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Commentary

'GENERAL Requirements for Electronic Parts of Assessed Quality' is the title of a draft British Standard which has recently been circulated for comment. If widely adopted, as all users and most manufacturers must hope it will be, this Standard could do much to clarify the somewhat muddled position which at present obtains in the electronic component industry and could also have a pronounced effect on the reliability of components in general.

This draft Standard follows on from the recommendations of the Burghard Report which was issued in May 1965. The body which formulated this Report was the Committee on Common Standards for Electronic Parts which was composed of representatives of British manufacturers and users of active and passive electronic parts; it met under the auspices of the Ministry of Aviation and had as its Chairman Rear Admiral G. F. Burghard, C.B., D.S.O.

After consideration of the Burghard Report by all the Government and industrial organizations concerned, the BSI was formally requested, early in April of this year, to implement the scheme. However, in anticipation of the official signal to go ahead, nine months preparatory work had been put in by an unofficial working party representing the interests chiefly concerned and the result is the Draft Standard already mentioned. Apart from the usefulness of the work itself, the BSI and the other parties involved are to be heartily congratulated on the speed with which this complicated task has been carried through.

The widely differing environmental and circuit conditions in which electronic parts are used coupled with the lack, up to now, of a comprehensive system of specifications, has resulted in a multiplicity of specifications to which parts are purchased. These range from complex Government specifications to manufacturers' data sheets and it is frequently difficult, if not impossible, to translate adequately from one to the other. The object, therefore, is to replace existing Ministry of Defence, Ministry of Aviation, British Standard and other specifications by a single common standards system.

An important point in the philosophy of the Burghard Report and the ensuing Standard is the recognition of the fact that in reality very few parts are made solely for military or other government purposes and parts bought to different specifications have frequently come off the same production line and differ only in the amount of testing to which they were subjected, or in their finish. The new system is so arranged that it will cover the bulk of military and industrial requirements so that the maximum economic benefits can be derived from more extensive mass production and testing; also reliability information may be obtained from cumulative test data based on this larger scale production.

The system is designed to cover three groups of component usage as follows:

General application, where the electrical, mechanical, environmental and endurance performance is specified and assessed to meet the bulk of requirements, without discrimination of the Services and the capital goods industry;

Special application, where a higher degree of performance at the environmental stress levels of the general application group and/or proven performance at other stress levels will be specified and assessed;

Commercial application, where a reduced level of testing at the environmental stress levels of the general application group, or the same performance at lower levels will be specified and assessed.

The system will also enable the sole manufacturer of a particular part to write and publish a specification which conforms with the requirements specified in the British Standard for the family of parts to which the particular component belongs. This will enable special or improved parts to be introduced with a minimum of delay.

With regard to Qualification Approval it has been agreed that the Electrical Inspection Directorate of the Ministry of Aviation* will act as an executive agent of BSI for the approval of manufacturers and test houses and for the general supervision of inspection and test. An approved manufacturer wishing to supply a particular electronic part to common standards must then carry out a test procedure under the supervision of BSI's executive agent to establish Qualification Approval. Any electronic parts for which Qualification Approval has been granted will be added to the Qualified Electronics Parts List which will be issued by BSI.

The Burghard Report emphasizes the need for international co-operation and specifically recommends that the technical content of these new standards should conform with IEC Recommendations, and where these do not exist with basic NATO specifications. If neither IEC nor NATO specifications exist, then the standards should not differ unnecessarily from American MIL specifications. In fact, the Burghard Report has already aroused significant interest in Europe and it is evident that some European countries are thinking along the same lines as Britain regarding the need for a set of standards which give some assurance of reliability†.

The Burghard Committee and the BSI have started on a very worthwhile task in a decidedly workmanlike and efficient manner but to carry the plan to fruition will require the wholehearted support of the entire industry; if carried out successfully, however, it should have a very beneficial effect on economy, efficiency and the more easy and realistic assessment of reliability figures for complete equipments.

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* *At the time of going to press it has been announced that the Ministry of Aviation is to be disbanded and it is presumed that one of the other Ministries will take over this task.*

† *See page 482 of this issue.*

A Brushless D.C. Stepper Motor with Binary-to-Decimal Conversion

By W. Leonchick*, B.S.E.E.

