximate power factor correction is required, but loads that fluctuate rapidly in demand or power factor will not upset the unit.

The standard unit is supplied in the chassis form, but if required can be supplied for rack-mounting or in a case complete with voltmeter, ammeter, etc.

The output voltage is 115V or 57V adjustable ±10V, the output power being 75VA.



EE 73 753 *for further details*

OSCILLOSCOPE CALIBRATOR

Telequipment Ltd, Chase Road, Southgate, London, N.14

(Illustrated above right)

A portable oscilloscope calibrator, type C1, has been introduced by Telequipment Ltd. It provides in one unit all waveforms and frequencies necessary for the complete alignment of Telequipment and other makes of oscilloscopes, and operates from fully stabilized power supplies.

Weighing 24 lb and measuring 13in by 6½in by 13in, the type C1 has four separate outputs available which may be used independently or, without interaction, concurrently. Using comparison techniques, the calibrator can also test other signal sources.

Signals available are a square wave switched at either 100kc/s or 1Mc/s repetition rate, a square wave switched at either 10kc/s or 1kc/s repetition rate, plus time marker pulses at switched rates of 1Mc/s, 100kc/s, 10kc/s, 1kc/s, 50c/s, and a timing comb, negative going with respect to earth.

The calibrator incorporates a crystal source with an accuracy of 0·2 per cent. A non-interlaced television waveform is available, switched positive with 200 lines approximately; amplitude 1V (sync + video).

EE 73 754 *for further details*

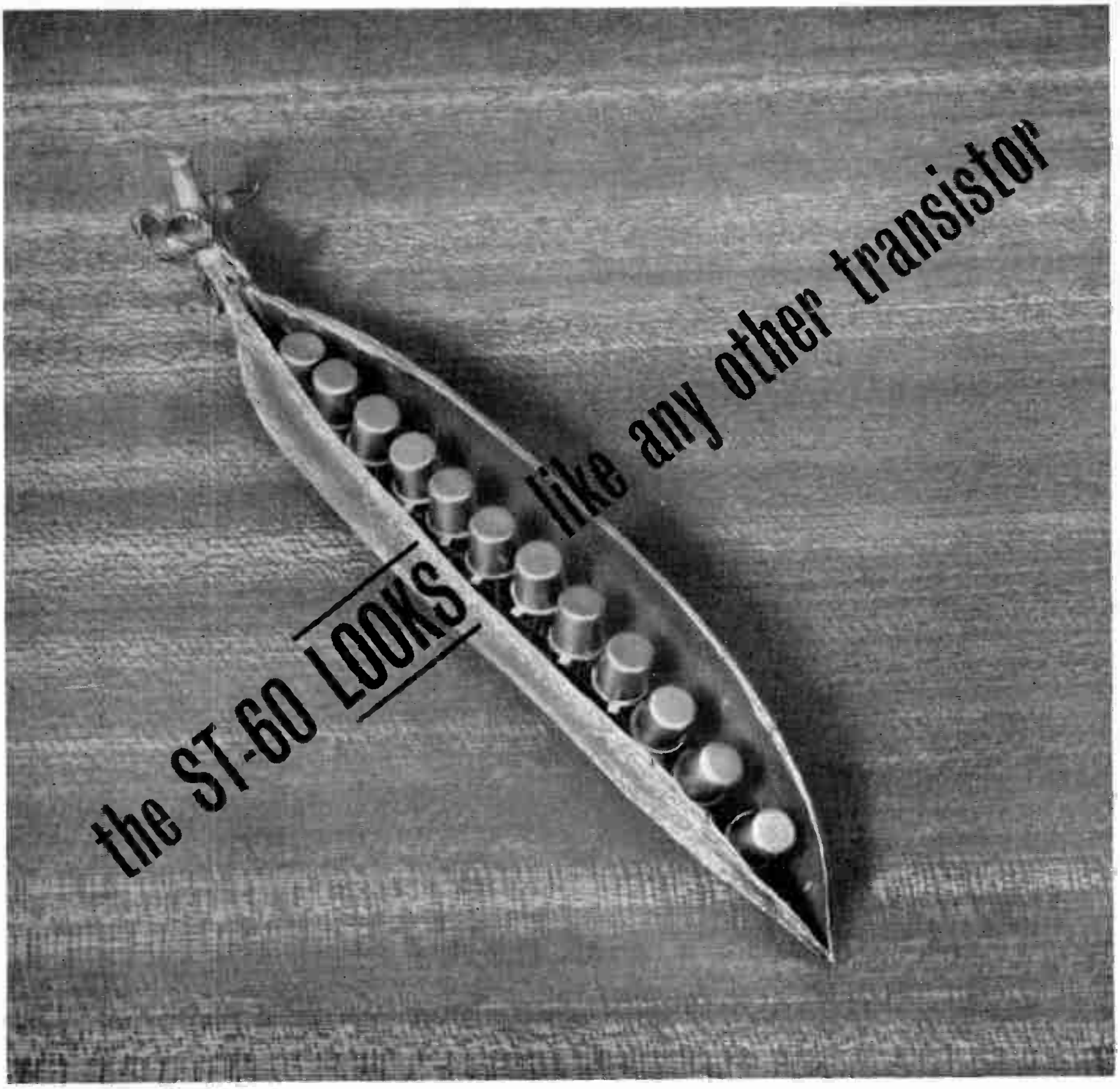
SUB-MINIATURE REED SWITCH

Flight Refuelling Ltd, Wimborne, Dorset

(Illustrated on page 637)

The Hamlin MSRS-2, a subminiature magnetic reed switch designed specifically for low level logic switching, is now available from Flight Refuelling Ltd.

A new silver alloy contact material is used to provide a uniform contact resistance throughout millions of operations, and in addition the high drop-out to pull-in ratio simplifies the design of basic AND logic circuits where the omis-



In fact, the ST60 Series are 'second generation planar, epitaxial transistors, characterised by the following typical parameters :

f_T of 550 Mc/s
C_{ob} of 3pF
t_s of 8 nanoseconds
Interdigitated design
All-aluminium construction for long term reliability

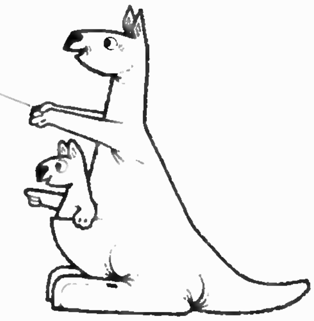
ST-60
SERIES

PLESSEY GROUP

Semiconductors Limited

CHENEY MANOR, SWINDON, WILTS.
Telephone: Swindon 6251.

EE 73 151 *for further details*



Honeywell Visicorder takes the strain for BBC 2 Midlands aerial

In preparation for the BBC-2 Midland service, Marconi engineers are investigating the behaviour of an existing 750 foot mast. This is in connection with the installation of an additional aerial and it is being carried out in accordance with BBC contracts. A Honeywell 1706 Visicorder oscillograph is being used in vital tests to establish the natural resonant frequency of the top of the mast as it oscillates in the wind. An anemometer and three strain gauges feed information to four independent channels of the Visicorder; by comparing the resulting traces with a known simulated strain, the oscillatory strains at various wind speeds can be determined. A Honeywell 1706 Visicorder oscillograph was chosen for this important task because of its high sensitivity, its versatility, and its ability to produce traces for immediate analysis without processing. A Honeywell Visicorder can help solve *your* research problems economically too.

Find out more from

Honeywell

SCIENTIFIC INSTRUMENTS

HONEYWELL CONTROLS LIMITED
Greenford Middlesex *Waxlow 2333*

HONEYWELL INTERNATIONAL *Sales and Service offices in all principal cities of the world. Manufacturing in United Kingdom, U.S.A., Canada, Netherlands, Germany, France, Japan.*

sion of one contributing condition will cause drop-out, or prevent pull-in, depending upon the state of the circuit.

Tests indicate a pull-in to drop-out ratio of 66 per cent to 75 per cent and a maximum contact resistance of 150mΩ after 100 million operations. The switch is rated at 0·5W d.c. with a maximum current of 10mA. The switch can be supplied with a sensitivity of 20 to 100 At according to requirements.

The switch has an overall length (including leads) of 2·25in, the length of the switch body is 0·80in, and the diameter of the glass envelope is 0·090in.

EE 73 755 *for further details*

SWEEP GENERATOR

KLB Electric Ltd, 335 Whitehorse Road, Croydon, Surrey

(Illustrated above)

The PACO G.32 sweep generator and marker adder is a frequency modulated signal generator with a centre frequency range of 3 to 213Mc/s in 5 overlapping bands. The sweep width is adjustable from 0 to 30Mc/s on the high range.

This instrument is suitable for aligning wide-band amplifiers of all types, a desirable feature being that markers are added after the signal has been through the component under test. Due to this, marker signals will not cause misleading results, as is often the case with this type of generator.

EE 73 756 *for further details*

CARBON MICROPHONES

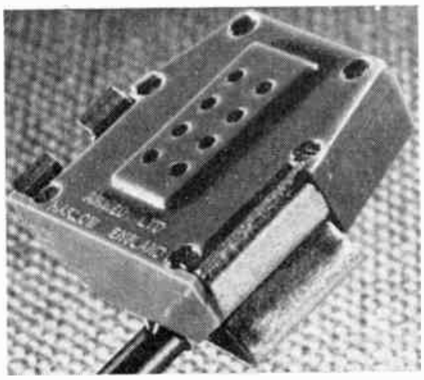
Airmed Ltd, Edinburgh Way, Harlow, Essex

(Illustrated above right)

Airmed Ltd has introduced two new carbon microphones specially designed to meet the needs of modern communication systems where cheapness, small size and reliability are of prime importance.

The type A.301 is a miniature pressure microphone with a sensitivity of −50dB/V/dyne/cm² at 1kc/s. It weighs less than 10g and can be mounted on a headset boom where light weight and small size are of importance. Frequency response has been specially shaped for the best speech intelligibility and covers 200 to 5 000c/s. It is flat from 200 to 1 000c/s then rising 8dB to a broad peak between 2 000 and 4 000c/s, and falling off rapidly above 5 000c/s.

The type A.303 (illustrated) is a carbon pressure differential noise cancelling microphone with high output designed for close speaking under conditions of high ambient noise. As a pressure differential microphone it responds to sounds originating from a small nearby source—the lips—at the same time discriminating against sounds originating from a distance—extraneous noise. The sensitivity is −50dB/V/dyne/cm² at 1 000c/s and response is smooth from 200 to 4 000c/s with a broad rise in the region 2 000 to 3 000c/s and falling away rapidly above 4 000c/s. The high impact injection moulded case has a special clip contact assembly for mounting on a headset boom.



EE 73 757 *for further details*

SENSING RELAY

Leland Instruments Ltd, 145 Grosvenor Road, Westminster, London, S.W.1

(Illustrated below)

The new Leland Leroux type RS1 Sensicon contactless measuring relay is a precision sensing and controlling relay of the contactless type. It comprises a moving-coil meter, similar to the Sensitact relay, which operates a self-contained transistorized electronic control system providing for the operation of associated control units when the pointer reaches a pre-selected position on the scale.

The relay is available in two types. Type N is available in six different versions, with sensitivities ranging between 100μA at 6000Ω and 10mA at 1·5Ω. Type S is available in five different versions with sensitivities ranging from 25μA at 6000Ω to 2·5mA at 1·5Ω.

The Sensicon measures 71mm by 38mm diameter and is presented in the form of a cylinder mounted on an octal plug-in base. At the top end of the cylinder which is of moulded insulating material there is screwed a transparent plastic cap through which appears the scale, the pointer and the control index. The adjustment of the index is made by unscrewing the plastic cap and turning the whole electronic circuit assembly in relation to the base. In this way the index can be moved from one end of the scale to the other. If so ordered these units can be supplied with the index adjusted and fixed to the required setting at the factory.

EE 73 758 *for further details*

SCINTILLATION COUNTERS

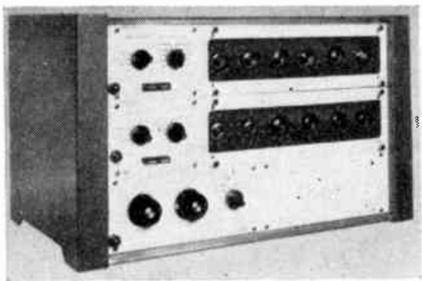
Panax Equipment Ltd, Holmethorpe Industrial Estate, Redhill, Surrey

(Illustrated below)

Two nucleonic counting instruments incorporating transistorized six-decade fast scalers have been added to the Panax range. They are intended as general-purpose laboratory instruments for both Geiger and scintillation counting and each has a maximum counting rate of 50 000 per second. A rotary selector switch enables the counts to be preset to any one of 9 values covering the range of 100 to 1 million. Output signals from the decades are available at coaxial sockets on the rear of the instrument for feeding into a suitable printer for automatic print-out at the end of a counting period. Push-button switches control starting and resetting.

The e.h.t. supply for the Geiger and scintillation counters is generated within the instrument and is fully stabilized against mains voltage fluctuations up to ±10 per cent. It is adjustable from 0 to 2kV by means of a ten-turn precision potentiometer with an accurately graduated control knob. The necessary discriminator bias voltages are also developed within the equipment and are similarly adjustable. A four-position switch is provided for selecting 'dead' times of 4, 20, 200 or 400μsec.

The above description applies equally to both these new instruments, which



will be known under the type numbers P.7602 and P.7702. The main difference lies in the latter instrument (illustrated) which incorporates a six-decade timer as well as the fast scaler, and therefore is more correctly known as an auto-scaler, with fully automatic preset count and preset time facilities.

EE 73 75) *for further details*

ELECTRONIC THERMOMETER

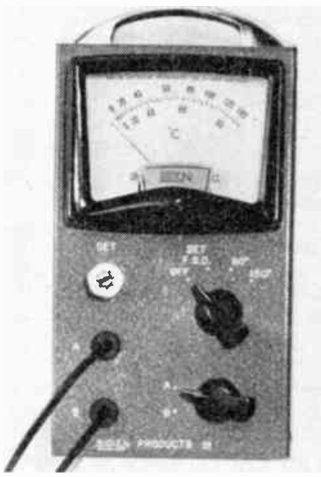
Sidien Products Ltd, 11 Birchwood Court, Edgware, Middlesex

(Illustrated below)

This electronic thermometer has two switch selected ranges of −10°C to 90°C and −10°C to 150°C.

Each thermometer has two probes which are connected to the meter box by 4ft of miniature cable, selection of a particular probe being made by a two-way switch.

The unit is operated by a mallory MN1300T2 cell which has a life in excess of two years.



The thermal time-constant is 20sec and the accuracy of the order of ±2°C.

EE 73 760 *for further details*

VARIABLE TRANSFORMERS

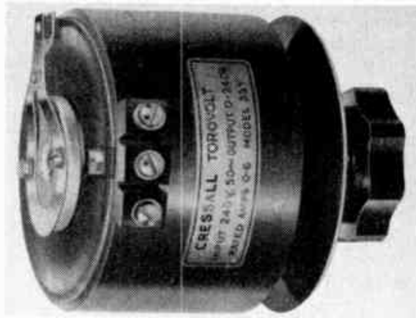
The Cressall Manufacturing Co. Ltd, Cheston Road, Birmingham 7

(Illustrated above right)

An encapsulated auto-wound variable transformer has recently been introduced into the Torovolt range of variable ratio units by the Cressall Manufacturing Co. Ltd.

The new Torovolt. model 33Y, provides an economical and reliable means of controlling a.c. apparatus and for continuously rated loads of up to 0·6A is of the smallest size consistent with long life.

The transformer is designed for direct connexion to a 240V a.c. supply, and enables smooth variations of output to be obtained from zero up to line voltage at a maximum current rating of 0·8A. The wound core is potted in a substantial insulated moulding and has



dimensions of 3 $^{3}/_{16}$in diameter, depth 2 $^{3}/_{16}$in. Connexions are by means of terminals mounted in blocks moulded integrally with the case.

Models can be supplied for front-of-board, or alternatively back-of-board mounting.

EE 73 761 *for further details*

TRANSISTORIZED TELEVISION CAMERA

Automatic Information and Data Services Ltd, 26 Sheen Road, Richmond, Surrey

(Illustrated below)

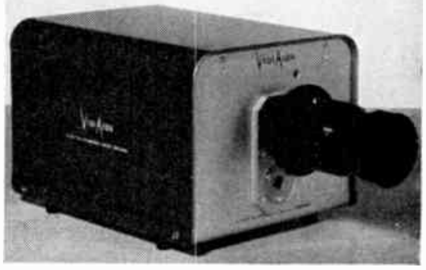
A fully transistorized camera is announced by A.I.D.S. Ltd. The unit is designed to extract maximum performance from a 1in Vidicon, used at a wall anode voltage of below 250V and for fully continuous operation.

Experience has shown that after an initial 'running-in' period, at ambient temperatures between −50° to plus 45°C, servicing is mainly confined to cleaning lenses and replacing the Vidicon tube after 4 000 hours. The camera can be set up to achieve continuous running in higher ambient temperatures, the limiting factor being, in this case, the performance of the Vidicon at elevated heat levels. The camera incorporates 21 transistors and 13 diodes, silicon devices being used where their improved performance, or heat resisting qualities are necessary.

Nominally designed for 625 line, free-running operation, other line standards can be supplied, and 2:1 interlace both to C.C.I.R. standards and A.I.D.S. waveforms can be supplied by an A.I.D.S. waveform generator.

The camera is normally supplied with a power supply contained in a die-cast box, and is fully automatic. A modular type power supply can be supplied with an auto/manual light level switch, for use with other A.I.D.S. modular units.

The camera is capable of resolving 650



television lines in the centre of the picture, and 400 television lines in the corners, where a 50 per cent modulation (black to white), occurs at 330 television lines in the centre and 250 television lines in the corners. (An aperture corrector in modular form can be supplied to increase the centre modulation to 90 per cent).

EE 73 762 *for further details*

PANEL METERS

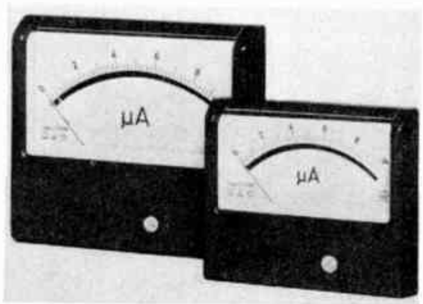
Smiths Industrial Division, Kelvin House, Wembley Park Drive, Wembley, Middlesex

(Illustrated below)

These small panel meters of extreme sensitivity and high precision are the latest addition to the range of recording and laboratory instruments offered by Smiths Industrial Division.

These meters have taut ligament suspended movements with glass pointers giving indication over a very clear scale. The minimum range of the meters is 0—6μA or 3—0—3μA with centre zero.

With horizontally mounted meters having a full scale of 25μA or more, the



accuracy can be as high as 0·5 per cent and a similar accuracy is offered with vertically mounted meters with a full scale of 100μA or above.

The standard size of meter is 72 by 84mm, these meters having a scale length of 63mm.

EE 73 763 *for further details*

T.R. CELL SUPPLIES

Ferranti Ltd, Ferry Road, Edinburgh 5

(Illustrated on page 639)

Ferranti Ltd has developed a new low weight and compact high voltage power supply which provides a 'keep alive' supply for the Ferranti t.r. cell type WF42 (CV2311). It has been designed to overcome the problem of obtaining a suitable high voltage d.c. supply in an aircraft.

Two versions are produced, one having an input voltage of 115V, 400c/s and the other 200V, 400c/s. Both versions have a striking voltage of 1 050V d.c. ±10 per cent which reduces to normal sustaining voltage after striking.

The assembly measures 1·4in by 1·4in by 1·5in with a weight of 2·7oz and is encapsulated in epoxy resin. It meets the Joint Service humidity classification H.6 and DEF spec. 5214 for resin cast

EE 73 152 *for further details*

DM 2005

a new addition

Derived from, and additional to, our existing range of instruments, the DM 2005 Digital Voltmeter offers excellent accuracy and stability with the reliability of proven design. It shares many interchangeable plug-in circuit boards* with our DM 2001, DM 2020 and DM 2022 models and inherits their virtues of long life and ease of maintenance. Manufactured to the highest standards, it is available now

* Epoxy Bonded glass fibre of course!

* 10uV to 2KV with Automatic Polarity Indication.
* Direct Reading Scale of 7999 with 10uV increments.
* Accuracy 0·0125% F.S.D. ± 0·02% of reading.
* > 25,000M Input Impedance.
* Common Mode Rejection, 120dB at D.C., 100dB at 50c/s A.C.
* 6 Operating Modes including full accuracy Maximum and Minimum Modes.
* Choice of 6 Digital Output Codes by Replaceable Plug-in Boards to drive Printers, Punches, etc.
* Price £500.

DM

DIGITAL MEASUREMENTS LIMITED

SALISBURY GROVE MYTCHETT ALDERSHOT HANTS TEL: FARNBOROUGH 3551

transformers. The ambient temperature range is $-40°C$ to $+100°C$, and the assembly has passed the voltage breakdown test at an altitude of 70 000ft.

The power supply will withstand a peak inverse voltage of 4kV for $10\mu sec$ and will not be damaged by an accidental short-circuit of the output.

Varients within the same frame size are possible for use with other Ferranti t.r. cell specifications.

EE 73 764 *for further details*

TORQUE TESTING EQUIPMENT

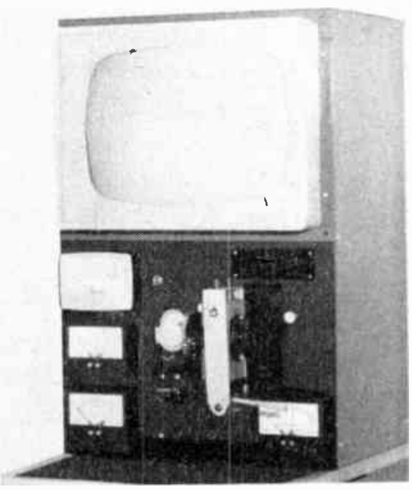
Mining & Scientific Equipment Ltd, 317 Kennington Road, London, S.E.11

(Illustrated below)

M.S.E. Ltd has announced the introduction of a new torque measuring instrument. Designed particularly for batch testing and quality control of fractional horse power motors, the instrument employs a new type of electronically controlled absorption dynamometer which enables torque readings to be accurately measured independently of rotational speed.

The model shown in the photograph has a torque measurement range of 5 to 50oz in, with a claimed accuracy of 1·5 per cent of full scale torque range; other ranges are available. The rotational speed of the motor can also be simultaneously measured and indicated up to 5 000rev/min.

Read-out can be obtained on an open



scale meter, or alternatively displayed on a 17in long-persistence oscilloscope which gives an immediate trace of the speed torque characteristics.

A graticule can be arranged on the face of the c.r.t. in order that the quality of the motor can be readily determined. The use of a long-persistence oscilloscope also allows photographs of the trace to be taken.

EE 73 765 *for further details*

SOLDERING IRON

Distributed by: Oliver Dow Ltd, 877a High Road, Finchley, London, N.12

(Illustrated below)

The 'BLIXT' soldering iron, which is made in Sweden, has in its handle a roll of flux-cored solder which is fed out to the print of the bit by a 'pistol' trigger. Thus one hand only is needed to use the iron and the operator has the other hand free to hold a component or a pair of pliers etc.

The iron is supplied with one spare bit and others are available. The solder is loaded on easily replaced plastic spools which hold 16ft.

EE 73 766 *for further details*

LEVEL MEASUREMENT

Vacuum Reflex Ltd, 6 Soho Street, London, W.1

(Illustrated above right)

Vacuum Reflex Ltd has developed a new range of instruments for continuous level indication and control. The system utilizes ultrasonic waves reflected from the free liquid or loose solid material interface, and is suitable for level measurement on all types of liquid, granular, solids and powders.

Two ultrasonic transducers, one a transmitter, the other a receiver are used mounted above the maximum level of the tank or bunker contents. To determine level, the time taken for the reflected pulse to return from the interface to the receiver is measured, and presented on a meter calibrated in level or tank contents. All external units are encapsulated and the entire equipment is transistorized, silicon or planar transistors only being incorporated.

With the transducers mounted above the tank contents, measurement of level depends only on reflection of ultrasonic waves from the surface of the material and, hence, the system is independent of the nature and condition of the contents of the tank or storage vessel. Materials which can be measured include water, milk, coal, ash, stone, rock, salt, etc. The system is also suitable for

measuring liquid/liquid interfaces, has been used for liquid/foam interfaces and may be used inside high pressure vessels.

Instruments are at present available having ranges up to 40ft (100ft in liquids) but development work is proceeding to extend the ranges.