A machine language translator utilizing a unique vectorial principle to obtain direct decimal read-out from binary data input is described. To ensure clarity, the development is purposely detailed. Since stepper motors of the non-contact variety are essentially digital input/output devices, the second half of the article deals with the utilization of the above translator as an open loop d.c. step-servo mechanism. Advantages are: (a) reliability, (b) maintenance-free operation, and (c) less space and weight as compared to equivalent a.c. servo circuits. The system described is not affected by varying environmental conditions.

Digital-to-analogue conversions as well as binary-to-decimal translating equipment is eliminated by the versatility of the electromechanical design.

(Voir page 488 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 495)

WITH the coming of the digital revolution, there is a great need for interface equipment in the field of input/output translation devices. D.C. stepper motors of the brushless type are making a comeback from their long hibernation. One of the reasons is due to the fact that d.c. motors utilizing brushes have a tendency to disintegrate in rarified atmospheres.

This article will describe in a step-by-step manner a simple electromechanical device that combines a unique machine language translator (in the binary-to-decimal direction) along with the driving power of a d.c. brushless motors of the non-contact type offering the following advanages:

(1) Non-ambiguity
(2) Large stepping angles
(3) Positive magnetic detent
(4) High starting torque
(5) Relatively fast response
(6) Rarified atmosphere operation
(7) Small size and weight
(8) Reliability and long life
(9) Economy
(10) Efficiency

In addition to its novel and principal advantage of translation, this conversion device can be used as a step-servo motor with the inherent advantages of the latter. Built-in versatility will permit it to be operated directly from a binary input or from the decimal system with the addition of a diode matrix. Some stepper motors require separate lines for clockwise (c.w.) or counterclockwise (c.c.w.) operation; other types offer c.w. or c.w.w. rotation depending on the polarity of the pulse input. The concept to be described illustrates one simple method of bi-directional output employing a common line and one polarity of signal input.

The basic theory will be presented with the hope of stimulating interest in the field of ingenious electromechanical translators that are suited for myriad tasks in the field of digital computers and other automatic devices. The design approach makes the unit adaptable to both industrial and military designs.

Binary-to-Decimal Translator

The decimal numbering system was undoubtedly developed from the fact that humans have ten fingers. In other words, ten is the radix or base of the decimal system. However, the latter is too complicated for use as machine language—or logic—as used in computers, because most mechanisms and electromechanical devices are binary in character, having inherently only two states or conditions such as on/off, start/stop, go/no-go, pulse/no-pulse, etc.

A compromise is reached in that the majority of digital computers operate internally in the binary number system with a radix or base two employing only two (2) symbols such as one and zero. Usually data to and from the computer is represented in the familiar decimal system.

Thus the need for highly specialized translators such as encoders and its more complex complement, the decoder. This section will describe a unique electromechanical read-out device with an integral decoder (or machine language translator) that converts binary coded data directly to the decimal system by means of a vectorial principle. In addition, the conversion device occupies a minimum of physical space.

As is well known, an encoder is just a network of logic OR circuits in which only one input is excited at a time to produce a multiplicity of outputs. A diode matrix makes an ideal decimal-to-binary encoder when wired as shown in Fig. 1.

Compared to the OR circuit of the encoder, a decoder is simply a network of logic AND circuits used to convert a combination of pulses or signals into one output of a group of signals. There are various techniques for the translation of encoded binary data signals into the decimal system. One of the simplest to understand is the relay tree wired as shown in Fig. 2.

It is well known that every number, expressed as a whole number or as a whole number plus a decimal fraction, may be represented in the binary system by a pair of symbols expressed in a definite array. Conversely, a binary coded number may be directly converted into a decimal number. This may best be illustrated with the tabulation in Table 1 of decimal/binary equivalents by representing each digit in the decimal number by a 4-bit binary number.

From this it follows that with 3-bits (binary digits) of information in the binary system, exactly eight (8) combinations or numbers can be represented in the decimal system as: $2^3 = 8$. Thus, when using a decimal-coded binary system, at least 4-bits of information must be

* *Argonne National Laboratory, U.S.A.*

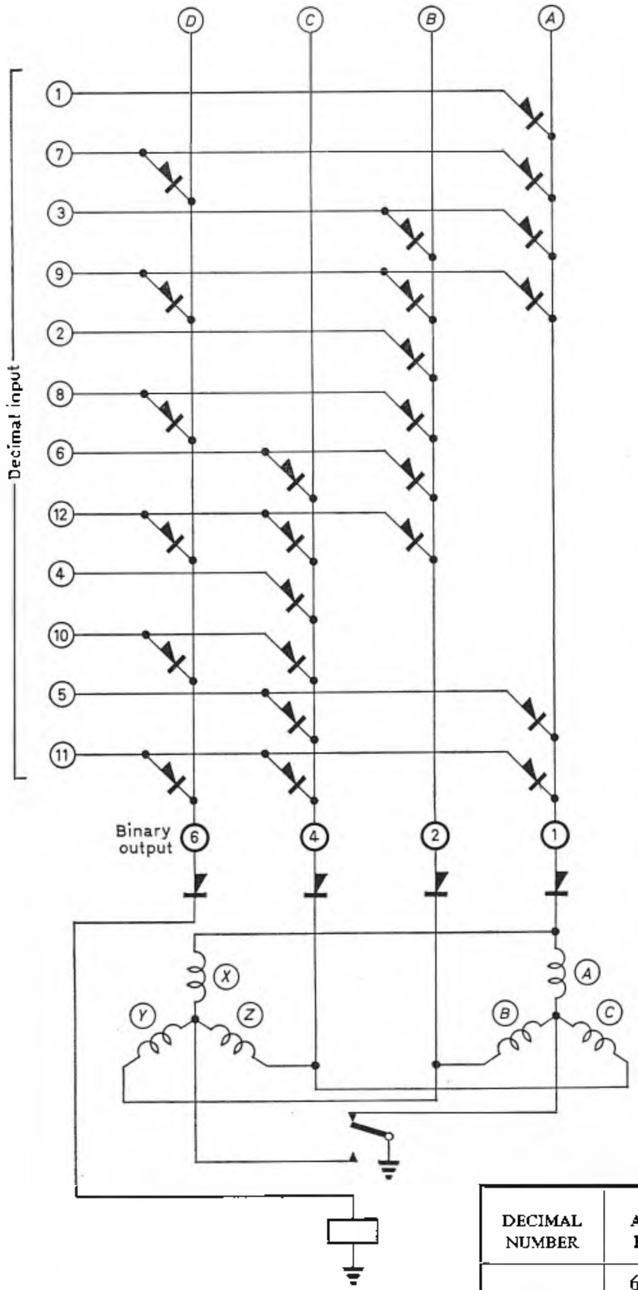


Fig. 1. Decimal-to-binary encoder

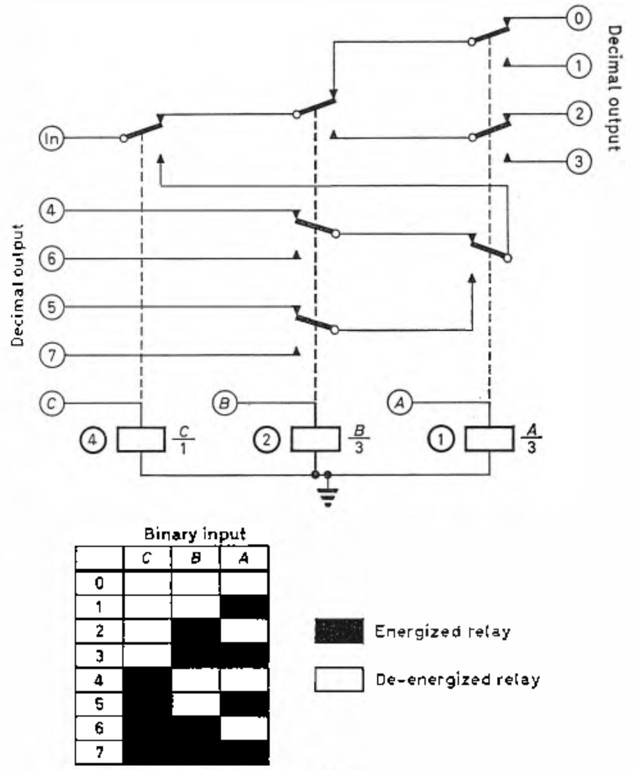


Fig. 2. Binary-to-decimal decoder

available for a maximum of 2^4 or 16. Of the latter configurations, usually only ten (10) combinations are required for a decimal system. Twelve (12) numerals will be used for the present example for obvious reasons as will be described in the text.

A simplified circuit is shown in Fig. 3 to illustrate the basic principle. Referring to Fig. 3(a), an internal equally spaced six-tooth lamination is utilized having a wye-winding of three-coils. Each coil is mounted on alternate teeth, and the latter are designated as coils *ABC* in a clockwise direction. Inside the stator is located a two-pole permanent magnet rotor (materials such as Alnico or ferrites can be used) with a suitable air-gap between the stator and the rotor. Thus the only moving part, the rotor, is free to rotate in either direction on a ball bearing or other similar device.