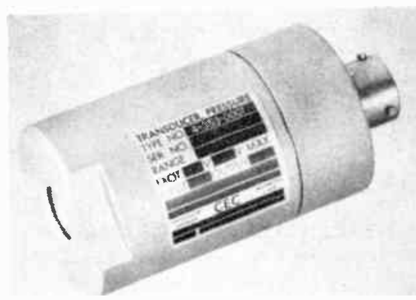
EE 73 767 *for further details*

PRESSURE TRANSDUCER

Consolidated Electrodynamics Division, Bell & Howell Ltd, 14 Commercial Road, Woking, Surrey

(Illustrated below)

A low-range pressure transducer designed for applications where small size and light weight are a requirement has been introduced by the Consolidated Electrodynamics Division of Bell & Howell Ltd.



The type 4-353 pressure transducer is claimed to be the smallest unit on the market in its absolute pressure range of 0 to 1 lb/in^2. It weighs seven ounces and is 2½in long and 1¼in in diameter.

The new transducer was designed for measurements of airborne high-altitude pressures, altitude chamber pressures, wind-tunnel pressures and exhaust pressures of steam turbines. Its first application was a survey of pressures generated by the firing of a rocket in an altitude chamber.

The 4-353 uses a new diaphragm design which achieves high efficiency in minimum space and size. All compensation is accomplished externally to the sealed active element. A mechanical over-pressure stop allows the instrument to tolerate 20 lb/in^2 without damage.

Operable temperature range is $-100°F$ to $+275°F$, sensitivity is 20mV, thermal sensitivity shift is within $\pm 0{\cdot}005$ per cent/°F over the compensated temperature range, and combined linearity and hysteresis is $\pm 0{\cdot}5$ per cent or better of all range output.

EE 73 768 *for further details*

KLYSTRON POWER SUPPLY

Distributed by: Miles Hivolt Ltd, Old Shoreham Road, Shoreham-by-Sea, Sussex

(Illustrated on page 640)

This unit type LS525R, which is

manufactured by Oltronic, of Sweden, is a universal power supply which is suitable for use with the majority of klystrons now available.

The following supplies are provided:

Beam voltage	−200V to −3·6kV
Reflector voltage	0 to 1kV, 50μA
Grid voltage	−300V to +150V 5mA
Heater voltage	6·3V, 3A

All outputs have excellent stability and low ripple; the reflector voltage having less than 200μV ripple.

Four modulation generators are built in and provide square waves, pulses, sawtooth and sine waves.

EE 73 769 *for further details*

SONIC GAUGE COMPARATOR

Westland Aircraft Ltd, East Cowes, Isle of Wight

(Illustrated below)

A portable comparator for use with vibrating wire sonic strain gauges has been designed and produced by Westland Aircraft Ltd. It has been developed in connexion with the company's contract to install sonic and foil strain gauges at the Oldbury-on-Severn nuclear power station.

The comparator has been designed with particular emphasis on simplicity of operation and indication. The controls are limited to an on-off switch, together with terminals for connecting the gauge under test. The instrument reads sonic vibration frequency directly by comparison with an internal quartz crystal, and frequency is indicated on a four-digit projection display. A source of energy for plucking the gauge wire is included in the unit together with drive circuits to make the measurement process completely automatic. On connecting a gauge to the comparator, the gauge is plucked, measured and the reading displayed; this process being repeated at four-second intervals.

The instrument is suitable for outdoor use in construction site conditions, being housed in a robust portable case with a lid protecting the controls from damage during transit. It is completely self-contained, including its own lightweight rechargeable battery, with facilities for recharging from an external charging unit. An internal stabilizer allows correct operation to be maintained over a wide range of battery conditions.

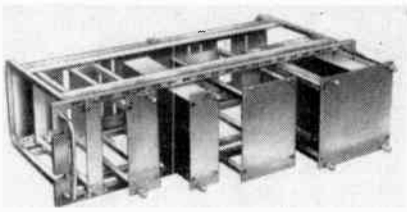
EE 73 770 *for further details*

CONSTRUCTION SYSTEM

A.P.T. Electronic Industries Ltd, Chertsey Road, Byfleet, Surrey

(Illustrated below)

A.P.T. Electronic Industries Ltd has introduced a new modular chassis construction system intended for electronic equipment, and marketed under the name MINAR. The system was designed by the BBC by whom it has been used experimentally for some time, and it is based on a standard 19 by 5¼in panel in the form of a miniature rack, into which units of various sizes may be



inserted. The system is particularly suitable for transistor or printed circuit assemblies.

The principal mounting assembly of the system is the frame assembly, which is intended for fitting directly into a 19in rack. The frame assembly is drilled and tapped to accept modules of various widths, all of which are multiples of the basic unit width of ⅞in. Modules up to 20 times this basic width may be fitted in any order in one frame assembly, and there is provision for permanent fixing of modules in the frame assembly or for making them readily withdrawable. as desired. Dimensions are arranged so that, once the first module is fitted, the fixing screws on other modules automatically line up with the corresponding holes in the frame.

The system includes a case and lid, into which a frame assembly may be fitted for portability, and covers for enclosing individual units or complete assemblies. Other items available are chassis and printed circuit boards of the correct size for fitting directly into the modules.

Most items in the system are supplied ready assembled, but in order to reduce the cost of despatch, the largest modules are supplied in kit form; the only tool required for assembly is a screwdriver, however, and no special skill is called for. The individual components from which frame assemblies, modules, etc., are constructed, are also available separately, so that users may make up special items to their own requirements, at minimum cost.

EE 73 771 *for further details*

ELECTRONIC MULTIMETER

Comark Electronics Ltd, Gloucester Road, Littlehampton, Sussex

(Illustrated below)

The type 130 electronic multimeter combines the versatility of the multimeter with the sensitivity of the valve-voltmeter. As it is battery operated and uses transistors throughout, it is completely free from hum and earthing problems, is very compact, and requires no warm-up time.

The instrument has 54 basic ranges, extending from 10mV to 300V and 10μA to 100mA a.c. and d.c. and 1μA and 1kV d.c. (all ranges are f.s.d.). with centre zero on all d.c. ranges.

Input resistance is approximately 2MΩ on all a.c. ranges and 1MΩ/V on all d.c. ranges except the highest. The voltage drop at the terminals on all current ranges is less than 12mV. Noise is negligible on all except the most sensitive a.c. voltage range, when it does not exceed 1mV when fed from 100kΩ source impedance. Frequency response is flat to 50kc/s with useful response to 250kc/s.

Resistance measurements are on linear scales, covering 100Ω to 1MΩ f.s.d. in five ranges. In addition there is a high resistance range of 1MΩ centre-scale. These ranges give a total coverage of 1Ω to 100MΩ. Resistance measurements up to 10 000MΩ and down to 0·01Ω may be made using external batteries.

A feature of the instrument is that three-terminal resistance measurements are possible; resistance may be measured directly in the presence of shunting to a third terminal (for instance to earth).

The instrument may be used as a voltage or current amplifier on both a.c. and d.c. ranges. Output is approximately

EE 73 154 *for further details*

NEW

PHILIPS

hf double beam oscilloscope

PM 3230

HIGH SENSITIVITY	LARGE BANDWIDTH
20 mV/div	0 - 10 Mc/s
2 mV/div	0 - 1.5 Mc/s
(1 division = 8 mm)	

Clear, sharp picture provided by new "side-by-side" double-gun 10 cm (4 in) cathode ray tube with full scale horizontal and vertical deflection of both beams.

Clear, logical lay-out of front plate, resulting in quick and easy operation.

19 calibrated sweep speeds from 0.5 μs/div to 0.5 s/div, sweep expansion up to 5×

Jitter free triggering with simple controls stability being preset, special TV frame position

Largely transistorized

Compact and lightweight (11 kg - 24 lbs)

Full set of accessories available

Complete manual with full service instructions supplied with the instrument

Sold and serviced all over the world

Compartment for mains flex provided

Transparant dust cover with pockets for all accessories provided

Low power consumption (70 W) and wide range of supply frequencies (50 - 400 c/s) making it suitable for use with an inverter

Facility for photographic recording, three preset levels of graticule illumination for simple camera setting

All components easily accessible for maintenance and repair

PHILIPS

PHILIPS

electronic measuring instruments

For the U.K.:
The M.E.L. Equipment Ltd.,
207 Kings Cross Road, London WC1

1V for the input selected on the range switch.

The circuit consists of a high-gain a.c./d.c. amplifier driving a 100μA 3·2in mirror scale meter. For d.c. measurements this is preceded by a transistor chopper pre-amplifier, giving very low drift. For a.c. measurements, a high impedance input stage is used. Gain is stabilized by heavy overall negative feedback.

EE 73 772 *for further details*

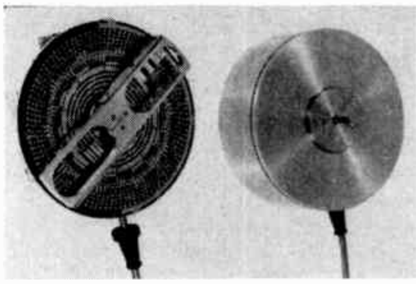
DIGITAL SHAFT ENCODER

Distributed by: B & K Laboratories Ltd, 4 Tilney Street, Park Lane, London, W.1

(Illustrated below)

The Peekel type PP3A1 digital shaft encoder is intended for attachment to potentiometric recorders or similar apparatus in which analogue information relates to the angular position of a rotary shaft. The PP3A1 unit converts this into digital form for direct feed into supplementary read-out or display units.

The PP3A1 unit is attached to the recorder or other instrument by means



of clamps and a mounting plate, and can either be directly driven or coupled to the shaft with cable and pulleys, etc. A special model is available for use with the Brüel and Kjær 2305 recorder.

The output available from the PP3A1 unit is presented in the form of three-decade decimal information from 0 to 999 related to one complete shaft revolution. This output, without further decoding, can be fed directly to a Peekel type PP5C1 digital indicator. It can be used to operate a digital printer through the medium of a Peekel type PP9BA1 print-out convertor, or a tape puncher through the medium of the Peekel type PP9BF1 punch tape convertor. Other Peekel units designed to be used in conjunction with the PP3A1 adaptor include a digital memory, a digital clock, a pulse counter, and input commutators which enable up to 200 channels to be connected to one digital instrument or system.

The illustration shows internal and external views of the encoder.

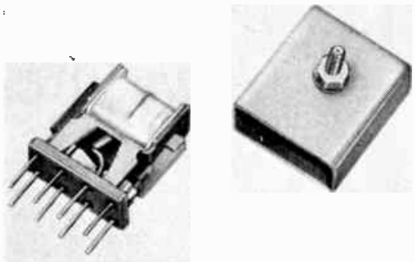
EE 73 773 *for further details*

SUB-MINIATURE RELAY

Plessey-UK Ltd, Abbey Works, Titchfield, Fareham, Hampshire

(Illustrated above right)

A new sub-miniature relay—type CF—from Plessey UK Ltd has received



approval to Ministry Specification DEF 5165. This component is a fully sealed rotary-action two-pole changeover relay with a maximum contact resistance of 0·030Ω ensured by twin gold-plated contacts.

Connecting pins are spaced on the standard 0·1in module for printed circuit mounting and are brought out through glass-to-metal seals. Coil and contact connexions to these pins are arranged symmetrically so that correct operation is ensured whichever way the relay is plugged into its holder.

Contact rating is 1A to 3A at 28V d.c. or 115V a.c. resistive, according to number of operations.

EE 73 774 *for further details*

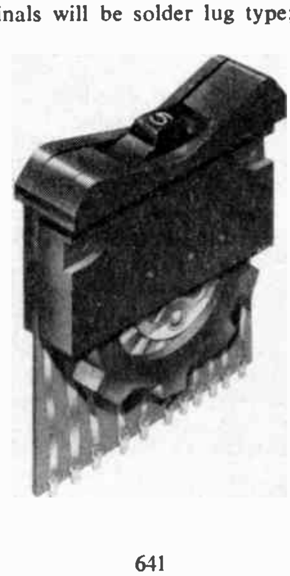
PRINTED CIRCUIT SWITCH

NSF Ltd, 31-32 Alfred Place, London, W.C.1

(Illustrated below)

NSF announce a printed circuit, single pole, 10-position edge control rotary switch, suitable for instrumentation and control apparatus. The switch is available in single units, or ganged units of up to four sections, controlled by a single thumb wheel, or in stacked assemblies of up to a maximum of twelve single units. The indexing mechanism is of double ball type, 36° indexing only and a stop will be included to limit rotation to any position between 2 and 10. Contacts are non-shorting.

The standard printed circuit board provides 1-pole, 10-way switching. Special boards can be tooled to order. Terminals will be solder lug type; alternatively, printed circuit connectors on a 0·2in module can be used.



Slots are provided for universal mounting to suit various panel thicknesses.

Thumbwheels are black phenolic, with white digits, and the bezel a medium grey Delrin. It will be possible to supply in other colours to order.

EE 73 775 *for further details*

MULTI-RANGE ELECTROMETER

Thomas Industrial Automation Ltd, Station Building, Altrincham, Cheshire

(Illustrated below)

Thomas Industrial Automation Ltd has introduced an additional model to their range of electrometers. This instrument is complementary to the VC.99 retaining the same outstanding features and specification, but offers the additional facility of a range doubler switch.

Known as the model VC.99A its function is:

(a) As a micro current meter measuring currents from 1pA up to 600μA in eight switched ranges. The top of the range has been designed to overlap the most sensitive range of the conventional multimeters. The use of heavy degeneration ensures a low voltage drop on current measurements.

(b) As an electrometer voltmeter direct measurements of potentials of either polarity from 0 to 2 000mV in two switched ranges can be made, with an equivalent input resistance of greater than $10^{13}\Omega$.

Because the electrometer can measure extremely small currents it can readily be arranged to measure very high resistances. Using an external potential source, resistances of up to $10^{14}\Omega$ can be measured. Where a 500V test supply is specified, as in some tests for capacitors and cables, a special highly stabilized 500V adaptor is available. Measurements up to 5 by $10^{14}\Omega$ can now be made using this adaptor.

The instrument can be housed in a robust Burma teak case, or a black laminated plastic case each fitted with a carrying handle and removable lid.

EE 73 776 *for further details*

Short News Items

The Electronics Division of the Institution of Electrical Engineers together with the Institution of Electronic and Radio Engineers and the United Kingdom and Eire Section of the Institute of Electrical and Electronic Engineers, is organizing a one-week conference on components and materials used in electronics engineering to be held at the I.E.E. from 17 to 21 May 1965. The meeting will be concurrent with the 1965 Radio and Electronic Component Show at Olympia.

Subjects to be discussed are recent developments in active and passive components, including integrated circuits, and in the materials of which they are made. Thermionic and cold-cathode valves and applications (except in so far as they influence the design of components or the use of particular materials) will not be included.

Contributions of about 1 500 words are invited and authors are asked to submit synopses of about 200 words, together with probable title without delay.

Completed contributions should be sent to the Secretary of the I.E.E. by January 1965.

Further information and registration forms are available from the Secretary, the Institution of Electrical Engineers, Savoy Place, London, W.C.2.

The International Exhibition of Electronic Components (Salon International des Composants Electroniques) organized by the French National Federation of Electronic Industries is to be held at the Porte de Versailles, Paris on 8 to 13 April 1965.

Further details are obtainable from the Fédération Nationale des Industries Electroniques (F.N.I.E.) at their new address:- 16 rue de Presles, Paris 15e.

The Institute of Sound and Vibration Research is holding a three day course on Practical Noise Control at the University of Southampton on 16 to 18 September.

The course is aimed at bringing together professions from many different fields who, at some time or another, may have to consider a problem of noise control, its hazards and implications.

The course will consist of ten lectures, supported by laboratory demonstrations and will cover acoustics, noise measurements, subjective assessments, technical principles of noise control, medical and legal implications.

The course fee is £4 4. 0. and forms of application and further information can be obtained from—Dr. I. Priede, Course Organizing Secretary, Institute of Sound and Vibration Research, The The University, Southampton.

The Electronics Group of the Association of Special Libraries and Information Bureau (ASLIB) has recently published Issue No 3 of its "Handlist of Basic Reference Material for Librarians and Information officers in Electrical and Electronic Engineering."

Copies are available to Aslib members at 6s (non Aslib members 12s) on application to Miss B. Newman, Secretary, Aslib Electronics Group at Ericsson Telephones Ltd, Beeston, Nottingham.

IEC Publication 153-1, the first edition of which has just been issued, forms the first part of the complete IEC Recommendation for Hollow Metallic Waveguides, which relates to straight hollow metallic tubing for use as waveguides in electronic equipment. The object of this Recommendation is to specify the details necessary to ensure compatibility and, as far as essential, interchangeability, to lay down test methods and to prescribe uniform requirements for electrical and mechanical properties. Copies of this Publication 153, Part 1, may be obtained from the BSI Sales Branch, 2 Park Street, London, W.1, price 20s. 3d.

A contract to supply an automatic message switching centre for the communications of North Atlantic Treaty Organization forces in southern Europe has been announced by the International Telephone and Telegraph Corporation Data and Information Systems Division at Paramus, New Jersey, an associate of Standard Telephones and Cables Ltd. The system has been ordered by the Italian Ministry of Defence as the host NATO Nation of the activity. The new installation, which is scheduled to be in operation within 18 months, will be part of an overall NATO high-speed communication network called Telegraph Automated Relay Equipment (TARE).

The Report of the Research Council of the Department of Scientific and Industrial Research on the need for National Priorities in Research has recently been published by Her Majesty's Stationery Office. Reconsideration of the level of support for research in established areas, particularly in those requiring large teams or expensive equipment is indicated by the Council in their Report for 1963.

The British Radio Valve Manufacturers Association (BVA) and the Electronic Valve and Semiconductor Manufacturers Association (VASCA) have recently published the second edition of the Guide to the British Electronic Valve and Semiconductor Industry.

In a foreword it is stated that British production of electronic valves and tubes is in the region of 100 million devices annually (90 million two years ago). The current annual rate of manufacture of semiconductor devices in the U.K. is about 45 million transistors (32 million in 1962), 44 million diodes (28 million in 1962), and 10 million rectifiers, including selenium (compared with five million). Direct exports of the whole industry have been growing steadily and now run at about 20 per cent of production.

Primarily intended as a brief for British Commercial Attaches and Information Officers abroad, the booklet is available post free from the Secretary, B.V.A., Mappin House, 156 Oxford Street, London, W.1.

A new h.f. radio station is to be built at Belize, British Honduras, for Cable and Wireless Ltd. It will cost in the region of £100 000 and take 18 months to complete.

The station will be built on a site at Pine Ridge, which has an area of about 200 acres, and work is expected to start on it in six months' time.

Present plans provide for transmitters and receivers to be in separate buildings about 800 yards apart, the transmitters being remotely controlled from the receiver building.

Four transmitters of 1kW each will be installed and there will be five receivers. The new station will be linked to the Central Telegraph Office in Cattouse Building, Belize, by a multi-channel v.h.f. link.

Circuits to be operated eventually will include telegraph to Jamaica, Mexico and Miami, and telephone to Jamaica, Guatemala and Honduras.

Enamelled and Cotton Covered Round Copper Wire, B.S. 1815, has recently been published. Part 1 'Round Wire' introduces optional requirements and tests for the chemical purity of cotton yarn. These requirements are similar to those specified in B.S. 1791 for cotton-covered copper conductors, and bring the Standard into line with others in the series. Copies of B.S. 1815, Part 1, may be obtained from the BSI Sales Branch, 2 Park Street, London, W.1, price 6s.

The Institute of Physics and The Physical Society is arranging a conference on Elementary Particles to be held at the University of Birmingham on 5 to 7 April, 1965.

The aim of the conference will be to review the current experimental and theoretical situation in the strong and weak interactions of elementary particles. Registration is necessary and further details and application forms will be available in due course from The Administration Assistant, The Institute of Physics and The Physical Society, 47 Belgrave Square, London, S.W.1.

The British Radio Valve Manufacturers' Association (BVA) and the Electronic Valve and Semiconductor Manufacturers' Association (VASCA) have announced the following sterling values of their members' sales during the three months ended 31 March 1964:

Valves and tubes	£12·8M
Semiconductor devices	£5·7M
Total	£18·5M

The total value of the sales for the whole of 1963, published in April last, was £61·6M.

The BBC Research Department Line-Store Electronic Standards Convertor is now able at the turn of a switch to convert television signals either from 625 lines to 405 lines or from 405 lines to 625 lines.

This standards convertor which was developed at Kingswood Warren was originally installed at the Television Centre in time for the opening of BBC-2. It works on a somewhat similar principle to the one-way convertor developed by the BBC Designs Department.

The standards convertor, which is contained in a single bay of equipment 3ft wide, stores one scanning line of the television picture in 576 capacitors. There are 1 152 high-speed switches, each of which can charge or discharge one of these capacitors in one twenty-millionth of a second.

The Post Office Research Station at Dollis Hill will be moved to the vicinity of Martlesham, in Suffolk. The aim is to start the move about 1968.

The Research Station, which employs some 1 400 staff. has occupied its present site at Dollis Hill, London, since 1912. It has been recognized for some time that the Station would have to be moved because the present site is severely congested and there is not enough space to allow for development.

The area chosen is under two hours' travelling time from London, and is near the University of Essex.

The Research Branch of the Post Office Engineering Department has as its main objective the improvement of the efficiency of the communications services of the Post Office by the study of new phenomena, materials, techniques, and development leading to new systems and apparatus. It also co-operates with other Government and industrial laboratories doing similar work, and has liaison with universities on research contracts, advice on research projects for higher degrees and the organization of specialist training.

The International Federation of Automatic Control (IFAC) announces the third Congress to be held in London from 20 to 26 June 1966 on invitation of the United Kingdom Automation Council (UKAC).

Authors who wish to submit a paper should approach the IFAC National member organization of their country, the address of which may be requested from the IFAC Secretary, Postfach 10250, Dusseldorf, Germany. The address in the U.K. is United Kingdom Automation Council, c/o The Institution of Electrical Engineers, Savoy Place, London, W.C.2. Printed invitation to authors with all the details for the presentation of the papers will be available by the end of October 1964. Papers should be submitted to the IFAC national member organization of the author's country early in spring 1965 (the exact date to be fixed by the national member organization). The papers should deal with topics, components, applications or the theory of automatic control. The theory papers should preferably bridge the gap between theory and practice. Some papers will be admitted on biological problems.

The first direct telephone route from Europe to Japan using submarine cables for all its transoceanic routes has recently been brought into service.

Hitherto telephone calls to Japan have been carried in the TAT Cables to the U.S.A. and then by radio to Tokyo. They will now be carried via CANTAT and the trans-Canada microwave link to Vancouver, thence in the COMPAC cable from Vancouver to Hawaii, where COMPAC is linked with the recently completed American/Japanese trans-Pacific cable (T.P.C.), to Japan. Telegraph and telex circuits to Japan will soon be routed in the same way to give an improved 24-hour service. At present the radio circuits to Japan are affected by fading for several hours a day.

The United Kingdom will contribute about £6M to the global communications satellite system under an agreement reached in Washington. This represents 8·4 per cent, comparing with 6·1 per cent each for France and Germany, 3·75 per cent for Canada, 2·75 per cent for Australia and 61 per cent for the United States.

The U.K. contribution may be increased to £9M if additional funds are required. Two agreements, closely inter-related, were reached at the international conference in Washington. These provide that the design, deveopment, construction and establishment of the 'space segment' of the satellite system, i.e. the satellites themselves and the ground installations for their control will be a co-operative international enterprise. The telecommunications ground stations will be owned by the countries, or groups of countries in which they are located. Control will be exercised by an International Committee, comprising one representative from each signatory or group of signatories which has made a contribution of not less than 1·5 per cent of the estimated costs. The U.K. will have a representative on the Committee, which at the outset will consist of 12 members, one each from the United States, Canada, Australia and Japan and the remaining seven from the countries of Europe, in addition to the U.K.

The United States Communication Satellite Corporation will act as the manager for the space system, pursuant to the general policies of the International Committee and in accordance with specific determination which may be made by the Committee.

The Dover Harbour Board has installed closed-circuit television, supplied by Pye Telecommunications Ltd, to assist the control of cars at the Dover Car Ferry Terminal en route to Calais, Ostend and Boulogne. During peak periods 7 000 cars a day pass through the terminal to board ferries arriving and departing every half an hour.

A control room with two 19in television screens capable of showing pictures from six cameras enables the traffic controller to spot any likely bottlenecks and to give instructions over the public address system to maintain an easy flow of traffic.

The cameras are remotely controlled by a telephone dialling system, which automatically selects the required camera for focus, rotation and tilt action.

Drivers arriving at the waiting area are given visual instructions from a large illuminated sign. This shows sailing and loading times as well as destinations. The information board is operated by push-buttons direct from the control room.

More than 50 British firms will be taking part in the British Electronic Component and Instrument Exhibition to be held in Stockholm from 13 to 16 October, 1964.

Promoted by the Radio and Electronic Component Manufacturers Federation, this exhibition is the latest in the series regularly held in Stockholm over the last fifteen years and is by far the largest in firms represented and space occupied.

W. H. Sanders (Electronics) Ltd, of Stevenage, Herts have received orders for 23 of their Language Laboratories.

The equipment has been purchased

by educational authorities including the London County Council, Hertfordshire Education Department and Derbyshire Education Committee. Interest shown by private educational establishments has resulted in the sale of language laboratories to Bradford College of Reading and the Teresian Institute at Kingston-upon-Thames.

The Belfast College of Technology has also ordered an equipment and a language laboratory has been installed in Stockholm University.

EMI Electronics Ltd, in conjunction with its associated company in America, the Hughes Aircraft Company, has been selected as the main contractor for simulators at the Royal Navy Polaris School now under construction at Faslane in Scotland. This school will provide full synthetic training for the R.N. personnel, using a weapon control simulator manufactured by the Hughes Aircraft Company. EMI engineers will assist in the installation, testing and tuning of this device and will also train naval personnel in its use.

The Sixteenth Annual Exhibition of Cardio-Pulmonary Apparatus organized by the Society of Cardiological Technicians of Great Britain is to be held at the Londoner Hotel, Welbeck Street, London, W.1. on Friday, 30 October (5.30 p.m. to 9.0 p.m.) and on Saturday 31 October, 1964 (9.30 a.m. to 1.0 p.m.).

Complimentary tickets are obtainable from Miss Margaret Hale, Exhibition Secretary, c/o Cardiac Research Dept., Guy's Hospital, London, S.E.1.

The British Standards Institution has issued an amendment to B.S. 3494 'Memorandum on Light Current Semiconductor Devices,' Part 1 'Essential Ratings and Characteristics.' The amendmeht gives details of the essential ratings and characteristics of tunnel diodes. This information is regarded as the minimum data which should be provided by a manufacturer when describing his product for general sale. Copies (to be ordered as PD5239) may be obtained from the BSI Sales Branch, 2 Park Street, London, W.1. Price 2s. each. (Postage will be charged extra to non-subscribers.)

Ebauches S.A. of Switzerland has produced a new electronic chronometer of very small dimensions and high precision.

This model is insensitive to vibrations and is specially designed as a time unit for military and civil use as well as air and sea navigation. By virtue of its built-in power source it can be used for both indoor and field operation and its internal capacity is sufficient for 5 days operation.

This timepiece is only 3in × 3in × 6in and the dial diameter 2¼in.

The chronometer maintains an accuracy to within ±0·01 seconds per day.

The Corporation of Trinity House, London, which is the principal pilotage authority in Great Britain, has recently improved the working of the cruising Pilot Cutters at Dungeness, the Sunk and the Nab by the co-ordinated use of radar and v.h.f. radio. Until the adoption of this new technique, the Pilot Cutters have been able to close a ship in fog by using radar, but in very thick weather they have been unable to transfer the pilot because of the likelihood of the motor boarding boat failing to locate the vessel. To overcome this difficulty, each cruising Pilot Cutter now carries a Redifon GR.336 hand-portable v.h.f., f.m. radiotelephone, which can be used by the coxswain of the boarding boat to talk back to the parent Cutter and receive directions from the Officer of the Watch, who can track the boarding boat on the radar.

The boarding boats are equipped with a radar reflector to enable them to be picked up by the Pilot Cutters' radar at very close or long range; the reflector is found to double the range at which the motor boat can be tracked. Experience at Dungeness and Sunk Pilot stations has shown that communication can be maintained over a far greater distance than the boat normally has to travel from the Pilot Cutter when transferring a pilot.

By 'talking' the boat to and from the Cutter it is now possible to ship and land pilots in conditions of considerably reduced visibility that would previously have made it impossible. This application of v.h.f., in conjunction with radar, assists in maintaining the flow of shipping to London and Southampton during fog, and increases the safety of Pilots and boats' crews.

The Fourth International Television Symposium is to be held at Montreux, Switzerland on 24 to 28 May, 1965.

The main topics for which papers are invited will be—(a) The role of television for education purposes, particularly in developing countries, (b) Industrial and other applications, (c) Colour television, (d) Television by satellite.

Further details are obtainable from the Montreux Television Symposium, Postfach 97, Switzerland.

Aer Lingus, Irish International Airlines, has placed an order with Plessey-UK Limited for automatic flight-information recorders to be installed in its B.A.C. One-Eleven jet airliners.

The Irish airline is the second major international operator to order Plessey flight recorders. Last March, BEA placed a £500 000 contract for similar equipment.

Primary use of the flight-information recorder—designed by Plessey-UK in collaboration with S. Davall & Sons Ltd —is to monitor continuously information such as height, air speed, pitch attitude, directional heading and acceleration of an aircraft during flight, so that causes and circumstances of faults can quickly be traced.

In addition to these functions, the system will monitor any nine other instruments. The design also allows the basic parameters to be extended to seventy if required by the addition of sub-multiplexers. Information fed into the instrument is recorded on a stainless steel magnetic wire finer than a human hair. Ministry of Aviation regulations which come into force on 1 July, 1965, require all new passenger and transport aircraft above a certain size to be fitted with a data recording system. The same rule will be extended to cover other aircraft a year later.

The Electronic Engineering Association has issued 'A Guide to a Method of Measuring the Solderability of Round Wires and Component Termination Wires' dealing with the precise and objective methods of assessing the solderability of round wires and component terminations, in order to meet the more exacting requirements of printed wiring as well as conventional circuit wiring. Copies of the Guide are available, free of charge, from The Electronic Engineering Association, 61 Green Street, Mayfair, London, W.1.

Rank Cintel, a division of The Rank Organization, has received an order from the Dunlop Rubber Co Ltd for the development and production of a Resilience Measuring Equipment.

This new equipment will be based on the principles employed on the Projectile Velocity Measuring Equipment and Ballistics Timing Equipment which Rank Cintel has manufactured for a number of years.

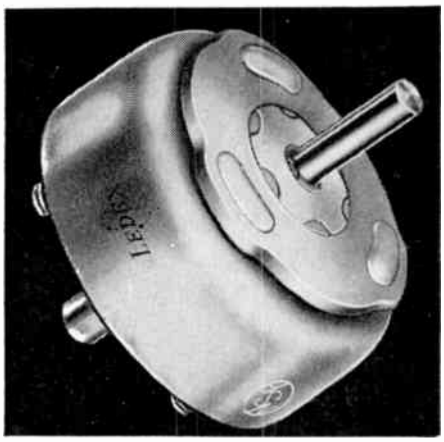
In the first instance the new equipment will be used for the measurement of the resilience of golf balls. In this case a pneumatically propelled projectile is made to impinge upon a stationary golf ball in such a way that both the ball and the projectile pass down a range on a common axis at different speeds, but in reasonably close proximity. Situated across this range are two light screens and their associated light sources. The light screens are fitted with photo-voltaic cells having an extremely fast response.

Interruption of the light beam which is focused on to the voltaic cell causes electrical output pulses to be generated. The pulses pass into a trigger separator which is arranged so that the first interruption of the light beam causes a pulse to arrive at the gate circuit of a microsecond chronometer.

The first light screen generates 'start' pulses and the second screen generates 'stop' pulses, thus the microsecond chronometer times the passage of the golf ball while the digital delay and microsecond chronometer times the passage of the projectile.

EE 73 156 *for further details*

NSF



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A solution to many remote control problems

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CARBON TRACK POTENTIOMETERS
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- Frictionless, snap action
- Up to 98 lb/in. torque
- Rotary strokes of 25° to 95°, clockwise or anti-clockwise
- Substantially linear and level work-to-rotation curve

Enquiries invited for Solenoids with special torque-to-stroke characteristics.

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—combinations of Solenoid and Rotary Wafer Switch sections, permitting a versatility of complicated circuit switching with innumerable applications.

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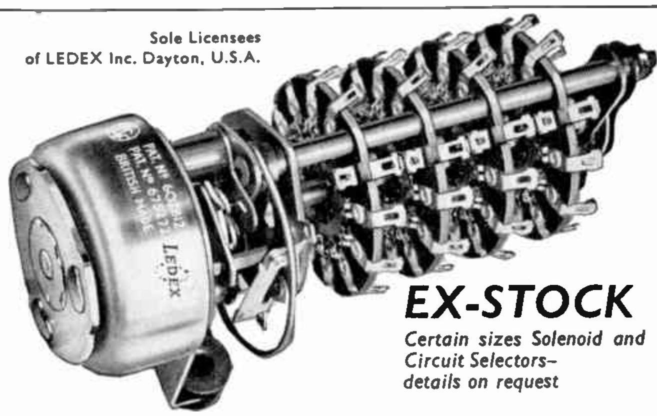

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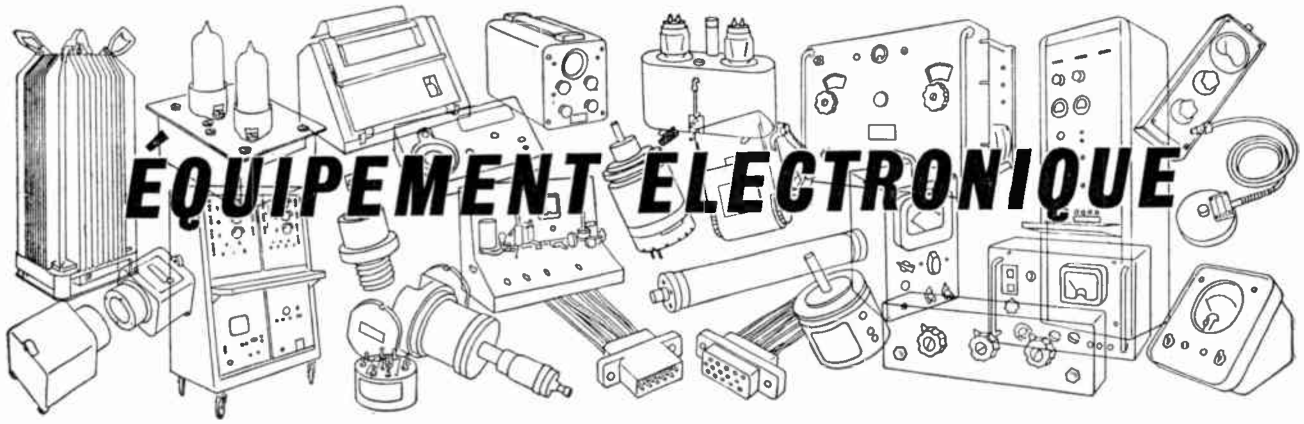


Une description basée sur des renseignements fournis par les fabricants de nouveaux organes. accessoires et instruments d'essai *Traduction des pages 636 à 641*

THERMOMÈTRE DE POINT DE ROSÉE

Shaw Moisture Meters, Rawson Road, Westgate, Bradford 1, Yorkshire

(Illustration à la page 636)

Ce nouvel appareil pour la mesure du point de rosée donne des lectures absolues des températures de l'air ambiantes ou au point de rosée. Il a été spécialement conçu pour indiquer, sur un cadran, les températures de l'air et de points de rosée à des fins météorologiques.

L'indicateur a été prévu pour être placé à l'intérieur et l'appareil de mesure peut être situé à une distance de 12 m de l'indicateur installé dans un écran de Stevenson (la distance requise doit être indiquée à la commande).

La précision de la lecture de la température du point de rosée est inhérente à l'appareil. La lumière de la lampe, actionnée à partir d'une source de tension stabilisée est réfléchie sur une cellule photoélectrique lorsque la rosée se forme sur une surface de miroir. Le courant du thermo-module utilisé pour refroidir le miroir cesse à ce moment-là, de sorte que la surface du miroir, dont la température est mesurée, effectue un cycle continu entre les conditions de rosée et de non rosée.

Ce processus empêche la dérive dans l'un quelconque des circuits de provoquer le refroidissement du miroir au-dessous de la température du point de rosée. Un voyant lumineaux sur le panneau frontal indique le maintien de l'état ou d'absence de rosée. Cette méthode de fonctionnement est considérée comme un progrès technique important par rapport aux systèmes d'asservissement utilisés auparavant pour la mesure automatique du point de rosée. L'appareil est entièrement muni de composants constitués de corps solides.

EE 73 751 *pour plus amples renseignements*

TERMINAISONS MICROONDES

Morganite Resistors Ltd, Bede Trading Estate, Jarrow, Co. Durham

La société Morganite Resistors fournit maintenant, dans une nouvelle gamme de terminaisons microondes de grande puissance d'un encombrement réduit, des terminaisons à ailettes d'aluminium pouvant être soumises à une pression de régime de 2,812 kg/cm². Ces terminaisons type 'A4 sont prévues pour des fils de 1,63 mm de diamètre d'une puissance nominale de 500 W à une température de surface maxima de 300°C. Le taux d'ondes stationnaires de 8.4 à 100 Hz est de 0,95 ou davantage, et les dimensions hors tout ne sont que de 16,51 cm × 10,16 cm. Des terminaisons analogues ont été mises au point pour la gamme de guide d'ondes d'un diamètre de 3,25 mm à 1,22 mm.

Pour les applications à puissance réduite, la série du type 'M' peut être fournie. Ces charges sont à revêtement métallique et dans le modèle à 1,63 mm une puissance de 350 W peut être dissipée avec une température de surface approximative de 600°C. La gamme type 'W' de terminaisons à refroidissement par l'eau a été conçue pour les applications à très haute puissance. Dans le modèle à 1.63 mm une puissance moyenne de 1500 W peut être dissipée.

EE 73 752 *pour plus amples renseignements*

BLOC D'ALIMENTATION DE 400 HZ

Ashgrove Instruments Ltd, 96 Amyand Park Road, Twickenham, Middlesex

(Illustration à la page 636)

Ce bloc d'alimentation a été conçu pour l'alimentation en puissance de synchros, de servomoteurs, de tachygénérateurs, de capteurs et, en général, de tout matériel nécessitant une alimentation de 400 Hz.

On peut régler la commande de resistance de sortie de manière à ce que la tension de sortie monte, demeure constante, ou tombe lorsque la charge est appliquée.

La chute de tension dans les lignes d'alimentataion, fusibles, instruments de mesure, etc. peut être compensée de cette façon le cas échéant.

Il est nécessaire de corriger de manière approximative le facteur de puissance, mais les charges qui peuvent fluctuer rapidement suivant la consommation ou le facteur dc puissance ne dérangent pas l'appareil.

Le modèle standard est fourni sous forme de châssis mais il peut également être fourni, sur demande, pour montage sur bâti ou dans un coffret avec voltmètre, ampèremètre, etc.

La tension de sortie est de 115 V ou 57 V réglables ± 10 V, la puissance de sortie étant de 75 VA.

EE 73 753 *pour plus amples renseignements*

CALIBREUR D'OSCILLOSCOPE

Telequipment Ltd, Chase Road, Southgate, London, N.14

(Illustration à la page 636)

La société Telequipment Ltd. a réalisé un calibreur d'oscilloscope portatif, type C1, qui fournit en un seul appareil toutes les formes d'ondes et toutes les fréquences nécessaires à l'étalonnage complet des oscilloscopes Telequipment et ceux d'autres marques. Il fonctionne sur alimentations entièrement stabilisées.

Ne pesant que 11 kg et mesurant 33 cm × 16,5 cm × 33 cm, le type C1 comporte quatre sorties indépendantes pouvant être utilisées séparément ou conjointement, sans interaction. Le calibreur peut également contrôler, à l'aide des méthodes de comparaison, d'autres sources de signaux.

Les signaux prévus comprennent les ondes carrées commutées à une vitesse de répétition de 100 kHz ou de 1 MHz, les ondes carrées commutées à une vitesse de répétition de 10 kHz ou de 1 kHz, ainsi que les impulsions de marquage du temps commutées à des vitesses de répétition de 1MHz, 100 kHz, 10 kHz, 1 kHz, 50 Hz, et un dispositif de minutage négatif par rapport à la masse.

Le calibreur comprend en outre une source piézoélectrique d'une précision de 0,2% Enfin, il fournit une forme d'onde de télévision non entrelacée, commutée positivement avec environ 200 lignes. L'amplitude est de IV (syne + vidéo).

EE 73 754 *pour plus amples renseignements*

COMMUTATEUR À LAME VIBRANTE SUBMINIATURE

Flight Refuelling Ltd, Wimborne, Dorset

(Illustration à la page 637)

Le Hamlin MSRS-2 est un commutateur à lame vibrante magnétique subminiature, spécifiquement conçu pour la commutation logique à niveau réduit. Il est fourni par la société Flight Refuelling Ltd.

Un nouveau matériau de contact en alliage d'argent assure une résistance de contact uniforme pendant des millions d'opérations. De plus, le rapport élevé entre l'intensité de désexcitation et la tension maximum en charge simplifie la réalisation des circuits logiques et de base lorsque l'omission d'une des conditions contributives peut causer une tension de désexcitation ou empêcher la tension maximum en charge, selon l'état du circuit.

Les essais indiquent un rapport de tension maximum en charge/intensité de désexcitation de 66% à 75% et une résistance de contact maxima de 150 mégohms après un million d'opérations. La tension nominale du commutateur est de 0,5 W/c.c. avec un courant maximum de 10 mA. Il peut être prévu avec une sensibilité de 20 à 100 At suivant les besoins.

La longueur hors-tout du commutateur (conducteurs y compris) est de 5,71 cm. La longueur du corps du commutateur est de 20,3 mm et le diamètre de l'enveloppe de verre est de 2,3 mm.

EE 73 755 *pour plus amples renseignements*

GÉNÉRATEUR DE BALAYAGE

KLB Electric Ltd, 335 Whitehorse Road, Croydon, Surrey

(Illustration à la page 637)

Le générateur de balayage et intégrateur de marqueur PACO G.32 est un générateur de signaux modulé en fréquence ayant une gamme de fréquence centrale de 3 à 213 MHz dans cinq bandes se chevauchant. La largeur de balayage est réglable de 0 à 30 MHz dans la gamme supérieure.

L'appareil est prévu pour l'alignement des tous types d'amplificateurs à large bande. Il se caractérise en particulier par le fait que les marqueurs sont additionnés après le passage du signal à travers le composant soumis à l'essai. Grâce à cette particularité, les signaux de marqueurs ne donnent pas lieu à des résultats pouvant induire en erreur, comme c'est souvent le cas pour ce type de générateur.

EE 73 756 *pour plus amples renseignements*

MICROPHONES AU CHARBON

Airmed Ltd, Edinburgh Way, Harlow, Essex

(Illustration à la page 637)

La société Airmed Ltd. vient de mettre au point deux nouveaux microphones au charbon spécialement prévus pour répondre aux besoins des systèmes de communication modernes qui exigent des accessoires sûrs et de dimensions réduites.

Le microphone type A.301 est un élément miniature à pression d'une sensibilité de—50 dB/V/dyne/cm² à 1 kHz. Il pèse moins de 10 g et il peut être monté sur une antenne de casque nécessitant un accessoire de poids léger et de format miniature. La réponse de fréquence a été spécialement prévue pour assurer une excellente intelligibilité de parole et elle s'étend de 200 à 5000 Hz. Elle est linéaire de 200 à 1000 Hz, puis monte de 8 dB à une pointe de 2000 à 4000 Hz, tombant rapidement au dessus de 5000 Hz.

Le type A.303, qu'on voit dans notre gravure, est un microphone au charbon à pression neutralisant les bruits différentiels et à débit élevé. Il est conçu pour parler de près dans des conditions de bruit ambiant élevé. En tant que microphone différentiel à pression il répond à des sons provenant d'une source faible et proche, c'est à dire les lèvres, tout en discriminant contre des sons provenant d'une certaine distance, à savoir des bruits de l'extérieur. Sa sensibilité est de —50 dB/V/dyne/cm² à 1000 Hz et sa réponse est uniforme de 200 à 4000 Hz avec une large montée autour de 2000 à 3000 Hz et une chute rapide au dessus de 4000 Hz. Le boîtier moulé par injection et à impact élevé comporte un contact à pince spécial pour permettre le montage sur antenne de casque.

EE 73 757 *pour plus amples renseignements*

RELAIS SENSIBLE

Leland Instruments Ltd, 145 Grosvenor Road, Westminster, London, S.W.1

(Illustration à la page 637)

Le nouveau relais de mesure sans contact "Sensicon", type RS1, de la société Leland Leroux est un relais de contrôle sensible et précis. Il comprend un instrument de mesure à cadre mobile, analogue au relais "Sensitact", qui actionne un système de contrôle électronique transistorisé et autonome permettant la mise en oeuvre d'éléments de commande connexes au moment où l'aiguille atteint une position prédéterminée sur l'échelle.

Le relais est livrable en deux modèles. Le type N existe en six versions différentes dont les sensibilités s'étendent, de 100 μA, à 6000 ohms, à 10 mA à 1,5 ohm.

Le type S existe en cinq versions différentes dont les sensibilités vont de 25 μA—à 6000 ohms—à 2,5 mA à 1,5 ohm.

Le "Sensicon" mesure 71 mm × 38 mm de diamètre et se présente sous la forme d'un cylindre monté sur une base à fiches octale. Un capuchon en matière plastique transparente est vissé au sommet du cylindre en matière isolante moulée. Le capuchon laisse voir l'échelle, l'aiguille et l'index de contrôle. Le réglage de l'index se fait en dévissant le capuchon plastique et en tournant l'ensemble du circuit électronique par rapport à la base. On peut ainsi déplacer l'index d'une extrémité à l'autre de l'échelle. Ces éléments peuvent être fournis sur commande avec l'index réglé et fixé à la position voulue au départ de l'usine.

EE 73 758 *pour plus amples renseignements*

COMPTEURS À SCINTILLATIONS

Panax Equipment Ltd, Holmethorpe Industrial Estate, Redhill, Surrey

(Illustration à la page 637)

Deux instruments de comptage nucléoniques, comportant des échelles transistorisées à six décades à action rapide, viennent d'être ajoutées à la gamme Panax. Ils ont été prévus comme instruments universels de laboratoire pour le comptage Geiger et à scintillations et leur taux de comptage maximum est de 50 000 coups/minute. Un sélecteur rotatif permet de prérégler le comptage à n'importe laquelle des neufs valeurs couvrant la gamme de 100 à 1 million. Les signaux de sortie des décades sont fournis aux douilles coaxiales à l'arrière de l'instrument et ils sont injectés à un imprimeur approprié pour l'impression automatique au terme d'une période de comptage. Des commutateurs à bouton-poussoir commandent la mise en marche et le réenclenchement.

L'alimentation très haute tension pour les compteurs Geiger et à scintillations est produite à l'intérieur de l'instrument et elle est pleinement stabilisée jusqu'a ±10% à l'égard des fluctuations de tension secteur. Elle peut être réglée de 0 à 2kV au moyen d'un potentiomètre à 10 tours de précision et à bouton de commande gradué. Les tensions nécessaires de polarisation de discrimination sont également produites à l'intérieur de l'appareil et elles sont également réglables. Un commutateur à quatre directions permet de choisir des temps morts de 4, 20, 200 ou 40 μsec.

Cette description s'applique de manière égale à ces deux nouveaux instruments, qui porteront les références de type P.7702. La principale différence réside dans le type P.7702 (voir notre gravure) qui comporte une minuterie à six décades ainsi que l'échelle à action rapide, plus connue sous le nom précis d'échelle automatique, vu qu'elle comprend des dispositifs automatiques de comptage et de temps préréglés.

EE 73 759 *pour plus amples renseignements*

THERMOMÈTRE ÉLECTRONIQUE

Sidien Products Ltd, 11 Birchwood Court, Edgware, Middlesex

(Illustration à la page 638)

Ce thermomètre électronique comporte deux gammes, choisies par sélecteur, de −10° C à 90° C et de −10° C à 150° C.

Chaque thermomètre comprend deux palpeurs, reliés à la boîte de mesure par un conducteur miniature de 121 cm de long, le choix du palpeur voulu s'effectuant au moyen d'un commutateur à deux directions.

L'appareil fonctionne sur pile Mallory MN1300T2, dont la durée de vie dépasse deux ans.

La constante de temps thermique est de 20 sec et le degré de précision est de l'ordre de ±2° C.

EE 73 760 *pour plus amples renseignements*

TRANSFORMATEURS VARIABLES

The Cressall Manufacturing Co. Ltd, Cheston Road, Birmingham 7

(Illustration à la page 638)

Un transformateur variable encapsulé à bobinage automatique vient d'être ajouté à la gamme Torovolt d'éléments à rapport variable réalisé par la Cressall Manufacturing Co. Ltd.

Le nouveau Torovolt, modèle 33Y, constitue un instrument sûr et économique pour le contrôle d'appareils à courant alternatif, ainsi que pour le contrôle de charges nominales continues allant jusqu'a 0,6 A. Ses dimensions sont les plus réduites possibles par rapport à la longue durée de vie qu'on peut en exiger.

Le nouveau transformateur a été conçu pour être relié directement à une alimentation en alternatif de 240 V et il permet d'obtenir des variations régulières de sortie du zéro à la tension de ligne, à un courant nominal maximum de 0,8 A. Le noyau bobiné est enrobé dans un moulage isolé et mesure 8 cm de diamètre sur 5,5 cm de long. Les connexions sont effectuées au moyen de bornes montées dans des blocs moulés faisant corps avec le coffret.

Des modèles peuvent être fournis pour montage soit à l'avant soit à l'arrière de la plaquette.

EE 73 761 *pour plus amples renseignements*

CAMÉRA DE TÉLÉVISION TRANSISTORISÉE

Automatic Information and Data Services Ltd, 26 Sheen Road, Richmond, Surrey

(Illustration à la page 638)

Une caméra entièrement transistorisée vient d'être réalisée par A.I.D.S. Ltd. Elle est destinée à tirer une performance maxima du Vidicon de 2,54 cm, utilisé à une tension d'anode murale inférieure à 250 V, ainsi que pour le fonctionnement continu.

L'expérience a montré qu'après une période initiale de "rodage" à des températures ambiamtes variant entre −50° et +45°C, l'entretien se limite au nettoyage des lentilles et au remplacement du tube Vidicon après 4000 heures. La caméra peut être réglée pour le fonctionnement continu à des températures ambiantes plus élevées, le facteur limitatif étant, dans ce cas, la performance du Vidicon à des niveaux de chaleur élevés. La caméra comprend 21 transistors et 13 diodes, des dispositifs au silicium étant utilisés lorsqu'un meilleur rendement ou une plus grande résistance à la chaleur sont nécessaires.

Bien que nominalement prévue pour le fonctionnement libre à 625 lignes, d'autres étalons de ligne peuvent être fournis. On peut, en outre, fournir, au moyen d'un générateur de formes d'onde A.I.D.S. un entrelacement de 2:1 pour étalons C.C.I.R. et formes d'ondes A.I.D.S.

La caméra est normalement fournie avec un bloc d'alimentation dans un coffret coulé en matrice et elle est entièrement automatique. Un bloc d'alimentation de type modulaire peut être fourni avec un commutateur de niveau de lumière automatique/manuel pour l'emploi avec d'autres éléments modulaires A.I.D.S.

La caméra peut résoudre 650 lignes de télévision au centre de l'image et 400 lignes de télévision dans les coins lorsqu'une modulation de 50% (blanc/noir) se produit à 330 lignes de télévision au centre et à 250 lignes de télévision dans les coins. (Un correcteur d'ouverture de forme modulaire peut être fourni pour porter la modulation centrale à 90%).

EE 73 762 *pour plus amples renseignements*

INSTRUMENTS DE MESURE POUR PANNEAUX

Smiths Industrial Division, Kelvin House, Wembley Park Drive, Wembley, Middlesex

(Illustration à la page 638)

Ces petits instruments de mesure pour montage sur panneaux sont d'une sensibilité extrême et d'une très grande précision et constituent la plus récente réalisation dans la gamme d'instruments d'enregistrement et de laboratoire de la Smiths Industrial Division.

Ils comportent des mouvements à suspension par ligaments tendus, avec aiguilles de verre donnant des indications sur une échelle très claire. La gamme minima de ces instruments est de 0-6 μA ou de 3-0-3 μA avec zéro central.

La précision des instruments à montage horizontal dont l'échelle totale est de 25 μA ou davantage peut atteindre 0,5% et une précision analogue peut être obtenue avec des instruments à montage vertical d'une échelle totale de 100 μA ou davantage.

Les dimensions courantes de l'instrument de mesure sont: 72 × 84 mm, la longueur de l'échelle étant de 63 mm.

EE 73 763 *pour plus amples renseignements*

ALIMENTATION DE CELLULES ÉMISSION-RÉCEPTION

Ferranti Ltd, Ferry Road, Edinburgh 5

(Illustration à la page 638)

La société Ferranti Ltd vient de mettre au point un nouveau bloc d'alimentation à haute tension, à la fois compact et d'un poids réduit. Ce bloc fournit une alimentation d'entretien à la cellule TR Ferranti type WF4(CV2311). Il a été conçu pour résoudre le problème que pose l'alimentation en tension continue élevée des avions.

Deux versions sont prévues, l'une donnant une tension d'entrée de 115 V, 400 Hz et l'autre une tension de 200 V, 400 Hz. Les deux versions ont une tension d'amorçage de 1050 V c.c. ±10%, qui se réduit à une tension d'entretien normale après l'amorçage.

L'ensemble mesure 3,47 cm × 3,47 cm × 3,78 cm et pèse 85 grammes. Il est enfermé dans une capsule en résine d'époxyde. Il répond à la classification d'humidité H6 des services interarmes, ainsi qu'à la spécification DEF.5214 pour les transformateurs à moulage de résine. La gamme de températures ambiantes prévues s'étend de −40°C à 100°C et l'appareil a passé avec succès le contrôle de panne de tension à une altitude de 21336 m.

Le bloc peut résister à une tension de pointe inversée de 4 kV pendant 10 μsec et ne risque pas d'être endommagé par un court-circuit accidentel de la puissance de sortie.

Des variantes de mêmes dimensions peuvent être fournies pour l'emploi avec d'autres spécifications de cellules TR Ferranti.

EE 73 764 *pour plus amples renseignements*

CONTRÔLEURS DE COUPLE

Mining & Scientific Equipment Ltd, 317 Kennington Road, London, S.E.11

(Illustration à la page 639)

La société M.S.E. Ltd vient d'annoncer la création d'un nouvel instrument de mesure du moment de torsion. Conçu en particulier pour le contrôle par lots et le contrôle de qualité de moteurs à fraction de cheval, le nouvel instrument utilise un type inédit de dynamomètre à absorption à commande électronique qui permet de mesurer les indications de couple avec précision et indépendamment de la vitesse rotationnelle.

Le modèle qu'on voit dans notre illustration a une gamme de mesure du couple de 155,5 grammes à 1555 grammes, la précision étant de 1,5 de l'échelle totale de la gamme de couple. Il existe aussi d'autres gammes. La vitesse de rotation du moteur peut aussi être mesurée simultanément et indiquée jusqu'a 5000 tours/minute.

La lecture peut être obtenue sur un instrument de mesure à échelle ouverte. En variante, elle peut être affichée sur un oscilloscope à longue persistance de 42,5 cm, donnant une trace immédiate des caractéristiques du couple de vitesse.

Un micromètre peut être disposé sur la face de l'écran cathodique afin de pouvoir déterminer facilement la qualité du moteur. L'emploi d'un oscilloscope à longue persistance permet également de prendre des photographies de la trace.

EE 73 765 *pour plus amples renseignements*

FER À SOUDER

Distributeurs: Oliver Dow Ltd, 877a High Road, Finchley, London, N.12

(*Illustration à la page 639*)

Le fer à souder BLIXT, construit en Suède, contient dans sa poignée un rouleau de soudure à noyau de flux qui alimente l'étampe de la barre à souder par un déclencheur à "pistolet". L'opérateur n'a donc besoin que d'une seule main pour utiliser le fer, l'autre main demeurant libre pour tenir une pièce ou des pinces.

Le fer est fourni avec une barre de réserve, d'autres pouvant être obtenues sur commande. La soudure est enroulée sur des bobines en matière plastique pouvant recevoir jusqu'à 487 cm de soudure.

EE 73 766 *pour plus amples renseignements*

MESURE DE NIVEAU

Vacuum Reflex Ltd, 6 Soho Street, London, W.1

(*Illustration à la page 639*)

La société Vacuum Reflex Ltd a mis au point une nouvelle gamme d'instruments pour l'indication et le contrôle continu du niveau. Le système utilise des ondes à ultrasons réfléchies par l'interface à liquide libre ou à matières solides lâches. Il permet la mesure du niveau de tous les types de liquides, de matières granulaires, de solides et de poudres.

L'appareil comprend deux transducteurs à ultrasons, dont un émetteur et un récepteur, montés au dessus du niveau maximum du contenu du réservoir ou de la soute. Pour déterminer le niveau, on mesure le temps mis par l'onde réfléchie pour retourner de l'interface au récepteur. Ce temps est indiqué sur un enregistreur étalonné en niveau ou en contenu de réservoir. Tous les éléments extérieurs sont encapsulés et l'ensemble de l'appareil est transistorisé, ne comprenant que des transistors au silicium ou planaires.

Lorsque les transducteurs sont montés au dessus du contenu du réservoir, la mesure du niveau ne dépend plus que de la réflexion des ondes à ultrasons par la surface du matériau. Le système est, par conséquent, indépendant de la nature et de l'état du contenu du réservoir ou de la soute. Les matériaux pouvant être mesurés comprennent l'eau, le lait, le charbon, la cendre, la pierre, le roc, le sel, etc. Le système se prête également à la mesure des interfaces liquide/liquide. Il a été utilisé pour mesurer les interfaces de liquide/écume et il peut être employé à l'intérieur des réservoirs à haute pression.

Les appareils livrables actuellement sont prévus pour des gammes allant jusqu'à 12.192 m (30,48 m dans les liquides) mais des travaux de mise au point sont en cours pour étendre ces gammes.

EE 73 767 *pour plus amples renseignements*

TRANSDUCTEUR À PRESSION

Consolidated Electrodynamics Division, Bell & Howell Ltd, 14 Commercial Road, Woking, Surrey

(*Illustration à la page 639*)

Un transducteur à pression à gamme réduite, conçu pour les applications exigeant des composants légers et d'un faible encombrement, a été créé par la Consolidated Electrodynamics Division de la société Bell & Howell Ltd.

Le transducteur à pression type 4-353, serait, en effet, le plus petit composant du genre existant sur le marché et sa gamme de pression absolue va de 0 à 0.070 kg/cm². Il ne pèse que 198,45 grammes et mesure 6,35 cm de long sur 3,81 cm de diamètre.

Le nouveau transducteur a été conçu pour la mesure des pressions à haute altitude de bord, les pressions de chambres d'altitude, les pressions de souffleries et les pressions d'échappement de turbines à vapeur. Sa première application consistait en une étude des pressions produites par la mise à feu d'une fusée dans une chambre d'altitude.

Il utilise un nouveau modèle de diaphragme qui assure une grande efficacité de fontionnement dans un minimum d'espace et de dimensions. Toutes les compensations s'effectuent à l'extérieur de l'élément actif scellé. Une butée mécanique de suppression accorde à l'instrument une tolérance de 1,4 kg/cm² sans dommage.

La gamme de température réalisable s'étend de $-100°F$ à $+275°F$, la sensibilité est de 20 mV, le décalage de sensibilité thermique est de $\pm 0.005\%$ par degré Farenheit dans la gamme de température compensée. La linéarité et l'hystérésis combinés sont de $\pm 0,5\%$ ou supérieurs à la sortie totale des gammes.

EE 73 768 *pour plus amples renseignements*

BLOC D'ALIMENTATION À KLYSTRON

Distributeurs: Miles Hivolt Ltd, Old Shoreham Road, Shoreham-by-Sea, Sussex

(*Illustration à la page 640*)

Ce bloc type LS52R, construit par Oltronic de Suède, est un bloc d'alimentation universel pouvant être utilisé avec la plupart des klystrons actuellement sur le marché.

Il assure les alimentations suivantes:
Tension de faisceau: -200 V à $+3.6$ kV
Tension de réflexion: 0 à 1 kV, 50 μA
Tension de grille: 300 V à $+150$ V, 5 mA
Tension de chauffage: 6,3 V 3 A.

Toutes les sorties ont une excellente stabilité et une faible ondulation, la tension de réflexion ayant une ondulation inférieure à 200 μV.

Le bloc comprend quatre générateurs de modulation incorporés et fournit des ondes carrées, des impulsions, des ondes sinusoïdales et des dents de scie.

EE 73 769 *pour plus amples renseignements*

COMPARATEUR D'EXTENSOMÈTRES ACOUSTIQUES

Westland Aircraft Ltd, East Cowes, Isle of Wight

(*Illustration à la page 640*)

La société Westland Aircraft Ltd a conçu et produit un comparateur portatif pouvant être utilisé avec des extensomètres acoustiques à fil vibrant. Il a été réalisé à la suite du contrat obtenu par cette société pour l'installation d'extensomètres acoustiques et à feuille de métal à la centrale nucléaire d'Oldbury-on-Severn.

Les réalisateurs ont tenu à simplifier autant que possible le fonctionnement et l'indication. Les commandes se limitent à un commutateur arrêt-marche, ainsi qu'à des bornes de branchement pour l'extensomètre soumis à l'essai. L'instrument lit la fréquence des vibrations acoustiques en les comparant directement avec un cristal de quartz intérieur, la fréquence étant indiquée sur un dispositif de projection à quatre chiffres. L'appareil comprend une source d'énergie pour le fil extensométrique, ainsi que des circuits d'entraînement pour rendre le processus entièrement automatique. Lorsqu'on relie le comparateur à l'extensomètre, ce dernier est instantanément contrôlé et son indication est affichée. Ce processus est répété à intervalles de quatre secondes.

L'instrument peut être utilisé à l'extérieur, en particulier sur des chantiers, car il est logé dans un coffret portatif fort robuste protégeant les commandes contre tout risque d'endommagement en transit. Il est entièrement autonome et cela s'applique également à la batterie rechargeable poids léger. On peut le recharger à partir d'un élément de charge extérieur. Un stabilisateur interne permet de maintenir le fonctionnement dans une série étendue de conditions de batterie.

EE 73 770 *pour plus amples renseignements*

SYSTÈME DE CONSTRUCTION

A.P.T. Electronic Industries Ltd, Chertsey Road, Byfleet, Surrey

(*Illustration à la page 640*)

La société A.P.T. Electronic Industries Ltd. a réalisé un nouveau système de construction à châssis modulaire, prévu pour le matériel électronique et vendu sous le nom de MINAR. Ce système a été conçu par la BBC qui l'a utilisé expérimentalement pendant quelques temps et il est basé sur un panneau standard de 48,26 cm × 13,33 cm ayant la forme d'un bâti miniature dans lequel peuvent s'insérer des éléments de diverses grandeurs. Le système convient particulièrement pour les assemblages à circuit imprimé ou à transistors.

L'assemblage principal de montage du système est le cadre, prévu pour s'insérer directement dans un bâti de 48,26 cm.

Le cadre est percé et taraudé pour pouvoir recevoir des modules de différentes largeurs, dont toutes sont des multiples de la largeur de l'élément de base, soit 2,22 cm. Des modules ayant vingt fois cette largeur de base peuvent être montés dans n'importe quel ordre sur un cadre et on peut fixer les modules de manière permanente dans le cadre ou les rendre retractables, selon les besoins. Les dimensions sont prévues de manière à ce qu'aussitôt fixé le premier module, les vis de fixation des autres modules s'accordent automatiquement avec les trous correspondants dans le cadre.

Le système comprend un coffret et un couvercle, dans lequel peut être inséré un cadre pour le rendre portatif, ainsi que des capots pouvant renfermer des éléments individuels ou des assemblages complets. D'autres accessoires comprennent des plaquettes à châssis et à circuit imprimé de la grandeur voulue pour s'adapter directement aux modules.

La plupart des éléments du système sont fournis complètement assemblés mais, afin de réduire les frais d'expédition, le plus grands modules sont livrés sous forme de trousses. Le seul outil nécessaire à l'assemblage est un tournevis et aucune habileté spéciale n'est requise. Les composants individuels qui servent à la fabrication des cadres, modules, etc. sont également livrables séparément, de manière à permettre aux utilisateurs de réaliser les montages qui répondent à leurs besoins particuliers.

EE 73 771 *pour plus amples renseignements*

MULTIMÈTRE ÉLECTRONIQUE

Comark Electronics Ltd, Gloucester Road, Littlehampton, Sussex

(Illustration à la page 640)

Le multimètre électronique type 130 allie la souplesse d'emploi du multimètre à la sensibilité du voltmètre électronique. Fonctionnant sur batterie et étant entièrement transistorisé, il est totalement exempt de problèmes de bourdonnement et de masse. Il est très compact et n'exige pas de temps de chauffage.

Il comporte 54 gammes de base, s'étendant de 10 mV à 300 V, de 10 μA à 100 mA c.a. et c.c., et de 1 μA à 1 kV c.c. (toutes les gammes étant à déviation totale) avec zéro central sur toutes les gammes de courant continu.

La résistance d'entrée est d'environ 2 MΩ sur toutes les gammes de courant alternatif et de 1 MΩV sur toutes les gammes de courant continu à l'exception de la plus élevée. La chute de tension aux bornes sur toutes les gammes de courant est inférieure à 12 mV. Le bruit est négligeable sur toutes les gammes de tension alternative sauf la plus sensible où il ne dépasse pas 1 mV lorsqu'elle est alimentée par une source d'impédance de 100 kΩ. La réponse de fréquence est linéaire jusqu'à 50 kHz, avec réponse utile jusqu'à 250 kHz.

Les mesures de résistance s'effectuent sur échelles linéaires, s'étendant de 100 Ω à 1 MΩ de la totalité de l'échelle sur cinq gammes. De plus, il y a une gamme à haute résistane de 1 MΩ au centre de l'échelle. Ces gammes donnent une couverture totale de 1 Ω à 100 MΩ. Des mesures de résistance jusqu'à un maximum de 10 000 MΩ et un minimum de 0,01 Ω peuvent être effectuées à l'aide de batteries extérieures.

L'instrument se caractérise par la possibilité qu'il offre d'effectuer des mesures de résistance à trois bornes. La résistance peut être mesurée directement par déviation à une troisième borne (par exemple la masse).

L'appareil peut aussi être utilisé comme amplificateur de tension ou de courant aussi bien sur les gammes de courant alternatif que sur celles de courant continu. La sortie est d'environ 1 V pour l'entrée choisie sur le sélecteur de gamme.

Le circuit se compose d'un amplificateur à gain élevé c.a./c.c. actionnant un instrument de mesure à échelle à miroir de 8 cm de 100 μA. Pour la mesure de courant continu, l'instrument de mesure est précédé d'un préamplificateur à chopper transistorisé, donnant une très faible dérive. Pour la mesure de courant alternatif, un étage d'entrée à impédance élevée est utilisé. Le gain est stabilisé par une puissante contre-réaction totale.

EE 73 772 *pour plus amples renseignements*

ENCODEUR À ARBRE NUMÉRIQUE

Distributeurs: B & K Laboratories Ltd, 4 Tilney Street, Park Lane, London, W.1

(Illustration à la page 641)

L'encodeur à arbre numérique, Peekel type PP3A1, a été étudié pour être fixé à des enregistreurs potentiométriques ou à des appareils similaires dont l'information analogique se rapporte à la position angulaire d'un arbre rotatif. L'élément PP3A1 convertit cette information en une donnée numérique qui est injectée directement à des éléments supplémentaires de lecture ou d'affichage.

L'encodeur PP3A1 est fixé à l'enregistreur ou à tout autre instrument au moyen de pinces et d'une plaque de montage. Il peut être actionné directement ou accouplé à l'arbre au moyen d'un câble et de poulies. Un modèle spécial est prévu pour l'emploi avec l'enregistreur Bruel et Kjaer 2305.

La sortie fournie par l'élément PP3A1 est présentée sous forme d'information décimale à trois décades de 0 à 999 par rapport à un tour complet d'arbre. Cette sortie peut âtre alimentée directement, sans décodage aucun, à un indicateur numérique Peekel type PP5C1. Elle peut être utilisée pour actionner un imprimeur numérique, à l'aide d'un convertisseur d'impression Peekel type PP9BA1, ou un perforateur de bande, à l'aide d'un convertisseur de bande perforée Peekel type PP9BF1. D'autres éléments Peekel, conçus pour l'emploi en liaison avec l'adaptateur PP3A1, comprennent une mémoire numérique, une horloge numérique, un compteur d'impulsions et des commutateurs d'entrée permettant de relier jusqu'à 200 canaux à un seul instrument ou système numérique.

Notre gravure montre des vues intérieures et extérieures de l'encodeur.

EE 73 773 *pour plus amples renseignements*

RELAIS SUBMINIATURE

Plessey-UK Ltd, Abbey Works, Titchfield, Fareham, Hampshire

(Illustration à la page 641)

Le nouveau relais subminiature type CF de la Plessey UK Ltd a été approuvé par la spécification ministérielle DEF 5165. Il s'agit d'un relais à permutation bipolaire à action rotative, entièrement scellé et avec une résistance de contact maxima de 0,030Ω assurée par deux contacts dorés.

Les broches de connexion sont espacées sur le module standard de 2,3 mm pour le montage de circuits imprimés et elles passent à travers des scellements verre-métal. Les connexions de bobines et contacts à ces broches sont disposées de façon symétrique, de manière à ce que le fonctionnement correct puisse être assuré quel que soit le mode dont le relais est enfiché à son support.

La puissance nominale de contact est de 1 A à 3 A, à 28 V c.c. ou 115 V c.c. résistive, suivant le nombre d'opérations.

EE 73 774 *pour plus amples renseignements*

COMMUTATEUR À CIRCUIT IMPRIMÉ

NSF Ltd, 31-32 Alfred Place, London, W.C.1

(Illustration à la page 641)

La société NSF vient d'annoncer la mise au point d'un commutateur rotatif de contrôle unipolaire à circuit imprimé et à dix positions, destiné à l'instrumentation et à l'appareillage de contrôle. Ce commutateur est livrable en éléments individuels ou en éléments groupés à quatre sections, contrôlés par un seul volant à pouce, ou encore en assemblages empilés d'un maximum de douze éléments individuels. Le mécanisme à index est du type à double boule, l'index ne couvrant que 36°, et une butée sera ajoutée pour limiter la rotation à n'importe quelle position entre 2 et 10. Les contacts ne risquent pas de courts-circuits.

La plaquette standard à circuit imprimé assure la commutation unipolaire à 10 directions. Des plaquettes spéciales peuvent être outillées sur commande. Les bornes sont du type à cosse soudée. En variante, on peut employer des connecteurs de circuit imprimé sur module de 6 mm.

Des fentes permettent le montage universel sur panneaux de diverses épaisseurs.

Les volants à pouce sont en phénolique noir avec chiffres blancs et le couvercle est en gris Delrin. D'autres couleurs pourront être ajoutées par la suite sur commande.

EE 73 775 *pour plus amples renseignements*

ELECTROMÈTRE MULTIGAMMES

Thomas Industrial Automation Ltd, Station Building, Altrincham, Cheshire

(*Illustration à la page 641*)

La société Thomas Industrial Automation Ltd vient d'ajouter un nouveau modèle—le type VC.99A—à sa gamme d'électromètres. C'est un instrument complémentaire au VC.99, dont il garde toutes les remarquables caractéristiques. Il offre, cependant, l'avantage supplémentaire d'un commutateur doubleur de gamme. Il remplit une double fonction, à savoir:

(a) celle d'enregistreur de micro-courants, pouvant mesurer des courants atteignant 600 μA en huit gammes commutées. Le sommet de la gamme a été prévu pour chevaucher la gamme la plus sensible des multimètres classiques. L'emploi de la dégénération rapide assure une faible chute de tension sur les mesures de courant.

(b) celle de voltmètre électromètre, pouvant mesurer directement des potentiels des deux polarités de 0 à 2000 mV en deux gammes commutées, avec une résistance d'entrée équivalente, supérieure à 10^{13} Ω.

Vu que l'électromètre peut mesurer des courants extrêmement faibles, il peut être aisément disposé de manière à pouvoir mesurer des résistances très élevées. On peut ainsi mesurer des résistances allant jusqu'à 10^{14} Ω à l'aide d'une source extérieure de potentiel. Lorsqu'une alimentation de contrôle de 500 V est exigée, comme c'est le cas dans certains contrôles de condensateurs et de câbles, un adaptateur spécial à haute stabilité de 500 V peut être fourni. Des mesures atteignant 5×10^{14} Ω peuvent être effectuées à l'aide de cet adaptateur.

L'instrument peut être logé dans un robuste coffret Burma en bois de teck ou dans un coffret en matière plastique laminée noire, tous deux munis d'une poignée de transport et d'un couvercle amovible.

EE 73 776 *pour plus amples renseignements*

Résumés des Principaux Articles

Instruments microélectroniques par S. S. Forte

Résumé de l'article aux pages 586 à 590

La miniaturisation de matériel électronique se développe à une allure sans précédent, résultant de l'évolution des nouvelles méthodes basées sur les composants constitués de corps solides et de la nécessité d'une fiabilité sans cesse accrue.

La première partie de cet article se propose d'examiner les méthodes actuelles qui permettent d'améliorer la sûreté du rendement et de réduire les dimensions. Elle est suivie d'une description d'une calculatrice numérique expérimentale qui emploie des techniques applicables à ce genre d'appareil. La deuxième partie décrira les étapes du développement des circuits linéaires pour une installation expérimentale microminiaturisée de navigation aérienne.

Commutation automatique de thermocouples sensibles par J. L. Goldberg et H. M. King

Résumé de l'article aux pages 591 à 595

Les auteurs décrivent un appareil pour la commutation automatique de thermocouples sensibles à des intervalles à variation continue de quelques secondes à plusieurs dizaines de minutes. La partie mécanique se compose d'un système téléphonique unisélecteur à cliquet et rochet engrené à l'arbre d'un commutateur de thermocouple de haute qualité. Un contrôleur électronique fournit périodiquement un certain nombre d'impulsions à l'unisélecteur conformément au rapport d'engrenage, assurant ainsi le réglage voulu du commutateur de thermocouple.

L'appareil a été mis au point pour l'enregistrement automatique des températures et des gradients de température le long de standards de ligne et de bandes d'étude normales par l'emploi de thermocouples.

Un circuit simple de mesure de capacitance avec présentation numérique des valeurs de capacitance par S. L. Hurst

Résumé de l'article aux pages 596 à 599

Cet article traite d'une méthode simple, très rapide et à l'épreuve des erreurs, pour mesurer les valeurs de capacitance dans la gamme d'environ 0,001μF à environ 10μF. La précision de mesure est de $\pm 2\%$ près. Un degré de précision beaucoup plus élevé peut être atteint dans les gammes de valeurs plus réduites. La méthode peut être étendue également à la mesure de valeurs de résistance, mais avec certaines restrictions quant à la précision. L'équilibrage de condensateurs et de résistances peut être effectué, en outre, avec un degré de précision particulièrement élevé.

Opérateur d'intégration et de disjonction pour corrélateurs en série par R. M. Seeley

Résume de l'article aux pages 601 à 604

Le besoin d'un intégrateur à temps fini pour corrélateurs numériques en série se fait sentir depuis longtemps. Les conditions essentielles exigées sont le pouvoir d'accepter un taux élevé de chiffres binaires avec une intégration stable pour les taux aléatoires, une disjonction rapide et totale, un transfert stable du total à une sortie à faible impédance. Un dispositif répondant à ces critères est *décrit dans cet article, ainsi qu'un nouveau flip-flop rapide. La précision globale est d'environ 2% avec une entrée de données de 2 MHz, en utilisant une intégration de 300μsec et une période de disjonction de 2μsec.*

Un bloc d'alimentation asservie de 100V, 100mA, à transistors par R. E. Aitchison et W. S. Lamond

Résumé de l'article aux pages 604 à 605

Cet article décrit un module d'alimentation commandée de 100V, 100mA. Sa réalisation met en lumière certains principes pouvant être appliqués à l'étude et à la protection des circuits d'alimentation à transistors lorsque les tensions d'alimentation dépassent les tensions nominales de panne des transistors. Un certain nombre de ces modules peut être mis en cascade pour assurer une alimentation en tension plus élevée.

Réservoir de ferrite à accès séquentiel et à dispositif d'adresse par commutateur à plots par M. D. A. B. Rackowe

Résumé de l'article aux pages 612 à 615

Un réservoir de données numériques à bon marché a été réalisé à l'aide d'un commutateur à plots utilisé en liaison avec une matrice à noyau de ferrite et des circuits semi-conducteurs. Le réservoir est du type à accès par mots et celui qui est décrit dans cet article a une capacité de 10 mots par 10 chiffres binaires.

La simplicité résultant de l'emploi d'un commutateur à plots n'a pu être réalisée qu'en sacrifiant la vitesse et l'accès aléatoire serait également difficile à obtenir.

Un phasemètre à impulsions par R. E. King

Résumé de l'article aux pages 615 à 616

Il s'agit ici d'un circuit simple à semi-conducteurs donnant une indication de sortie proportionnelle à l'incidence de différence de temps entre deux trains d'impulsions d'entrée binaires. Le circuit comporte un réseau bistable classique et un circuit de porte différentiel donnant une sortie ternaire. L'adjonction d'un préamplificateur à chaque canal d'entrée permet de convertir l'instrument en un phasemètre à fréquence acoustique.

Quelques transformations de la matrice d'admittance nodale d'un circuit et son application à un amplificateur de différence par K. G. Nichols

Résumé de l'article aux pages 617 à 621

L'auteur présente une technique permettant d'établir la matrice d'admittance d'un circuit à orifice k, les orifices étant entièrement distincts. Le résultat est ensuite analysé afin d'obtenir les paramètres d'admittance d'un réseau quadripolaire. Par une méthode analogue, mais à l'aide d'une transformation différente de la matrice d'admittance, on analyse les réponses de déphasage et en phase d'une paire à longue queue ou amplificateur de différence.

Deux amplificateurs sélecteurs de fréquence à transistors par S. Harkness

Résumé de l'article aux pages 622 à 626

L'auteur traite de deux amplificateurs autonomes à sélection de fréquence entraînés par transistors. Les deux instruments utilisent une variante du pont de Wien décrit par Wigan qui assure le contrôle de fréquence au moyen d'une seule résistance variable, évitant ainsi les difficultés du dépistage précis lorsqu'il faut faire varier simultanément un certain nombre de résistances. Ces amplificateurs se prêtent particulièrement à l'utilisation comme détecteurs dans les ponts à courant alternatif. L'un d'eux est destiné au contrôle de la perte en fer et ne nécessite pas une sensibilité élevée mais exige, cependant, en raison de la présence possible d'un grand nombre d'harmoniques d'ordre inégal, une sélectivité très élevée. La couverture de fréquence s'étend de 20 à 600 Hz en trois gammes. La sortie est affichée sur un instrument de mesure à cadre mobile, sur lequel une déviation totale de 2% correspond à 5μV. L'autre amplificateur est un détecteur universel pour ponts à fréquence acoustique linéaire de la plus haute précision. Il s'agit surtout ici d'un instrument des plus sensibles, la déviation totale de l'enregistreur de sortie correspondant à un signal d'environ ½μV. La couverture de fréquence va de 20 Hz à 20 kHz en six gammes.

Expériences avec une ligne à retard Garent à bande X par J. H. Collins, B. Yazgan et J. Cochrane

Résumé de l'article aux pages 627 à 629

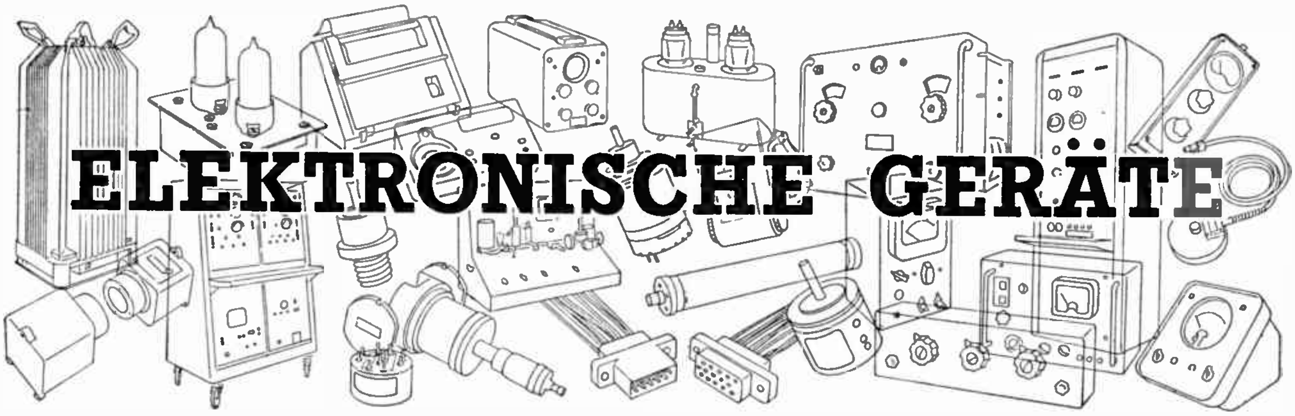
Les auteurs décrivent des expériences effectuées sur l'excitation d'ondes acoustiques rotationelles à bande X dans un disque à un seul cristal de 0,31cm de diamètre et aimanté axialement. Ce disque en yttrium est inséré dans une cavité résonante à une température ambiante de TM_{010}. La perte minima par insertion de ce dispositif est de 45dB sur la gamme de puissance d'entrée de −36 +7dBm.

Un analyseur électronique pour paires de transistors utilisées dans les amplificateurs équilibrés par R. R. Vierhout et A. J. H. Vendrik

Résumé de l'article aux pages 630 à 631

Il est montré dans cet article que le paramètre de transistor qui détermine principalement le rapport de rejet d'amplificateurs de tension équilibrés est de $r_b = h_{11}/(h_{21}+1)$

Une analyseur électronique qui indique cette valeur en fonction du courant émetteur pour deux transistors simultanément a été mis au point. Une possibilité d'étalonnage est prévue.



Beschreibung neuer Bauelemente, Zubehörteile und Prüfgeräte auf Grund der von Herstellern gemachten Angaben.

Übersetzung der Seiten 636 bis 641

Taupunkt-Thermometer

Shaw Moisture Meters, Rawson Road, Westgate, Bradford 1, Yorkshire

(*Abbildung Seite 636*)

Dieses neue Taupunkt-Messgerät gibt Absolutanzeige der Umgebungs- und Taupunkttemperatur der Luft und wurde besonders für die Anzeige der Luft- und Taupunkttemperaturen auf einer Messgerätskala für meteorologische Zwecke entwickelt.

Das Anzeigegerät ist für Innenmontage konstruiert, während das Messgerät bis zu 12 m vom Anzeiger entfernt in einem Stevenson-Schirm installiert werden kann; die Entfernung muss bei Bestellung aufgegeben werden.

Die Genauigkeit der Taupunkttemperaturanzeige beruht auf der Konstruktion. Wenn sich auf der Spiegeloberfläche Tau bildet. wird das Licht von der mit Konstantspannung betriebenen Lampe auf eine Fotozelle reflektiert. Der Strom zum Thermomodul. das den Spiegel kühlt, wird dann abgeschaltet. so dass die Spiegeloberfläche, deren Temperatur gemessen wird, dauernd zwischen dem betauten und taufreien Zustand pendelt.

Dadurch wird verhindert, dass der Spiegel durch Drift in der Schaltung zu einer Temperatur unter der des Taupunktes gekühlt wird. Man nimmt für diese Betriebsweise wesentliche technische Vorteile gegenüber den Servosystemen. die früher in automatischen Messungen der Taupunkttemperatur Verwendung fanden, in Anspruch. Es werden durchgehend Festkörper-Bauelemente benutzt.

EE 73 751 *für weitere Einzelheiten*

Mikrowellen-Abschlüsse

Morganite Resistors Ltd, Bede Trading Estate, Jarrow, Co. Durham

In einem neuen Programm von Hochleistungs-Mikrowellenabschlüssen kleiner Abmessungen sind nunmehr Absorber mit Aluminiumfinnen, die man unter 2,8 kg/cm² Betriebsdruck halten kann, lieferbar. Abschlüsse dieser Type "A" sind für Hohlleitergrössen WG 16 (WR 90) für eine mittlere Sollbelastbarkeit von 500 W bei maximaler Oberflächentemperatur von 300°C lieferbar. Bei einem Stehwellenverhältnis von 0,95 (1,05) oder besser zwischen 8,4 und 10 GHz sind die Aussenabmessungen nur 165 × 89 × 102 mm. Ähnliche Abschlüsse werden zur Zeit für Hohlleiter der Grössen WG 10 (WR 284) bis zu WG 18 (WR 62) entwickelt.

Für Verwendung bei niedriger Belastung steht die Serie "M" zur Verfügung. Diese Absorber haben einen Metallbelag und können in der Grösse WG 16 bei einer Oberflächentemperatur von ca. 600°C eine Leistung von 350 W zerstreuen. Die wassergekühlten Abschlüsse der Type "W" sind für sehr hohe Leistungen konstruiert. In der Grösse WG 16 kann eine Durchschnittsleistung von 1500 W zerstreut werden.

EE 73 752 *für weitere Einzelheiten*

400-Hz-Stromversorgung

Ashgrove Instruments Ltd, 96 Amyand Park Road, Twickenham, Middlesex

(*Abbildung Seite 636*)

Dieses Netzgerät wurde als Stromversorgung für Drehmelder, Drehfeldsysteme, Servomotoren, Tachogeneratoren, Abtastköpfe sowie alle Ausrüstungen, die 400-Hz-Stromversorgung benötigen, entwickelt.

Den Widerstandsregler im Ausgang kann man so einstellen. dass die Ausgangsspannung bei Anlegen der Last steigt, konstant bleibt oder fällt. Auf diese Weise lässt sich der Spannungsabfall in Leitungen, Sicherungen und Messgeräten auf Wunsch kompensieren.

Näherungsweise Korrektur des Leistungsfaktors ist erforderlich. Verbraucher, deren Stromaufnahme oder Leistungsfaktor schnellen Schwankungen unterliegt, beeinflussen das Arbeiten des Gerätes aber nicht.

In Standardausführung wird das Gerät als Einbauchassis geliefert, ist jedoch auf Wunsch auch für Gestelleinbau oder komplett mit Voltmeter und Ammeter im Gehäuse erhältlich.

Die Ausgangsspannung ist 115 V oder 57 V mit ±10 V Regelbereich, die Ausgangsleistung 75 VA.

EE 73 753 *für weitere Einzelheiten*

Oszillografen-Eichgerät

Telequipment Ltd, Chase Road, Southgate, London, N.14

(*Abbildung Seite 636*)

Telequipment hat ein tragbares Oszillografen-Eichgerät C1 eingeführt, das in einem Gerät alle zum Abgleichen von Oszillografen der Telequipment und anderer Marken erforderlichen Wellenformen und Frequenzen erzeugt und mit vollstabilisierten Stromversorgungen betrieben wird.

Bei Abmessungen von 33 × 16,5 × 33 cm wiegt der Typ C1 unter 11 kg und hat vier getrennte Ausgänge, die entweder einzeln oder ohne Wechselwirkung gleichzeitig benutzt werden können. Das Eichgerät lässt sich bei Einsatz nach dem Vergleichsverfahren auch zum Testen anderer Signalquellen einsetzen.

Abgegeben werden die folgenden Signale: eine mit einer Folgefrequenz von entweder 100 kHz oder 1 MHz umgeschaltete Rechteckwellenform plus Zeitmarkenimpulsen mit Folgefrequenzen von 1 MHz, 100 kHz, 10 kHz und 50 Hz, sowie ein in Bezug auf Masse negativgehender Zeitsteuerkamm.

Das Eichgerät hat eine eingebaute, mit 0,2 Prozent Genauigkeit arbeitende

Quarzquelle. Eine positive Fernseh-Wellenform mit ca. 200 Zeilen ohne Zeilensprung und mit 1 V Amplitude (Synchronisierung + Video) steht zur Verfügung.

EE 73 754 *für weitere Einzelheiten*

Subminiatur-Zungenschalter

Flight Refuelling Ltd, Wimborne, Dorset

(Abbildung Seite 637)

Der Hamlin MSRS-2 ist ein Subminiatur - Magnetzungenschalter, der speziell für logisches Schalten mit niedriger Energie entwickelt wurde und von Flight Refuelling Ltd erhältlich ist.

Eine neue Silberlegierung für Kontakte ergibt über Millionen Betätigungen gleichförmigen Kontaktwiderstand, und ausserdem vereinfacht das Abfall-Ansprechverhältnis den Aufbau von UND-Grundschaltungen, wo je nach Zustand der Schaltung Ausfall eines beitragenden Zustandes das Abfallen verursacht oder Ansprechen verhindert.

Tests deuten ein Ansprech-Abfallverhältnis von 66...75 Prozent und einen Höchstkontaktwiderstand von 150 mΩ nach 100 Millionen Betätigungen an. Der Schalter hat eine Nennleistung von 0,5 W– bei 10 mA Höchststrom. Der Schalter kann Kundenwünschen entsprechend mit einer Empfindlichkeit von 20...100 Aw geliefert werden.

Einschliesslich Zuleitungen hat der Schalter eine Gesamtlänge von 57,2 mm, der Schalterkörper eine von 20,3 mm und das Glasrohr einen Durchmesser von 2,3 mm.

EE 73 755 *für weitere Einzelheiten*

Wobbelsender

KLB Electric Ltd, 335 Whitehorse Road, Croydon, Surrey

(Abbildung Seite 637)

Der Wobbelsender und Markengeber PACO G.32 ist ein FM-Messender mit einem Mittenfrequenzbereich von 3 ... 213 MHz in fünf überlappenden Teilbereichen. Im hohen Teilbereich ist die Wobbelbreite von 0 ... 30 MHz regelbar.

Das Gerät ist für das Abgleichen von Breitbandverstärkern aller Art geeignet und hat die wünschenswerte Eigenschaft, dass die Marken eingegeben werden, nachdem das Signal durch den Prüfling gegangen ist. Dadurch können die Markensignale keine irreführenden Ergebnisse hervorrufen, wie das oft bei Sendern dieser Art geschieht.

EE 73 756 *für weitere Einzelheiten*

Kohlemikrofone

Airmed Ltd, Edinburgh Way, Harlow, Essex

(Abbildung Seite 637)

Airmed Ltd hat zwei neue Kohlemikrofone eingeführt, die besonders für moderne Nachrichtensysteme, in denen es hauptsächlich auf niedrigen Preis, kleine Abmessungen und Zuverlässigkeit ankommt, entwickelt wurden.

Typ A.301 ist ein Miniatur-Druckmikrofon mit −50 dB/V/Dyn/cm² Empfindlichkeit bei 1 kHz. Es wiegt unter 10 g und ist durch leichtes Gewicht und kleine Abmessungen für Montage an den Mikrofonbügel von Kopfhörern geeignet. Der Frequenzgang zwischen 200 und 5 000 Hz wurde vor allem auf beste Sprachdeutlichkeit ausgerichtet. Er ist zwischen 200 und 1 000 Hz geradlinig, steigt dann um 8 dB zu einer breiten Spitze zwischen 2 000 und 4 000 Hz und fällt über 5 000 Hz steil ab.

Der abgebildete Typ A.303 ist ein störschallunterdrückendes Druckdifferenzmikrofon hoher Ausgangsleistung für Nahbesprechung in geräuschvoller Umgebung. Ein Druckdifferenzmikrofon spricht auf Schall von einer nahen Quelle—den Lippen—an und diskriminiert gegen Schall, der von weiter entfernt herkommt, d.h. Störgeräusche. Die Empfindlichkeit ist −50 db/V/Dyn/cm² bei 1 kHz, und der Frequenzgang ist zwischen 200 und 4 000 Hz glatt mit breiter Anhebung zwischen 2 000 und 3 000 Hz und steilem Abfall über 4 000 Hz. Das schlagfeste, gespritzte Kunststoffgehäuse hat Klemmkontakte zur Befestigung am Kopfhörerbügel.

EE 73 757 *für weitere Einzelheiten*

Messrelais

Leland Instruments Ltd, 145 Grosvenor Road, Westminster, London, S.W.1

(Abbildung Seite 637)

Das neue kontaktlose Leland-Leroux-Messrelais Sensicon RS1 ist ein Mess- und Steuerrelais in Präzisionsausführung. Es besteht aus einem dem Sensitact-Relais ähnlichen Drehspulmessgerät, das ein geschlossenes, transistorisiertes elektronisches Steuersystem betätigt, das—wenn der Zeiger sich auf einen vorgewählten Skalenwert einstellt—zugeordnete Steuereinheiten in Betrieb setzt.

Das Relais kommt in zwei Typen, von denen Typ N in sechs verschiedenen Ausführungen mit Empfindlichkeiten von 100 µA bei 6 000 Ω bis 10 mA bei 1,5 Ω lieferbar ist.

Typ S wird in fünf verschiedenen Ausführungen mit Empfindlichkeiten zwischen 25 µA bei 6 000 Ω und 2,5 mA bei 1,5 Ω geführt.

Das als Zylinder auf einen Oktalsockel montierte Sensicon ist 71 mm lang und hat einen Durchmesser von 38 mm. Am oberen Ende des aus Isolierstoff gepressten Zylinders ist eine durchsichtige Kunststoffkappe aufgeschraubt, die Beobachtung von Skala, Zeiger und Steuermarke erlaubt. Die Steuermarke wird nach Abschrauben der Kappe durch Verdrehen der die elektronische Schaltung enthaltenden Baugruppe in Bezug auf den Sockel eingestellt und kann auf diese Weise über die ganze Skalenlänge verschoben werden. Auf Wunsch sind diese Einheiten mit in der Fabrik eingestelltem Feststeuerpunkt lieferbar.

EE 73 758 *für weitere Einzelheiten*

Szintillationszähler

Panax Equipment Ltd, Holmethorpe Industrial Estate, Redhill, Surrey

(Abbildung Seite 637)

Das Panax-Programm wurde durch zwei kernphysikalische Zählgeräte mit transistorisierten 6-Dekadenuntersetzern erweitert, die als Universal-Laborgeräte für Geiger- und Szintillationszählungen gedacht sind und beide eine maximale Zählgeschwindigkeit von 50 kHz haben. Mittels eines Drehschalters kann jeder beliebige von neun Zählwerten, die den Bereich 100 ... 1 Million überstreichen, vorgegeben werden. Ausgangssignale der Dekaden, die man koaxialen Buchsen auf der Rückseite des Gerätes entnehmen kann, sind für Speisung passender Druckwerke zum automatischen Ausschreiben nach Ablauf der Zählperiode geeignet. Drucktasten steuern Start und Rückstellung.

Die Hochspannung für Geiger- und Szintillationszähler wird im Gerät erzeugt und ist gegen Netzschwankungen von bis zu ±10 Prozent stabilisiert. Sie kann mittels eines 10gängigen Potentiometers mit genau markiertem Knopf zwischen 0 und 2 kV geregelt werden. Die erforderlichen Diskriminatorvorspannungen werden auch im Gerät erzeugt und auf gleiche Weise eingeregelt. Totzeiten von 4, 20, 200 oder 400 µs lassen sich über einen 4stufigen Drehschalter einstellen.

Die obige Beschreibung trifft auf beide neuen Geräte zu, die mit P.7602 und P.7702 bezeichnet sind. Der Hauptunterschied liegt darin, dass der letztere (abgebildete) Typ ausser dem schnellen Untersetzer noch einen 6-Dekadenzeitgeber hat und daher besser als automatisches Zählgerät (mit vollautomatischen vorwählbaren Zähl- und vorwählbaren Zeiteinrichtungen) bezeichnet wird.

EE 73 759 *für weitere Einzelheiten*

Elektronisches Thermometer

Sidien Products Ltd, 11 Birchwood Court, Edgware, Middlesex

(Abbildung Seite 638)

Dieses elektronische Thermometer hat zwei durch einen Schalter einstellbare Messbereiche von −10°...+90° C und −10°...+150° C.

Jedes Thermometer hat zwei Messfühler, die durch ein 1,22 m langes Miniaturkabel mit dem Gerät verbunden sind; der gewünschte Messfühler wird mittels eines Umschalters angelegt.

Das Gerät wird aus einer Mallory-Batterie MN1300T2 gespeist, die eine Lebensdauer von über zwei Jahren hat.

Die thermische Zeitkonstante ist 20 Sekunden und die Messunsicherheit in der Grössenordnung von ±2° C.

EE 73 760 *für weitere Einzelheiten*

Regeltransformatoren

The Cressall Manufacturing Co. Ltd, Cheston Road, Birmingham 7

(*Abbildung Seite 638*)

Das Torovolt-Programm für Regeltransformatoren der Cressall Manufacturing Co Ltd wurde nunmehr durch ein eingekapseltes Modell in Sparwicklungstechnik ergänzt.

Das neue Torovolt-Modell 33Y ist ein wirtschaftliches und zuverlässiges Mittel zur Steuerung von Wechselstromgeräten und für Dauerbelastung bis zu 0,6 A; es hat die kleinsten mit langer Lebensdauer zu vereinbarenden Abmessungen.

Der Transformator ist für direkten Anschluss an ein 240-V-Wechselspannungsnetz bemessen und ermöglicht gleichförmige Änderungen der Ausgangsspannung von Null bis zur Netzspannung bei einem Höchstnennstrom von 0,8 A. Der bewickelte Kern ist in einem dauerhaften Isolierstoffgehäuse vergossen und hat bei einer Tiefe von 55,6 mm einen Durchmesser von 81 mm. Verbindungen werden über Anschlüsse in mit dem Gehäuse zusammengepressten Blöcken hergestellt.

Modelle sind entweder für Aufbau oder Einbau lieferbar.

EE 73 761 *für weitere Einzelheiten*

Transistorisierte Fernsehkamera

Automatic Information and Data Services Ltd, 26 Sheen Road, Richmond, Surrey

(*Abbildung Seite 638*)

Eine volltransistorisierte Kamera wird von A.I.D.S. Ltd angekündigt. Sie wurde entwickelt, um aus einem 1-Zoll-Vidikon die höchste Leistung herauszuholen, arbeitet mit einer Anodenspannung von unter 250 V und ist für vollen Dauerbetrieb bemessen.

Erfahrungen haben gezeigt, dass Service nach einer anfänglichen Einlaufzeit und bei Umgebungstemperaturen zwischen −50° und +45° C hauptsächlich auf Reinigen der Objektive und Auswechseln der Vidikonröhre nach 4 000 Stunden beschränkt ist. Die Kamera kann auch für Dauerbetrieb bei höheren Umgebungstemperaturen eingestellt werden, jedoch ist in diesem Fall die Leistung des Vidikons bei erhöhten Temperaturen der begrenzende Faktor. Die Kamera ist mit 21 Transistoren und 31 Dioden bestückt, und zwar werden—wo es auf bessere Leistung oder wärmebeständige Eigenschaften ankommt—Siliziumhalbleiter verwendet.

Die nominell für 625 Zeilen und freilaufenden Betrieb konstruierte Kamera kann auch für andere Zeilennormen geliefert werden, und 2:1 Zeilensprung sowohl nach C.C.I.R.-Norm als auch A.I.D.S.-Wellenform kann einem A.I.D.S.-Wellenformgeber entnommen werden.

Die Kamera wird üblicherweise mit einer in einem Spritzgussgehäuse untergebrachten Stromversorgung geliefert und ist vollautomatisch. Für Verwendung mit anderen modularen A.I.D.S.-Einheiten ist auch eine modulare Stromversorgung mit automatisch-manuellem Niveauschalter erhältlich.

Die Kamera hat in der Mitte des Bildes ein Auflösungsvermögen von 650, in den Ecken eins von 400 Fernsehzeilen, wo eine 50% Modulation (schwarz-zu-weiss) bei 330 Fernsehzeilen im Zentrum und 250 in den Ecken vorkommt. (Zur Erhöhung der Modulation in der Mitte um 90% kann eine Blendenkorrektur in modularer Form geliefert werden.)

EE 73 762 *für weitere Einzelheiten*

Schalttafelinstrumente

Smiths Industrial Division, Kelvin House, Wembley Park Drive, Wembley, Middlesex

(*Abbildung Seite 638*)

Diese kleinen Schaltfeldmessgeräte höchster Empfindlichkeit und Präzision sind die jüngsten Ergänzungen des Programmes der Smiths Industrial Division für Registrier- und Laborgeräte.

Die Messwerke der Instrumente haben Spannbandlagerung und sind mit Glaszeigern für Anzeige auf einer sehr deutlichen Skala ausgestattet. Der kleinste Messbereich ist 0...6 μA oder 3...0...3 μA bei beiderseitigem Ausschlag.

Für horizontal montierte Messgeräte mit Skalenendwerten von 25 μA oder mehr kann die Messunsicherheit so niedrig wie 0,5 Prozent sein; eine ähnliche Messunsicherheit haben vertikal montierte Instrumente mit Skalenendwerten von 100 μA oder mehr.

Diese Messgeräte, deren Skalen 63 mm lang sind, haben Standardabmessungen von 72 × 84 mm.

EE 73 763 *für weitere Einzelheiten*

Stromversorgungen für Sende-Empfangsweichen

Ferranti Ltd, Ferry Road, Edinborgh 5

(*Abbildung Seite 639*)

Ferranti Ltd hat eine neue leichte Hochspannungs-Stromversorgung entwickelt, die die zur Aufrechterhaltung der Entladung in der Ferranti-Sende-Empfangsweiche WF42 (CV2311) erforderliche Spannung abgibt. Durch diese Entwicklung wurden die Schwierigkeiten in der Bereitstellung einer ausreichend hohen Gleichspannung in Flugzeugen behoben.

Je eine Ausführung wird für Eingangsspannungen von 115 V, 400 Hz und 200 V, 400 Hz hergestellt, und beide haben eine Zündspannung von 1050 V− ±10%, die nach dem Zünden auf die normale Brennspannung absinkt.

Die Abmessungen des 85 g wiegenden Bausteins sind 3,47 × 3,47 × 3,78 cm einschliesslich der Epoxydharz-Einkapselung. Er entspricht den Anforderungen der Feuchtigkeitsklasse H.6 des Joint Service-Pflichtenblattes und der britischen Vorschrift DEF 5214 für harzvergossene Transformatoren. Der Umgebungstemperaturbereich ist −40°... +100° C, und der Baustein hat den Spannungsdurchschlagtest für eine Höhe von 21 300 m bestanden.

Die Stromversorgung kann Spitzensperrspannungen von 4 kV für 10 μs aushalten und wird durch zufälligen Kurzschluss der Ausgangsklemmen nicht beschädigt.

Varianten in denselben Abmessungen lassen sich für Verwendung mit anderen Ferranti-Sende-Empfangsweichen entsprechend deren technischen Daten herstellen.

EE 73 764 *für weitere Einzelheiten*

Drehmoment-Tester

Mining & Scientific Equipment Ltd, 317 Kennington Road, London, S.E.11

(*Abbildung Seite 639*)

M.S.E. Ltd kündigt die Einführung eines neuen Drehmoment-Messgerätes an. Es wurde speziell für das Testen von Kleinserien sowie die Qualitätskontrolle von Kleinmotoren entwickelt und ist mit einem elektronisch gesteuerten Bremsdynamometer neuer Konstruktion ausgerüstet, mit dem man das Drehmoment unabhängig von der Umdrehungsgeschwindigkeit messen kann.

Das in der Abbildung gezeigte Modell hat einen Drehmoment-Messbereich von 360...3600 g. cm, für den eine Messunsicherheit von 1,5 Prozent des Bereichendwertes angegeben wird; andere Bereiche sind lieferbar. Die Umdrehungsgeschwindigkeit des Motors lässt sich gleichzeitig bis zu 5000 UPM messen und anzeigen.

Die Anzeige erfolgt auf einem Instrument mit offener Skala; sie kann auch auf einem 43-cm-Oszillografenschirm mit langer Nachleuchtdauer dargestellt werden, auf dem direkt eine Geschwindigkeits-Drehmomentkennlinie geschrieben wird.

Auf dem Schirm der Oszillografenröhre kann ein Raster angeordnet werden, mit dessen Hilfe man die Qualität des Motors schnell bestimmen kann. Verwendung eines Schirmes mit langer Nachleuchtdauer ermöglicht das Fotografieren der Kennlinie.

EE 73 765 *für weitere Einzelheiten*

Lötkolben

Vertrieb: Oliver Dow Ltd, 877a High Road, Finchley, London, N.12

(Abbildung Seite 639)

Der in Schweden hergestellte "BLIXT"-Lötkolben hat in seinem Handgriff eine Rolle Kombinationslötdraht, der mittels eines "Pistolenabzuges" zur Lötspitze vorgeschoben wird. So kann der Kolben mit nur einer Hand benutzt werden, und der Arbeiter hat die andere frei, um ein Bauelement, eine Zange oder dergleichen zu halten.

Der Lötkolben wird mit einer Ersatzspitze geliefert und andere sind erhältlich. Der auf eine leicht auswechselbare Kunststoffrolle gewickelte Lötdraht ist 5 m lang.

EE 73 766 *für weitere Einzelheiten*

Füllstandwächter

Vacuum Reflex Ltd, 6 Soho Street, London, W.1

(Abbildung Seite 639)

Vacuum Reflex Ltd hat eine Reihe neuer Geräte für laufende Füllstandanzeige und -regelung entwickelt. Das System arbeitet mit Überschallwellen, die von der Trennschicht einer Flüssigkeit oder eines Schüttgutes reflektiert werden, und ist für Füllstandmessungen von Flüssigkeiten sowie körnigen, festen und pulvrigen Stoffen geeignet.

Zwei Überschallwandler—ein Geber und ein Empfänger—werden über dem Höchststand des Tank- oder Bunkerinhaltes anmontiert. Zur Bestimmung des Füllstandes wird die vom reflektierten Impuls für die Rückkehr von der Trennschicht benötigte Zeitspanne gemessen und auf einem in Füllstand oder Tankinhalt geeichten Messgerät angezeigt. Alle externen Teile sind gekapselt, und die ganze Ausrüstung ist transistorisiert; es werden nur Silizium- oder Planartransistoren verwendet.

Da die Wandler über dem Tankinhalt angebracht sind und die Füllstandmessung nur auf der Reflektion von Überschallwellen durch die Oberfläche des Tankinhaltes beruht, ist das System von der Natur und dem Zustand des Tank- oder Speicherbehälterinhaltes völlig unabhängig. So lässt sich z.B. der Stand von Wasser, Milch, Kohle, Asche, Steinen, Felsbrocken, Salz usw. messen. Das System ist auch zum Messen von Flüssigkeit — Flüssigkeittrennschichten geeignet und wurde bereits zum Messen des Trennschichtniveaus zwischen Flüssigkeit und Schaum in Hochdruckbehältern eingesetzt.

Zur Zeit lieferbare Ausrüstungen haben einen Messbereich von 12,2 m (30,5 m in Flüssigkeiten); Arbeiten zur Erweiterung der Bereiche sind im Gange.

EE 73 767 *für weitere Einzelheiten*

Druckgeber

Consolidated Electrodynamics Division, Bell & Howell Ltd, 14 Commercial Road, Woking, Surrey

(Abbildung Seite 639)

Die Consolidated Electrodynamics Division der Bell & Howell Ltd hat einen Druckgeber mit kleinem Messbereich angekündigt, der speziell für Verwendungszwecke entwickelt wurde, bei denen es auf kleine Abmessungen und geringes Gewicht ankommt.

Vom Druckgeber 4-353 wird behauptet, dass er für seinen absoluten Druckmessbereich von 0 ... 0,07 kg/cm² der kleinste auf dem Markt ist. Er wiegt nur 198 g, ist 63,5 mm lang und hat einen Durchmesser von 38,1 mm.

Der neue Wandler wurde für das Messen von Höhendrücken während des Fluges, Höhenkammerdrücken, Windtunneldrücken und Dampfaustrittsdrücken von Turbinen entwickelt und erstmalig für die Untersuchung der beim Abfeuern einer Rakete in einer Höhenkammer erzeugten Drücke eingesetzt.

Durch eine neue Membranenkonstruktion erzielt der 4-353 hohe Leistungsfähigkeit im kleinsten Raum bei kleinen Abmessungen. Kompensation erfolgt ausserhalb des dichten aktiven Elementes. Durch einen mechanischen Überdruckstopp kann das Instrument ohne Schaden bis zu 1,4 kg/cm² vertragen.

Der Betriebstemperaturbereich ist −73° ... +135°C, die Empfindlichkeit 20 mV, die durch Wärmeempfindlichkeit hervorgerufene Verschiebung über den kompensierten Temperaturbereich ±0,009%/°C und die kombinierte Linearität und Hysterese des gesamten Bereichausganges ±0,5 Prozent.

EE 73 768 *für weitere Einzelheiten*

Klystron-Stromversorgung

Vertrieb: Miles Hivolt Ltd, Old Shoreham Road, Shoreham-by-Sea, Sussex

(Abbildung Seite 640)

Das von Oltronic in Schweden hergestellte Gerät LS525R ist eine für die Mehrzahl der zur Zeit lieferbaren Klystrons geeignete Universal-Stromversorgung.

Folgende Spannungen werden abgegeben:

Anodenspannung	−200 V ... −3,6 kV
Reflektorspannung	0 ... 1 kV, 50 µA
Gitterspannung	−300 V ... +150 V, 5 mA
Heizerspannung	6.3 V, 3 A

Alle Ausgänge haben ausgezeichnete Konstanz und geringe Restwelligkeit: die Restwelligkeit der Reflektorspannung ist niedriger als 200 µV.

Vier Modulationsgeneratoren sind eingebaut und geben Rechteck-, Impuls-, Sägezahn- und Sinuswellenformen ab.

EE 73 769 *für weitere Einzelheiten*

Schall-Dehnungsmesser-Vergleichsgerät

Westland Aircraft Ltd, East Cowes, Isle of Wight

(Abbildung Seite 640)

Ein tragbares Vergleichsgerät für das Testen von Dehnungsmessern mit schwingenden Saiten wird von Westland Aircraft Ltd hergestellt. Die Firma entwickelte es im Zusammenhang mit dem Auftrag, im Kernkraftwerk Oldbury-on-Severn Schall- und Streifendehnungsmesser einzubauen.

Bei der Entwicklung des Vergleichers wurde besonders auf einfache Bedienung und Anzeige Wert gelegt. Die Bedienungselemente sind auf einen Ein-Ausschalter und die Klemmen zum Anschluss des zu prüfenden Dehnungsmessers beschränkt. Das Gerät misst die Frequenz der Schallschwingungen unmittelbar durch Vergleich mit einem internen Quarzkristall und zeigt sie als vierstellige Projektionsdarstellung an. Im Gerät ist eine Energiequelle für das Zupfen des Messdrahtes zusammen mit den Treiberschaltungen vorhanden, die das Messverfahren völlig automatisch machen. Bei Anschluss des Dehnungsmessers an das Vergleichsgerät wird derselbe erregt, gemessen und das Ergebnis angezeigt; dieser Vorgang wird alle vier Sekunden wiederholt.

Durch sein robustes, tragbares Gehäuse und den die Bedienungselemente gegen Transportschaden schützenden Deckel ist das Gerät für den Aussendienst auf Baustellen geeignet. Es ist eine völlig geschlossene Ausrüstung mit eingebauter, aufladbarer Batterie mit Einrichtungen für Anschluss an ein externes Ladegerät. Korrekter Betrieb über einen breiten Umfang von Batteriezuständen wird durch einen internen Konstanthalter gewährleistet.

EE 73 770 *für weitere Einzelheiten*

Konstruktionssystem

A.P.T. Electronic Industries Ltd, Chertsey Road, Byfleet, Surrey

(Abbildung Seite 640)

A.P.T. Electronic Industries Ltd hat ein neues, für elektronische Geräte gedachtes modulares Chassis-Konstruktionssystem eingeführt, das unter dem Namen MINAR auf den Markt kommt. Das System wurde von der British Broadcasting Corporation entwickelt, die es für einige Zeit versuchsweise erprobte. Zugrunde liegt ein Standardfeld von 48,3 cm (19″) × 13,35 cm in Form eines Miniaturgestelles, in das Bausteine verschiedener Grösse eingesetzt werden können. Das System ist vor allem für Transistor- oder Druckkarten-Schaltungen geeignet.

Die hauptsächliche Montageeinheit ist ein Rahmenzusammenbau, den man direkt in ein 19″-Gestell einbauen kann. Im Rahmen sind zur Befestigung von Moduln verschiedener Breite, die alle Vielfache des Rastergrundmasses 7/8″ (22,2 mm) sind, Gewindelöcher vorgesehen. Moduln bis zu 20× Rastermass lassen sich in jeder beliebigen Anordnung in einem Rahmenzusammenbau unterbringen; sie können nach Wunsch entweder fest oder als Einschub eingebaut werden. Die Abmessungen sind so gewählt, dass nach Einbau des ersten Moduls die Befestigungsschrauben der anderen Moduln automatisch mit den entsprechenden Gewindelöchern im Rahmen ausgerichtet sind.

In dem System sind auch Gehäuse und Deckel vorgesehen, in die ein Rahmenzusammenbau montiert werden kann, um die Ausrüstung transportabel zu machen. Auch Hauben sind für einzelne Bausteine oder komplette Ausrüstungen lieferbar. Weiterhin kann man Chassis und Leiterplatten in entsprechenden Abmessungen für die Moduln beziehen.

Die meisten Teile des Systems werden zusammengebaut geliefert, die grössten Moduln aber in Bausatzform zwecks Einsparung von Transportkosten. Für den Zusammenbau, der keine Fachkenntnisse erfordert, benötigt man als einziges Werkzeug einen Schraubenzieher. Einzelteile des Rahmenzusammenbaus, der Moduln usw. können auch einzeln bezogen werden, so dass Kunden die Möglichkeit gegeben ist, mit geringsten Kosten Sonderkonstruktionen auszuführen.

EE 73 771 *für weitere Einzelheiten*

Elektronisches Vielfach-Messgerät

Comark Electronics Ltd, Gloucester Road, Littlehampton, Sussex

(*Abbildung Seite 640*)

In dem elektronischen Vielfach-Messgerät Typ 130 ist die Vielseitigkeit des Universal-Messgerätes mit der Empfindlichkeit des Röhrenvoltmeters kombiniert. Da es batteriebetrieben und durchweg mit Transistoren bestückt ist, bestehen keinerlei Brumm- oder Erdungsprobleme; es ist kompakt und benötigt keine Anheizzeit.

Das Instrument hat 54 Grundbereiche, deren Skalenendwerte zwischen 10 mV und 300 V und zwischen 10 μA und 100 mA Gleich- und Wechselstrom, sowie bei 1 μA und 1 kV— liegen; alle Gleichstrombereiche haben den Nullpunkt in der Skalenmitte.

Der Eingangswiderstand ist ca. 2MΩ für alle Wechselstrombereiche und 1 MΩ/V für alle Gleichstrombereiche mit Ausnahme des höchsten. Der Spannungsabfall an den Klemmen liegt für alle Strombereiche unter 12 mV. Rauschen ist vernachlässigbar klein; eine Ausnahme bildet jedoch der empfindlichste Wechselspannungsbereich, für den es bei Speisung aus einer 100-kΩ-Quelle 1 mV nicht überschreitet. Der Frequenzgang ist bis zu 50 kHz linear und bis zu 250 kHz nutzbar.

Widerstandsmessungen erfolgen in fünf Bereichen mit Vollausschlag von 100 Ω bis zu 1 MΩ. Bei Benutzung externer Batterien kann man Widerstände bis zu 10 GΩ und bis zu 0,01 Ω herunter messen.

Eine Eigenschaft des Messgerätes ist die Möglichkeit, mit ihm dreipolige Widerstandsmessungen durchzuführen; der Widerstand kann in Gegenwart eines Nebenschlusses zum dritten Pol, z.B. zur Masse, direkt gemessen werden.

Das Gerät kann sowohl in Gleichstrom- wie Wechselstrombereichen als Spannungs- oder Stromverstärker Verwendung finden. Der Ausgang ist für den mit dem Bereichschalter eingestellten Eingang ca. 1 V.

Die Schaltung besteht aus einem Hochleistungs-Allstromverstärker, der ein mit einer 81 mm langen Spiegelskala ausgestattetes 100-μA-Messgerät treibt. Für Gleichstrommessungen ist ein Transistor-Chopperverstärker vorgeschaltet, der sehr niedrige Drift gewährleistet. Für Wechselstrommessungen wird eine hochohmige Eingangsstufe benutzt. Die Verstärkung wird durch starke Gegenkopplung stabilisiert.

EE 73 772 *für weitere Einzelheiten*

Digital-Drehgeber

Vertrieb: B & K Laboratories Ltd, 4 Tilney Street, Park Lane, London, W.1

(*Abbildung Seite 641*)

Der Peekel-Digital-Drehgeber PP3A1 ist als Zusatz für Potentiometerschreiber oder ähnliche Apparate gedacht, in denen Analoginformation mit der Winkelstellung einer Drehwelle in Beziehung steht. Der PP3A1 setzt diese für direkte Speisung in eine zusätzliche Anzeigevorrichtung oder Darstellung in Digitalform um.

Der PP3A1 wird mittels Spannvorrichtungen und einer Montageplatte an den Schreiber oder das andere Instrument angebaut und kann entweder direkt getrieben oder mittels eines Seiles und Seilscheiben usw. mit der Welle gekuppelt werden. Für den Bruel und Kjoer-Schreiber 2305 steht ein Spezialmodell zur Verfügung.

Der Drehgeber PP3A1 gibt für eine vollständige Umdrehung der Welle einen Ausgang in Form von Dreidekaden-Dezimalinformation von 0...999 ab. Dieser Ausgang kann ohne weitere Entschlüsselung in die Peekel-Digitalanzeige PP5C1 gespeist werden. Er kann über den Peekel-Druck-Umsetzer PP9BA1 ein Digitaldruckwerk, oder über den Peekel - Lochstanzen - Umsetzer PP9F1 eine Lochstanze treiben. Andere Peekel-Geräte, die man mit dem Adapter PP3A1 einsetzen kann, sind z.B. ein Digitalspeicher, eine Digitaluhr, ein Impulszähler und ein Eingangskommutator, der Anschluss von bis zu 200 Kanälen an ein Digitalinstrument oder -system ermöglicht.

Die Abbildung zeigt interne und externe Ansichten des Drehgebers.

EE 73 773 *für weitere Einzelheiten*

Kleinstrelais

Plessey-UK Ltd, Abbey Works, Titchfield, Fareham, Hampshire

(*Abbildung Seite 641*)

Ein neues Kleinstrelais—Typ CF— der Plessey-UK Ltd hat die Typengenehmigung nach der britischen Vorschrift DEF 5165 erhalten. Es ist ein völlig abgedichtetes Relais mit Drehbetätigung und zweipoligem Umschalter, dessen maximaler Kontaktwiderstand von 0,030 Ω durch vergoldete Zwillingskontakte gewährleistet wird.

Verbindungsstifte haben den Rasterabstand von 2,5 mm für Montage auf gedruckte Schaltungen und werden durch Glas-Metallverschmelzungen herausgeführt. Die Spulen- und Kontaktverbindungen mit diesen Stiften sind symmetrisch angeordnet, so dass korrekte Arbeitsweise unabhängig von der Richtung, in der das Relais in die Fassung eingesteckt wird, gewährleistet ist.

Die Schaltleistung ist je nach Anzahl der Betätigungen 1...3 A bei 28 V— oder 115 V~ (rein ohmisch).

EE 73 774 *für weitere Einzelheiten*

Gedruckter Schalter

NSF Ltd, 31-32 Alfred Place, London, W.C.1

(*Abbildung Seite 641*)

NSF kündigt einen einpoligen Drehschalter mit zehn Stellungen in Druckschaltungstechnik mit Daumenradantrieb an, der für Instrumentenausrüstungen und Steueranlagen geeignet ist. Der Schalter ist entweder als Einzelbauelement, oder als Mehrfachschalter mit bis zu vier Ebenen mit einem Daumenradantrieb, oder als übereinander montierte Baugruppe von bis zu zwölf Einzelschaltern lieferbar. Das Rastwerk ist mit Doppelkugeln ausgeführt und hat einen Rastwinkel von nur 36°; ein Stopp ist vorhanden, um das Drehen auf jede Stellung zwischen 2 und 10 zu begrenzen. Die Kontakte schliessen nicht kurz.

Die Standard-Druckschaltungsplatte ist für einpoliges Schalten in zehn Schaltstellungen. Spezialplatten können im Sonderauftrag hergestellt werden. Als Anschlüsse sind Lötösen vorgesehen, auf Wunsch kann man aber auch Steckverbindungen für gedruckte Schaltungen mit 5,1 mm Rastermass benutzen.

Für universelle Montage sind den verschiedenen Plattendicken entsprechende Schlitze vorhanden.

Die Daumenräder sind aus schwarzem Phenolharz gepresst, haben weisse Ziffern, und die Deckplatte besteht aus grauem Delrin, kann jedoch auf Wunsch auch in anderen Farben geliefert werden.

EE 73 775 *für weitere Einzelheiten*

Mehrbereich-Elektrometer

Thomas Industrial Automation Ltd, Station Building, Altrincham, Cheshire

(Abbildung Seite 641)

In ihrem Elektrometer-Programm hat Thomas Industrial Automation Ltd ein neues Modell eingeführt. Dieses neue Instrument komplementiert das Modell VC.99, behält dessen hervorstechende Eigenschaften und technische Daten bei, bietet aber zusätzlich einen Behreichverdopplerschalter. Die Funktionen des mit VC99A bezeichneten Modells sind:

a) Als Mikrostrommesser misst es Ströme von 1 pA bis zu 600 μA in acht umschaltbaren Bereichen. Der höchste Bereich überlappt daher mit dem empfindlichsten eines herkömmlichen Vielfachmessgerätes. Die Anwendung starker Gegenkopplung gewährleistet bei der Strommessung einen kleinen Spannungsabfall.

(b) Als Elektrometer-Voltmeter misst es unmittelbar Potentiale beider Polaritäten von 0...2000 mV in zwei umschaltbaren Bereichen mit einem Äquivalenteingangswiderstand, der grösser als 10^{13} Ω ist.

Da ein Elektrometer äusserst niedrige Ströme messen kann, lässt es sich auch ohne Schwierigkeiten zum Messen sehr hoher Widerstände einsetzen. Mit einer externen Potentialquelle kann man Widerstände bis zu 10^{14} Ω messen. Wenn Testspannungen von 500 V vorgeschrieben sind—wie z.B. für gewisse Kondensatoren und Kabel—kann man sich eines lieferbaren hochkonstanten Spezialadapters für 500 V bedienen und mit dessen Hilfe Messungen bis zu 5×10^{14} Ω durchführen.

Das Instrument ist in einem robusten, mit Traggriffen und abnehmbarem Deckel ausgerüsteten Gehäuse aus Burma-Teakholz oder schwarzem Schichtstoff untergebracht.

EE 73 776 *für weitere Einzelheiten*

Zusammenfassung der wichtigsten Beiträge

Die Mikroelektronik in der Gerätekonstruktion von S. S. Forte

Die mit beispielloser Geschwindigkeit fortschreitende Miniaturisierung elektronischer Ausrüstungen wurde durch die Entwicklung der neuen Festkörpertechnik und durch Betonung der Forderungen nach grösserer Zuverlässigkeit verursacht.

Es ist beabsichtigt, im ersten Teil dieses Beitrages die derzeitig zur Verfügung stehenden Methoden zur Verbesserung der Zuverlässigkeit und Reduzierung der Grösse zu besprechen und anschliessend einen versuchsmässigen Digitalrechner, in dem die auf diese Ausrüstungsgruppe anwendbaren Konstruktionsmethoden verwendet werden, zu beschreiben. Im zweiten Teil sollen die Stufen in der Entwicklung linearer Schaltungen für eine versuchsweise mikrominiaturisierte Bordnavigationshilfe diskutiert werden.

Zusammenfassung des Beitrages auf Seite 586-590

Automatisches Schalten empfindlicher Thermoelemente von J. L. Goldberg und H. M. King

Eine Vorrichtung zum automatischen Schalten empfindlicher Thermoelemente in von wenigen Sekunden bis zu mehreren Zehnerminuten kontinuierlich regelbaren Intervallen wird beschrieben. Der mechanische Teil besteht aus der Ratschenschaltung eines Drehwählers, die über ein Getriebe mit der Welle eines Thermoelementschalters hoher Qualität gekoppelt ist. Eine elektronische Steuerung beaufschlagt die Drehwählerspule periodisch mit einer dem Übersetzungsverhältnis entsprechenden Anzahl von Impulsen, wodurch korrektes Rasten des Thermoelementschalters bewirkt wird.

Die Vorrichtung wurde für das automatische Registrieren von Temperaturen und Temperaturgefällen von Strichmassen und Messbandnormalen mittels Thermoelementen entwickelt.

Zusammenfassung des Beitrages auf Seite 591-595

Eine einfache Kapazitätsmessschaltung mit digitaler Darstellung der Kapazitätswerte von S. L. Hurst

Ein einfaches, sehr schnelles und narrensicheres Verfahren zum Messen von Kapazitäten im Bereich von ungefähr 0,001 μF bis 10 μF mit einer Messunsicherheit von ± 2 Prozent wird beschrieben. Innerhalb kleinerer Messbereiche ist eine viel höhere Messgenauigkeit möglich. Das Verfahren kann auch auf das Messen von Widerstandswerten ausgedehnt werden, unterliegt aber möglicherweise bei dieser Anwendung der Begrenzung minderwertigerer Genauigkeit. Das Zusammenpassen von Kondensatoren und Widerständen kann auch mit grosser Genauigkeit vorgenommen werden.

Zusammenfassung des Beitrages auf Seite 596-599

Integrier- und Abwurfschaltung für Serienkorrelatoren von R. M. Seeley

Es besteht Bedarf an einem endlichen Integrator für Verwendung in Serien-Digitalkorrelatoren. Die Hauptanforderungen sind die Fähigkeit, hohe Bit-Geschwindigkeiten mit konstanter Integration für regellose Geschwindigkeiten aufzunehmen, schneller und vollständiger Abwurf, sowie konstante Summenübertragung zu einem niederohmigen Ausgang. Ein diesen Bedarf deckendes Gerät wird hier

Zusammenfassung des Beitrages auf Seite 601-604

beschrieben und ausserdem eine neuartige schnelle Flip-Flopschaltung bekanntgegeben. Die durchschnittliche Genauigkeit bei einem 2-MHz-Datenstromeingang, 300 µs für Integrieren und 2 µs für Abwerfen, ist 2 Prozent.

Eine Konstant-Stromversorgung für 100 Volt, 100 mA mit Transistoren von R. E. Aitchison und W. S. Lamond

Zusammenfassung des Beitrages auf Seite 604-605

Dieser Beitrag beschreibt einen Konstant-Stromversorgungsmodul für 100 V, 100 mA. Die Konstruktion illustriert einige Prinzipien, die auf Entwurf und Schutz von Transistor-Stromversorgungsschaltungen angewendet werden können, wenn die Netzspannung höher ist als die Nenndurchbruchspannung der Transistoren. Eine Anzahl dieser Moduln können für eine Stromversorgung höherer Spannung in Kaskade geschaltet werden.

Folgezugriff-Kernspeicher mit Schrittschalteradressieren von M. D. A. B. Rackowe

Zusammenfassung des Beitrages auf Seite 612-615

Ein preiswerter Digitalspeicher wird aus einem Schrittschalter in Verbindung mit einem Kernspeicher und Halbleiterschaltungen gebildet. Der Speicher arbeitet nach dem Wortzugriff-Verfahren und hat in der beschriebenen Ausführung eine Kapazität von 10 Worten von je 10 Bits Länge.

Die sich aus der Anwendung des Schrittschalteradressierens ergebende Unkompliziertheit kann nur durch Geschwindigkeitseinbusse erzielt werden, und beliebiger Zugriff wäre auch nur mit Schwierigkeiten zu erreichen.

Ein Impulsphasenmesser von R. E. King

Zusammenfassung des Beitrages auf Seite 615-616

In einer beschriebenen einfachen Schaltung mit Halbleitern ist die Ausgangsanzeige proportional zur Eintreff-Zeitdifferenz zweier Eingangs-Binärimpulsreihen. Die Schaltung umfasst eine herkömmliche bistabile sowie eine Differentialtorschaltung und gibt einen Ternärausgang ab. Mit einem zusätzlichen Vorverstärker in jedem Eingangskanal kann man das Gerät als Tonfrequenz-Phasenmesser benutzen.

Einige auf einen Differentialverstärker anwendbare Transformationen der Knotenadmittanzmatrix eines Netzwerkes von K. G. Nichols

Zusammenfassung des Beitrages auf Seite 617-621

Eine Methode für das Aufstellen der Admittanzmatrix eines Netzwerkes mit k Eingangs- und Ausgangsklemmen, in dem die Klemmen völlig voneinander getrennt sind, wird beschrieben. Das Ergebnis wird dann auf die Leitwertparameter eines Vierpols spezialisiert. Nach dem gleichen Verfahren, jedoch mit einer anderen Transformation der Admittanzmatrix, kann der phasenverschobene und phasenstarre Frequenzgang eines "long-tailed pair" oder Differenzverstärkers analysiert werden.

Zwei frequenzselektive Transistorverstärker von S. Harkness

Zusammenfassung des Beitrages auf Seite 622-626

Der Beitrag beschreibt zwei geschlossene frequenzselektive Transistorverstärker. Beide sind Abwandlungen der von Wigan beschriebenen Wien-Brücke, in der die Frequenz mittels eines Regelwiderstandes eingestellt wird, wodurch die durch gleichzeitige Regelung mehrerer Widerstände auftretenden Schwierigkeiten des genauen Gleichlaufs vermieden werden. Die Verstärker sind besonders für Verwendung als Detektoren in Wechselstrombrücken geeignet. Einer ist für Verwendung in der Prüfung auf Eisenverluste bestimmt, wo es nicht auf hohe Empfindlichkeit, dagegen wegen des Vorkommens sehr grosser ungeradzahliger Harmonischer auf sehr hohe Selektivität ankommt. Der in drei Bereichen überstrichene Frequenzumfang ist 20 . . . 600 Hz. Der Ausgang wird auf einem Drehspulinstrument angezeigt, auf dem 2 Prozent des Skalenendwertes 5 µV entsprechen. Die andere Schaltung ist ein Mehrzweckdetektor für Tonfrequenzbrücken höchster Präzision. Hier wird hohe Empfindlichkeit betont, und der Vollausschlag des im Ausgang liegenden Messgerätes enspricht einem Signal von ca. 1/2 µV. Die Frequenz von 20 Hz . . . 20 kHz wird in sechs Bereichen überstrichen.

Versuche mit einer X-Band-Granat-Verzögerungsleitung von J. H. Collins, B. Yazgan und J. Cochrane

Zusammenfassung des Beitrages auf Seite 627-629

Versuche, in einer achsenmagnetisierten Einkristallscheibe aus Yttriumeisengranat von 3,18 mm Durchmesser, die bei Raumtemperatur in eine TM_{010} Resonanzkammer eingesetzt wurde, mit Impulsen spinakustische Wellen zu erregen, werden beschrieben. Der Mindesteinfügungsverlust des Elementes war 45 dB über einen Eingangsleistungsbereich von −36 . . . +7 dBm.

Ein elektronischer Abtaster für Transistorpaare für symmetrische Vorverstärker von R. R. Vierhout und A. J. H. Vendrik

Zusammenfassung des Beitrages auf Seite 630-631

Es wird gezeigt, das der den Unterdrückungsfaktor symmetrischer Spannungs-Vorverstärker hauptsächlich bestimmende Transistor-Parameter $r_b^* = h_{11}/(h_{21}+1)$ ist.

Ein elektronischer Abtaster, der diesen Wert gegenüber dem Emitterstrom gleichzeitig für zwei Transistoren darstellt, wurde entwickelt. Weiterhin wird eine Eichmöglichkeit besprochen.

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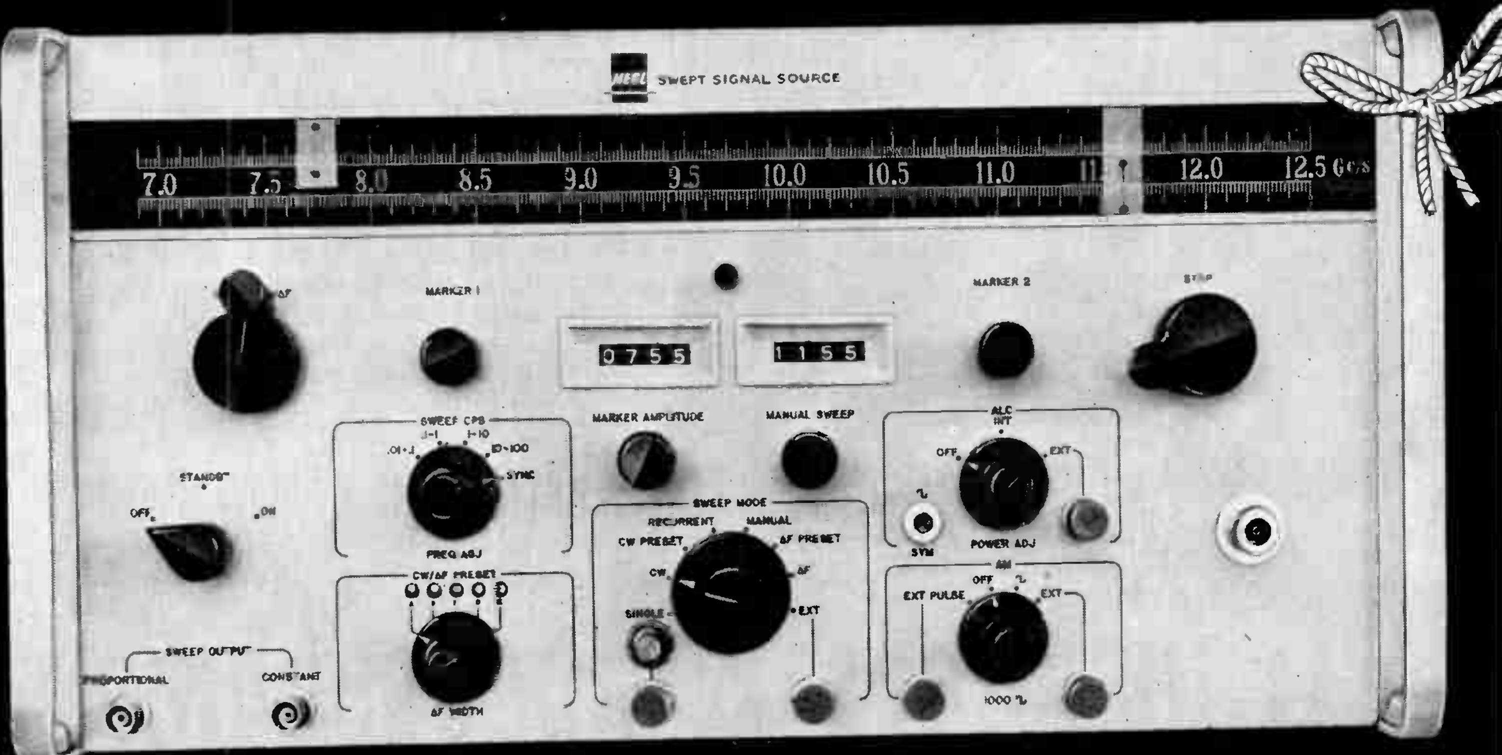

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SPECIFICATIONS

- **SWEEP VOLTAGE OUTPUT** ± 1% linear sawtooth + 25v peak amplitude
- **EXTERNAL FM** + 100v for full sweep width. D C to 10 Kc/s response. Stable + 150v rear panel output provided for use with external voltage divider
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Frequency Range (Gc/s)	Minimum Power Output mW	Model Number
0.5-1.0	50	MP880C
1.0-2.0	100	ML880C
1.0-4.0	25	MLS880C
1.7-4.2	20	MN880C
2.0-4.0	100	MS880C
2.4-5.3	75	MT880C
3.7-8.3	10	MW880C
4.0-8.2	25	MC880C
5.3-10.0	25	MJ880C
7.0-12.4	20	MH880C
8.0-12.4	25	MX880C
10.0-16.0	10	MD880C
12.4-18.0	10	MU880C
15.0-20.0	10	MY880C
18.0-26.5	5.0	MK880C
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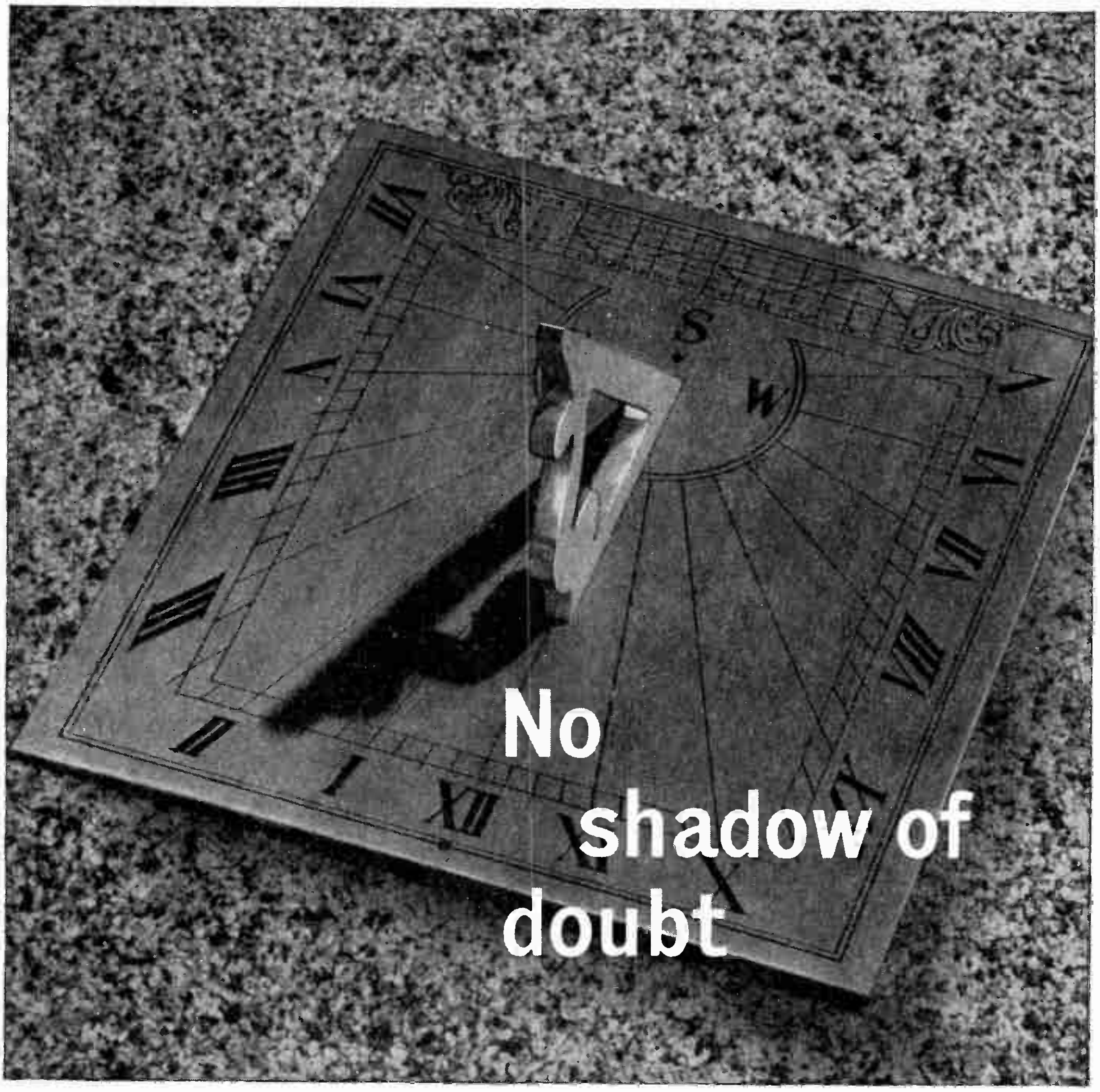

The Pullin Clear View range gives readability beyond the shadow of a doubt; because there can't be shadows. The case is translucent, the dial designed to be seen . . . not blurred.

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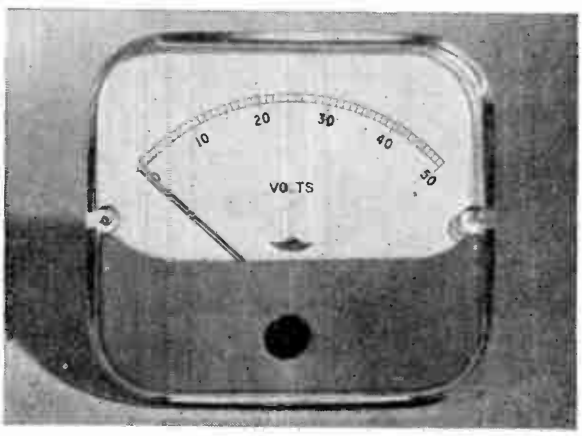

The new colour coded Series 35TC($3\frac{1}{2}$" *scale*) in standard or special colours—*with internal magnet moving coil movements and moving coil and moving iron movements, or with dynamometer wattmeter A.C. and D.C.*

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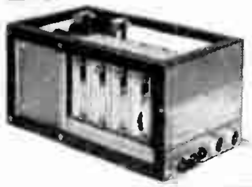
VACUUM TUBE

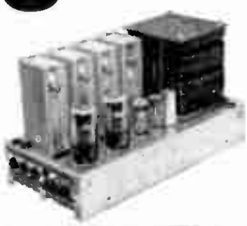
B Series rugged reliable modules



The B series is probably the most advanced open chassis design available. Extremely tough construction inter-services approved reliable valves and silicon junction rectifiers give durability coupled with the high efficiency of series pentode circuits. All units have an impedance of 0.1Ω and a stability of 0.02%.

Model	DC output (Floating) Volts	Current	AC outputs	Ripple mV	Size Inches W	D	H	UK price £
B series rugged reliable modules								
B101/100	100	30-150mA	2×6.3V3A CT	1.0	7½	9⅞	6½	55
B101/150	150	30-150mA	2×6.3V3A CT	1.0	7½	9⅞	6½	55
B101/200	200	0-150mA	2×6.3V3A CT	1.0	7½	9⅞	6½	55
B101/250	250	0-150mA	2×6.3V3A CT	1.0	7½	9⅞	6½	55
B101/300	300	0-150mA	2×6.3V3A CT	1.0	7½	9⅞	6½	55
B101/350	350	0-100mA	2×6.3V3A CT	1.0	7½	9⅞	6½	55
*B101/400/450/500	*	0-150mA	2×6.3V3A CT	1.0	7¾	10¾	7¾	75
*B101/550/600/650/700/750	*	0-150mA	2×6.3V3A CT	2.0	7¾	11½	8½	95
*B101/800/850/900/950/1000	*	0-150mA	2×6.3V3A CT	3.0	7¾	14½	8¾	120
B102/100	100	30-300mA	2×6.3V5A CT	1.0	7½	14⅜	6½	75
B102/150	150	30-300mA	2×6.3V5A CT	1.0	7½	14⅜	6½	75
B102/200	200	0-300mA	2×6.3V5A CT	1.0	7½	14⅜	6½	75
B102/250	250	0-250mA	2×6.3V5A CT	1.0	7½	14⅜	6½	75
B102/300	300	0-250mA	2×6.3V5A CT	1.0	7½	14⅜	6½	75
B102/350	350	0-200mA	2×6.3V5A CT	1.0	7½	14⅜	6½	75
*B102/400/450/500	*	0-300mA	2×6.3V5A CT	1.0	7¾	14½	8¾	95
*B102/550/600/650/700/750	*	0-300mA	2×6.3V5A CT	2.0	8	16⅜	9½	120
*B102/800/850/900/950/1000	*	0-300mA	2×6.3V5A CT	3.0	8⅜	17⅜	9½	140
B103/100	100	30-500mA	2×6.3V5A CT	1.0	8	16½	6½	95
B103/150	150	30-500mA	2×6.3V5A CT	1.0	8	16½	6½	95
B103/200	200	0-500mA	2×6.3V5A CT	1.0	8	16½	6½	95
B103/250	250	0-500mA	2×6.3V5A CT	1.0	8	16½	6½	95
B103/300	300	0-500mA	2×6.3V5A CT	1.0	8	16½	6½	95
B103/350	350	0-400mA	2×6.3V5A CT	1.0	8	16½	6½	95
*B103/400/450/500	*	0-500mA	2×6.3V5A CT	1.0	8⅜	17⅜	9½	120
*B103/550/600/650/700/750	*	0-500mA	2×6.3V5A CT	2.0	11⅜	17⅜	10	150
*B103/800/850/900/950/1000	*	0-500mA	2×6.3V5A CT	3.0	13	17⅜	10	175
B104/100	100	100mA-1A	2×6.3V5A CT	1.0	11	17⅞	8½	160
B104/150	150	100mA-1A	2×6.3V5A CT	1.0	11	17⅞	8½	160
B104/200	200	0-1A	2×6.3V5A CT	1.0	11	17⅞	8½	160
B104/250	250	0-1A	2×6.3V5A CT	1.0	11	17⅞	8½	160
B104/300	300	0-1A	2×6.3V5A CT	1.0	11	17⅞	8½	160
B104/350	350	0-800mA	2×6.3V5A CT	1.0	11	17⅞	8½	160

S Series single and twin regulated supplies



The S series embodies the same advanced design as the B series but originally included only twin units. Owing to their popularity they are now also offered on their own for single output voltages.

Model	DC output (Floating) Volts	Current	AC outputs	Ripple mV	Size Inches W	D	H	UK Price £
S series open chassis single output units								
S101/100	100	30-150mA	2×6.3V3A CT	1.0	6¾	10⅜	7 3/16	48
S101/150	150	30-150mA	2×6.3V3A CT	1.0	6¾	10⅜	7 3/16	48
S101/200	200	0-150mA	2×6.3V3A CT	1.0	6¾	10⅜	7 3/16	48
S101/250	250	0-150mA	2×6.3V3A CT	1.0	6¾	10⅜	3/16	48
S101/300	300	0-150mA	2×6.3V3A CT	1.0	6¾	10⅜	7 3/16	48
S101/350	350	0-100mA	2×6.3V3A CT	1.0	6¾	10⅜	7 3/16	48
S102/100	100	30-300mA	2×6.3V5A CT	1.0	6¾	14⅛	7 3/16	66
S102/150	150	30-300mA	2×6.3V5A CT	1.0	6¾	14⅛	7 3/16	66
S102/200	200	0-300mA	2×6.3V5A CT	1.0	6¾	14⅛	7 3/16	66
S102/250	250	0-250mA	2×6.3V5A CT	1.0	6¾	14⅛	7 3/16	66
S102/300	300	0-250mA	2×6.3V5A CT	1.0	6¾	14⅛	7 3/16	66
S102/350	350	0-200mA	2×6.3V5A CT	1.0	6¾	14⅛	7 3/16	66
S103/100	100	30-500mA	2×6.3V5A CT	1.0	6¾	17⅜	8⅜	88
S103/150	150	30-500mA	2×6.3V5A CT	1.0	6¾	17⅜	8⅜	88
S103/200	200	0-500mA	2×6.3V5A CT	1.0	6¾	17⅜	8⅜	88
S103/250	250	0-500mA	2×6.3V5A CT	1.0	6¾	17⅜	8⅜	88
S103/300	300	0-500mA	2×6.3V5A CT	1.0	6¾	17⅜	8⅜	88
S103/350	350	0-400mA	2×6.3V5A CT	1.0	6¾	17⅜	8⅜	88

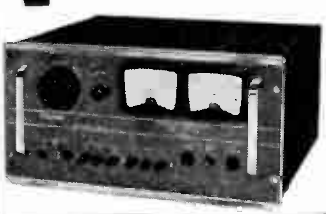
Model		Size Inches W	D	H	UK Price £
S Series front panel twin floating output units					
2 S101/volts S101/volts	Electrically identical to their single voltage counterparts. When ordering please state voltages required. Thus, a 300V, 500mA and 250V, 150mA twin unit would be 2 S103/300 S101/250. There is no need to state the required polarity as both sections are floating and completely, independent from each other.	19	10⅜	7	136
2 S102/volts S102/volts		19	14⅛	7	172
2 S101/volts S103/volts		19	17⅜	7	216
2 S101/volts S102/volts		19	14⅛	7	154
2 S102/volts S103/volts		19	17⅜	7	194
2 S103/volts S101/volts		19	17⅜	7	176

M Series regulated supplies

This series employs a chokeless design offering small size, excellent high frequency response (with stable operation down to at least 1 cps), reliability at high temperatures and safe operation with mains voltage changes up to ±10%. Stability 0.02%, Resistance 0.1Ω, Ripple Voltage 1.0 mV.

Model	DC output (Floating) Volts	Current	AC outputs	Ripple mV	Size Inches W	D	H	UK Price £
M series open chassis units								
M22	150	0- 20mA	nil	1.0	4¾	3¾	4½	26
M22A	200	0- 20mA	nil	1.0	4¾	3¾	4½	26
M23	150	0- 50mA	6.3V2A & 6.3V1A	1.0	6	5½	5¾	27.10
M23A	200	0- 50mA	6.3V2A & 6.3V1A	1.0	6	5½	5¾	27.10
M33	250	0- 50mA	6.3V2A & 6.3V1A	1.0	6	5½	5¾	27.10
M33A	300	0- 50mA	6.3V2A & 6.3V1A	1.0	6	5½	5¾	27.10
M14A	100	0-100mA	2×6.3V3A CT	1.0	10¾	7 3/16	6½	63
M24	150	0-100mA	2×6.3V3A CT	1.0	6¾	7⅜	6½	33.10
M24A	200	0-100mA	2×6.3V3A CT	1.0	6¾	7⅜	6½	33.10
M34	250	0-100mA	2×6.3V3A CT	1.0	6¾	7⅜	6½	33.10
M34A	300	0-100mA	2×6.3V3A CT	1.0	6¾	7⅜	6½	33.10
M25	150	0-200mA	2×6.3V3A CT	1.0	8½	8¾	7	52.10
M25A	200	0-200mA	2×6.3V3A CT	1.0	8½	8¾	7	52.10
M35	250	0-200mA	2×6.3V3A CT	1.0	8½	8¾	7	52.10
M35A	300	0-200mA	2×6.3V3A CT	1.0	8½	8¾	7	52.10
M26S	350	30-500mA	2×6.3V5A CT	1.0	14 1/16	11½	8 1/16	88
M26SA	200	0-500mA	2×6.3V5A CT	1.0	14 1/16	11½	8 1/16	88
M36S	250	0-500mA	2×6.3V5A CT	1.0	14 1/16	11½	8 1/16	88
M36SA	300	0-500mA	2×6.3V5A CT	1.0	14 1/16	11½	8 1/16	88
M28	150	30-350mA	2×6.3V5A CT	1.0	13 3/16	8¾	7½	66
M28A	200	0-350mA	2×6.3V5A CT	1.0	13 3/16	8¾	7½	66
M38	250	0-350mA	2×6.3V5A CT	1.0	13 3/16	8¾	7½	66
M38A	300	0-350mA	2×6.3V5A CT	1.0	13 3/16	8¾	7½	66
M series front panel units								
M26	150	30-500mA	2×6.3V5A CT	1.0	19	12½	8¾	97.10
M26A	200	0-500mA	2×6.3V5A CT	1.0	19	12½	8¾	97.10
M36	250	0-500mA	2×6.3V5A CT	1.0	19	12½	8¾	97.10
M36A	300	0-500mA	2×6.3V5A CT	1.0	19	12½	8¾	97.10
M27	150	50mA-1A	2×6.3V5A CT	1.0	19	16½	8¾	165
M27A	200	0-1A	2×6.3V5A CT	1.0	19	16½	8¾	165
M37	250	0-1A	2×6.3V5A CT	1.0	19	16½	8¾	165
M37A	300	0-1A	2×6.3V5A CT	1.0	19	16½	8¾	165
M29	150	100mA-2A	2×6.3V5A CT	1.0	19	18½	10½	255
M29A	200	0-2A	2×6.3V5A CT	1.0	19	18½	10½	255
M39	250	0-2A	2×6.3V5A CT	1.0	19	18½	10½	255
M39A	300	0-2A	2×6.3V5A CT	1.0	19	18½	10½	255

V Series variable voltage regulated supplies



The V series of laboratory units offers several exceptionally stable and one moderate specification design.

Model	Main DC output (Floating) Volts	Current	Negative Rail Volts	Current	Size Inches W D H	UK Price £
V series variable voltage regulated supplies						
V50-20D	0 to 500	0-200mA (restricted to 150mA below 200V)	0 to 85 via 100k pot.		19 10 7 Panel height 7"	79
V50-20R	200 to 500	0-200mA	nil		19 10 7 Panel height 7"	79
V50-50	0 to 500	0-500mA	250	0-50mA	19¾ 14½ 12 Panel height 10½"	155
2V35-25	(a) 0 to 350	0-250mA (restricted to 150mA below 200V)	250	0-40mA	19¾ 16¾ 10½ Panel height 8¾"	
	(b) 0 to 350	0-250mA (restricted to 150mA below 200V)				198
V100-25	200 to 1000	0-250mA	nil		19¾ 18½ 10½ Panel height 8¾"	170
V200-12	200 to 2000	0-125mA	nil		19¾ 21 15½ Panel height 14"	265

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SILICON

Variable voltage single and twin output instruments

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Output voltage is smoothly controlled from zero to maximum in one continuous sweep and a fine control is included. All units above 1A have four terminal systems, the twin voltage units having floating, independently controlled outputs.

Model	DC output (Floating) Amps	Volts	Ripple mV R.M.S.	Resistance Ohms	Size Inches W	D	H	UK Price
T series variable voltage single output instruments								£
T244	0.33	10-300	2.0	0.5	12⅞	10	6⅞	125
T245	0.5	10-200	2.0	0 5	12⅞	10	6⅞	125
T246	1	0-100	2.0	0.2	12⅞	10	6⅞	125
T247	2	0-50	1.0	0.05	12⅞	10	6⅞	125
T248	3	0-30	1.0	0.03	12⅞	10	6⅞	125
T229	0.5	10-300	2.0	0.5	19⅞	10	6⅞	145
T230	1	10-150	2.0	0.2	19⅞	10	6⅞	145
T236	2	0-75	2.0	0.08	19⅞	10	6⅞	145
T237	3	0-50	2.0	0.03	19⅞	10	6⅞	145
T238	5	0-30	1.0	0.02	19⅞	10¾	6⅞	145
T231	1	10-300	2.0	0.3	19⅞	12¾	6⅞	180
T232	2	10-150	2.0	0.2	19⅞	12¾	6⅞	180
T233	3	0-100	2.0	0.1	19⅞	12¾	6⅞	180
T234*	5	0-60	2.0	0.02	19⅞	12¾	6⅞	180
T235*	10	0-30	1.0	0.01	19⅞	12¾	6⅞	180
T239	1.5	10-300	2.0	0.2	19⅞	14	8¾	265
T240	3	10-150	2.0	0.1	19⅞	14	8¾	265
T241*	5	0-100	2.0	0.02	19⅞	14	8¾	265
T242*	10	0-50	2.0	0.01	19⅞	14	8¾	265
T243*	15	0-30	2.0	0.007	19⅞	14	8¾	265
T series variable voltage twin output instruments								
T249*	0.5/0.5	10-300/10-300	2.0	0.5	19⅞	15¼	10¼	255
T250*	1/1	10-150/10-150	2.0	0.2	19⅞	15¼	10¼	255
T251*	2/2	0-75/0-75	2.0	0.08	19⅞	15¼	10¼	255
T252*	3/3	0-50/0-50	2.0	0.03	19⅞	15¼	10¼	255
T253*	5/5	0-30/0-30	1.0	0.02	19⅞	15¼	10¼	255
T254*	1/1	10-300/10-300	2.0	0.3	19⅞	17	12	325
T255*	2/2	10-150/10-150	2.0	0.2	19⅞	17	12	325
T256*	3/3	0-100/0-100	2.0	0.1	19⅞	17	12	325
T257*	5/5	0-60/0-60	2.0	0 02	19⅞	17	12	325
T258*	10/10	0-30/0-30	1.0	0.01	19⅞	17	12	325

*Normally supplied for 200-250V mains or for 100-125V on request. All other units operate from both mains groups

Single preset voltage, open chassis units

Operating to at least 55°C ambient these units are reliably protected against overloads and short circuits, particularly where the output exceeds 30V. They can be preset to any voltage within their range, with other outputs available by adjustment within the unit. The T180 will operate in 65°C ambient and is therefore suitable to operate in close proximity to vacuum tube equipment. It also has two a.c. outputs, each 6.3V 5A.

Model	DC output (Floating) Amps	Volts	Effective Resistance ohms	Size Inches W	H	D	UK Price
T series open chassis single floating output units							£
T295	0.5	6-30	0.08	2⅞	5¼	4 7/16	39
T296	0.5	6-50	0.1	4¼	7¼	5¾	52
T297	0.5	6-100	0.15	5¼	8¼	6⅞	79
T298	1	6-15	0.04	2¾	6⅞	5¾	47
T299	1	6-30	0.05	4¼	7¼	5¾	49
T300	1	6-50	0.08	5¼	8¼	6⅞	65
T301	1	6-100	0.1	6	9	6	89
T302	3	6-15	0.04	6	9	6	74
T303	3	6-30	0.04	6	9	6	78
T190	3	6-50	0.05	6¼	11	7	85
T194	3	6-100	0.07	7	13	8	128
T304	5	6-15	0.02	6	13	6¼	97
T305	5	6-30	0.02	6	13	6¾	104
T191	5	6-50	0.03	6¼	11	7¼	115
T195	5	6-100	0.04	8¼	13¾	8¼	158
T306	10	6-15	0.01	6¼	17	7	137
T307	10	6-30	0.01	7	17	8	147
T192	10	6-50	0.01	8¼	13¾	8¼	185
T196	10	6-100	0.02	10¾	15¼	11	256
T180	0.5	100-300	0.1	4¾	11¼	6¾	95

Twin preset voltage, open chassis units

Operating to at least 55°C ambient, these units are reliably protected against overloads and short circuits, particularly where the output exceeds 30V. Complete stability is maintained even when mains are ±10% nominal.

Because these units have two independently controlled floating outputs there is no need to specify polarity when ordering. All sections rated above 1A have four terminal networks to eliminate lead resistance.

Model	DC output (Floating) Amps	Volts	Effective Resistance ohms	Size Inches W	D	H	UK Price
T series open chassis twin floating output units							£
T308	0.5/0.5	6-30/6-30	0.08/0.08	6	9	6¼	74
T309	0.5/0.5	6-50/6-50	0.1/0.1	6	9	6¼	99
T310	0.5/0.5	6-100/6-100	0.15/0.15	6¼	11	7	148
T311	1/1	6-15/6-15	0.04/0.04	6	9	6¼	88
T312	1/1	6-30/6-30	0.05/0.05	6	9	6¼	93
T313	1/1	6-50/6-50	0 08/0.08	6¼	11	7	116
T225	1/1	6-100/6-100	0.1/0.1	6¾	14¾	7¼	166
T314	3/3	6-15/6-15	0.04/0.04	6¼	11	7	140
T315	3/3	6-30/6-30	0.04/0.04	6¾	14¼	7¼	146
T221	3/3	6-50/6-50	0.05/0.05	7	13	8	154
T226	3/3	6-100/6-100	0.07/0.07	8¼	15	8¼	234
T316	5/5	6-15/6-15	0.02/0.02	6¾	14¾	7¾	182
T317	5/5	6-30/6-30	0.02/0.02	7	17	8	194
T222	5/5	6-50/6-50	0.03/0.03	8¼	15	8¼	208
T227	5/5	6-100/6-100	0.04/0.04	10¾	17	11	298
T318	10/10	6-15/6-15	0.01/0.01	9¾	17	8¾	252
T319	10/10	6-30/6-30	0.01/0.01	10¾	17	9¼	268
T223	10/10	6-50/6-50	0.01/0.01	10¾	17	11	345

T SERIES POWER SUPPLIES

GERMANIUM

Single preset voltage, open chassis units

These units will operate at full load, at 45°C ambient, even when mains are permanently ±10% nominal. They can be preset to any voltage within their range with other outputs available by adjustment within the unit. All units have safeguards against overload and short circuits, those above 1A have four terminal outputs to eliminate lead resistance.

T series open chassis single floating output units

Model	DC output (Floating) Amps	Volts	Effective Resistance ohms	Size Inches W	D	H	UK Price £
T201	0.5	0.06	0.04	2⅞	5¼	4 7/16	32
T98	0.5	6-30	0.08	2⅞	5¼	4 7/16	32
T202	0.5	6-50	0.1	4⅛	7¼	5¾	40
T149	0.5	0-50	0.1	4⅛	7¼	5¾	45
T203	0.5	6-100	0.15	5¼	8¼	6⅜	62
T197	1	0-6	0.04	2⅞	5¼	4 7/16	35
T165	1	6-15	0.04	2⅞	6⅜	5¾	38
T181	1	6-30	0.05	4⅛	7¼	5¾	40
T100	1	0-30	0.05	4⅛	7¼	5¾	45
T189	1	6-50	0.1	5¼	8¼	6⅜	49
T108	1	0-50	0.1	5¼	8¼	6⅜	55
T193	1	6-100	0.1	6	9	6	76
T169	2	0-15	0.04	4⅛	7¼	5¾	47
T170	2	0-30	0.05	5¼	8¼	6⅜	55
T109	2	0-50	0.05	5¼	10⅜	7¼	70
T198	3	0-6	0.04	4⅛	7¼	5¾	54
T166	3	6-15	0.04	6	9	6	62
T182	3	6-30	0.04	6	9	6	65
T137	3	0-30	0.04	5¼	10⅜	7¼	75
T161	3	0-50	0.05	7	12	6⅞	93
T162	4	0-40	0.04	7	12	6⅞	93
T199	5	0-6	0.02	5¼	8¼	6⅜	69
T167	5	6-15	0.02	6	13	6¾	79
T183	5	6-30	0.02	6	13	6¾	84
T110	5	0-30	0.02	8¼	13¼	7¾	98
T141	6	0-50	0.03	9	15	8¼	135
T163	7.5	0-40	0.02	9	15	8¼	135
T200	10	0-6	0.01	6	13	6¾	98
T168	10	6-15	0 01	6¼	17	7	115
T184	10	6-30	0.01	7	17	8	120
T114	10	0-30	0.01	10	17	8¼	135
T146	12	0-50	0.01	12	17¼	9¼	180
T164	15	0-30	0.01	12	17¼	9¼	180
T142	20	0-30	0.005	16¼	17¼	12	245
T135	25	0-20	0.005	16¼	17¼	12	245
T143	50	0-20	0.002	33	10	22	475
T157	50	0-20	0.002	17¼	20	22	475
T158	100	0-20	0.001	17¼	20	22	665

Twin preset voltage, open chassis units

Because these units have two independently controlled floating outputs there is no need to specify polarity when ordering. Thus, for example, a requirement for −18V 3A and +14V 3A would be met by the unit T116/18/15. All sections rated above 1A have four terminal networks to eliminate lead resistance. All units have safeguards against overload and short circuits.

T series open chassis twin floating output units

Model	DC output (Floating) Amps	Volts	Effective Resistance ohms	Size Inches W	D	H	UK Price £
T204	0.5/0.5	0-6/0-6	0.04/0.04	4⅛	7¼	5¾	62
T209	0.5/0.5	6-15/6-15	0.06/0.06	6	9	6¼	62
T214	0.5/0.5	6-30/6-30	0.08/0.08	6	9	6¼	62
T219	0.5/0.5	6-50/6-50	0.1/0.1	6	9	6¼	76
T224	0.5/0.5	6-100/6-100	0.15/0.15	6¼	11	7	118
T205	1/1	0-6 0-6	0.04/0.04	4⅛	7¼	5¾	65
T210	1/1	6-15/6-15	0.04/0.04	6	9	6¼	69
T115	1/1	0-20/0-20	0.04/0.04	6	10⅜	6	87.10
T185	1/1	6-30/6-30	0.05/0.05	6	9	6¼	72
T220	1/1	6-50/6-50	0.1/0.1	6¼	11	7	88
T151	1/1	0-50/0-50	0.1/0.1	8¼	12¼	7	120
T206	3/3	0-6/0-6	0.04/0.04	6	9	6¼	102
T211	3/3	6-15/6-15	0.04/0.04	6¼	11	7	114
T116	3/3	0-20/0-20	0.03/0.03	8¼	13⅞	7¾	130
T186	3/3	6-30/6-30	0.04/0.04	6¾	14¼	7¼	122
T152	3/3	0-50/0-50	0.05/0.05	10	16¼	9	178
T207	5/5	0.6/0.6	0.02/0.02	6¼	11	7	126
T212	5/5	6-15/6-15	0.02/0.02	6¾	14¼	7¼	144
T117	5/5	0-20/0-20	0.02/0.02	10	17⅜	8¼	176
T187	5/5	6-30/6-30	0.02/0 02	7	17	8	158
T153	5/5	0-50/0-50	0.03/0.03	14	17⅜	9¼	245
T208	10/10	0-6/0-6	0.01/0.01	8¼	17	8½	182
T213	10/10	6-15/6-15	0.01/0.01	9¾	17	8¼	210
T118	10/10	0-20/0-20	0.01/0.01	14	17⅜	9¼	245
T188	10/10	6-30/6-30	0.01/0.01	10¾	17	9¼	218
T144	20/20	0-30/0-30	0.003/0.003	33	10	22	475
T159	20/20	0-30/0-30	0.003/0.003	17¼	20	22	475
T145	25/25	0-20/0-20	0.003/0.003	33	10	22	475
T160	25/25	0-20/0-20	0.003/0.003	17¼	20	22	475

Variable voltage single and twin output instruments

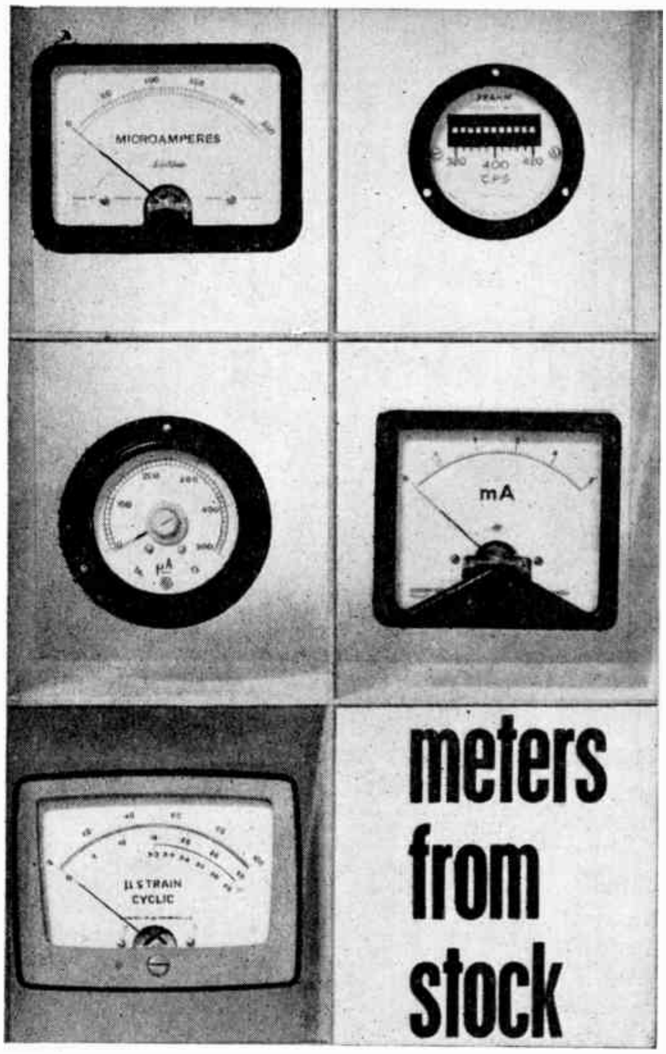
These precision instruments are available in either single or twin outputs, the latter having floating independently controlled outputs. Having 19 inch front panels, these units can also be supplied without the casing for rack mounting. Adequate safeguards against overloads and short circuits are built in.

T series variable voltage single output instruments

Model	DC output (Floating) Amps	Volts	Ripple mV	Effective Resistance ohms	Size Inches W	D	H	19" front panel height	UK Price £
T99	0.5	0-30	1.0	0.1	9¾	5¼	5	—	64.10
T155	0.5	0-50	1.0	0.2	11¼	8¼	6	—	83
T103	1	0-30	1.0	0.05	13¼	10¼	9¾	—	85
T101	1	0-50	1.0	0.1	13¼	10¼	9¾	—	93
T106	1	0-100	2.0	0.1	19¾	14¼	10¼	8¾	149
T102	2	0-30	1.0	0.05	16¾	10¼	10¾	—	100
T105	2	0-50	1.0	0.05	16¾	10¼	10¾	—	118
T107	2	0-100	2.0	0.1	19¾	14¼	10¼	8¾	183
T171	3	0-30	1.0	0 03	16¾	10¼	10¾	—	118
T111	5	0-50	2.0	0.01	19¾	18¼	10¼	8¾	155
T172	10	0-30	2.0	0 01	19¾	18¼	10¼	8¾	168
T112	10	0-50	2.0	0.01	19¾	20¼	12	10½	245
T173	15	0-30	2.0	0.01	19¾	20¼	12	—	265
T113	20	0-30	2 0	0.01	19¾	21	15¼	14	315
T320	30	0-30	3.0	0.005	19¾	21	19	17½	385
T321	50	0-30	5.0	0.002	21¼	25¼	38¼	31½	595
T322	100	0-30	6.0	0.001	21¼	25¼	38¼	31½	825

T series variable voltage twin output instruments

Model	DC output (Floating) Amps	Volts	Ripple mV	Effective Resistance ohms	Size Inches W	D	H	19" front panel height	UK Price £
T104	1/1	0-30/0-30	1.0	0.05	16¾	10¼	10¾	—	129
T134	1/1	0-50/0-50	1.0	0.1	19¾	14¼	10¼	8¾	166
T174	2/2	0-30/0-30	1.0	0.05	16¾	10¼	10¾	—	143
T175	3/3	0-30/0-30	1.0	0.04	19¾	14¼	10¼	8¾	166
T136	3/3	0-50/0-50	1.0	0.05	19¾	18¼	10¼	8¾	220
T176	5/5	0-30/0-30	1.0	0.01	19¾	18¼	10¼	8¾	220
T139	5/5	0-50/0-50	1.0	0.01	19¾	20¼	12	10½	275
T177	10/10	0-30/0-30	2.0	0.01	19¾	20¼	12	10½	275



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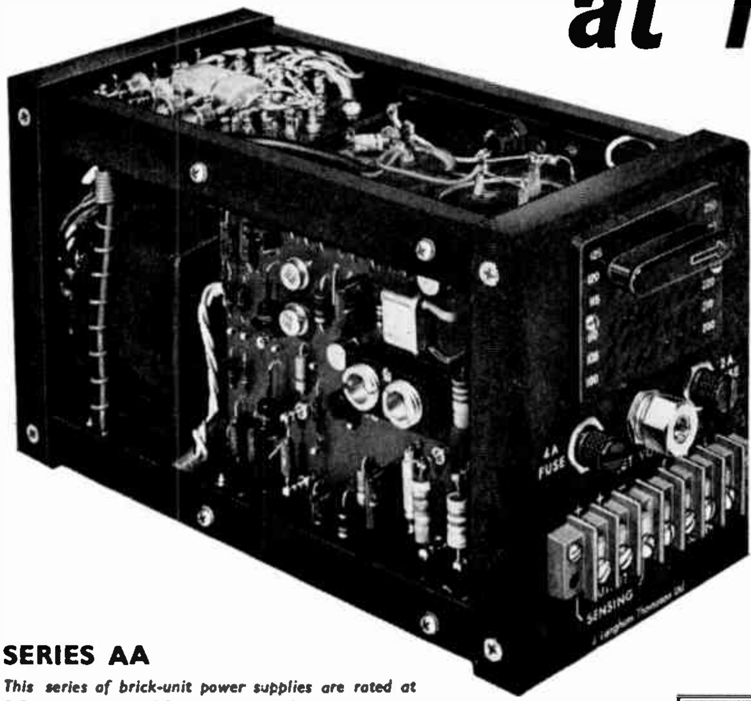

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