Now by applying a positive d.c. voltage to coil *A*, electromagnetic flux of the proper polarity would be set up inside this coil. The magnet will swing into position (by means of the shortest route—either clockwise or counterclockwise) and align itself in the manner shown in Fig 3(a). Energizing coils *B* and *C*, in turn, the north

TABLE 1

DECIMAL NUMBER	ASSIGNED 4-BIT BINARY VALUE				CONVERSION TO DECIMAL VALUE		
	6	4	2	1	BINARY	DECIMAL	SUM
1	0	0	0	1	0 + 0 + 0 + 1	0 + 0 + 0 + 1	= 1
2	0	0	1	0	0 + 0 + 1 + 0	0 + 0 + 2 + 0	= 2
3	0	0	1	1	0 + 0 + 1 + 1	0 + 0 + 2 + 1	= 3
4	0	1	0	0	0 + 1 + 0 + 0	0 + 4 + 0 + 0	= 4
5	0	1	0	1	0 + 1 + 0 + 1	0 + 4 + 0 + 1	= 5
6	0	1	1	0	0 + 1 + 1 + 0	0 + 4 + 2 + 0	= 6
7	1	0	0	1	1 + 0 + 0 + 1	6 + 0 + 0 + 1	= 7
8	1	0	1	0	1 + 0 + 1 + 0	6 + 0 + 2 + 0	= 8
9	1	0	1	1	1 + 0 + 1 + 1	6 + 0 + 2 + 1	= 9
10/0	1	1	0	0	1 + 1 + 0 + 0	6 + 4 + 0 + 0	= 10/0
11	1	1	0	1	1 + 1 + 0 + 1	6 + 4 + 0 + 1	= 11
12	1	1	1	0	1 + 1 + 1 + 0	6 + 4 + 2 + 0	= 12

pole of the rotor will swing to coil B, tooth 3, when it is energized and by the same process to coil C, tooth 5. Thus three distinct positions are achieved.

From the above description, it would be an advantage to shorten the flux path with a bifurcated south pole. That is, a tri-pole rotor can minimize the reluctance of the magnetic circuit; for example, with the N-pole at stator position 1 as in Fig. 3(a), the two S-poles would be located at stator positions 3 and 5. A view of the tri-pole magnet to be discussed later is shown in Fig. 6(a).

Another advantage of a three-pole magnet is to prevent 'freezing' of the magnetic rotor when a 180° reversal occurs. A two-pole rotor is more likely to create two symmetrical magnetic fields in the stator legs. With a tri-pole rotor, the probability of having the two S-poles of equal flux density are very slim.

By applying equal d.c. voltages simultaneously to coils A and B of Fig. 3(a), the magnetic flux of the rotor will align itself with the stator electromagnetic flux and come to rest midway between coils A and B at stator position 2. Completing this same sequence in a clockwise direction, three more magnet positions are obtained for a total of six distinct positions—one for each tooth. This principle is further explained by vectorial addition as shown in the vector diagram of Fig. 3(c). Numerical values can be assigned to the three coils as follows:

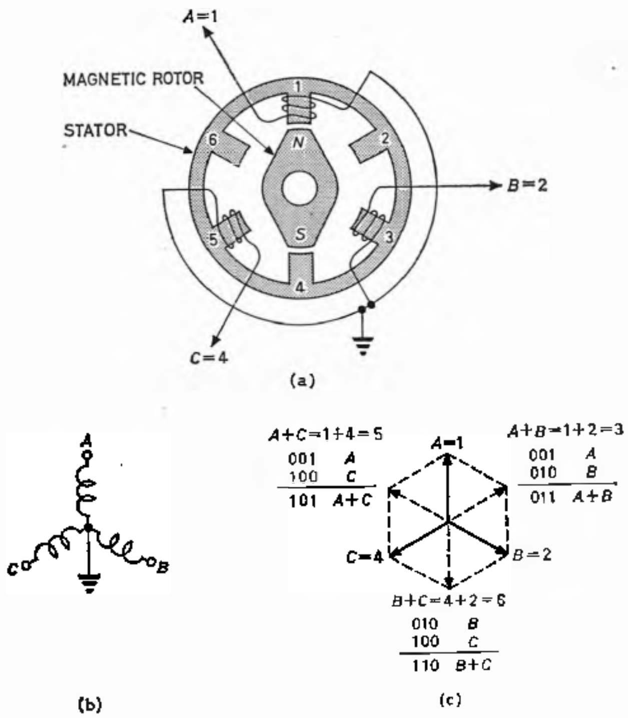
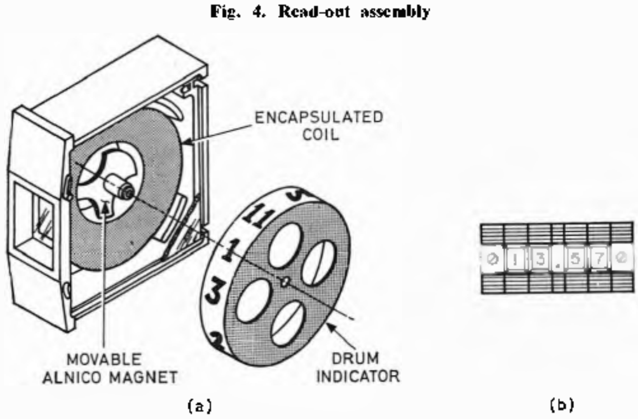


Fig. 3. Basic 6-pole decoder

Fig. 4. Read-out assembly



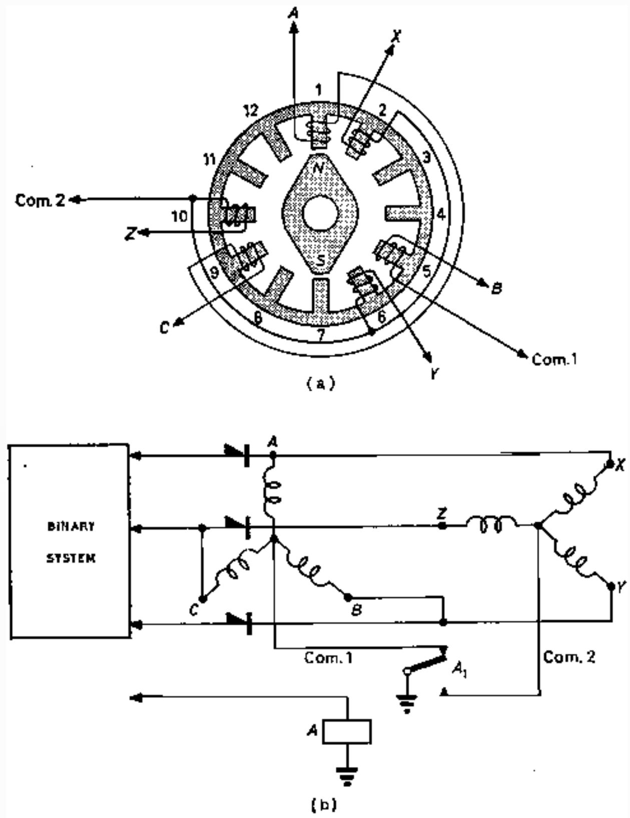


Fig. 5. Principle of 12-pole decoder

COIL	WEIGHTED VALUE
A	$2^0 = 1$
B	$2^1 = 2$
C	$2^2 = 4$

Attaching an external drum securely to the rotor assembly as shown in the exploded view of Fig. 4(a), the small stationary window can be used as the reference read-out position. Thus when coil A is energized, numeral 1 on the drum will be visible in the window. When coils B and C are energized in turn, numerals 2 and 4 would appear for read-out. Now when both coils A and C are energized simultaneously, number 5 will appear since $A + C = 1 + 4 = 5$. Using the same logic: $B + C = 2 + 4 = 6$ and $A + B = 1 + 2 = 3$. In other words, the numerical sequence on the drum due to the vectorial addition would follow the order: 1, 3, 2, 6, 4, 5 and back to the starting position of 1.

It is obvious that the procedure just performed was a binary-to-decimal translation yielding a decimal read-out without the use of complicated devices such as commutators, relay-trees, diode switching matrices, etc. The decoding and display mechanism would be satisfactory if there were only six numerals involved in the decimal system. At least four (4) more numerals or positions on the drum will be required for the decimal system—and preferably a couple of spare positions for miscellaneous duties, say, a total of twelve positions.

On first thought, since a wye-winding was used to obtain six positions, why not a double wye-winding with a twelve tooth stator to obtain the desired twelve decimal

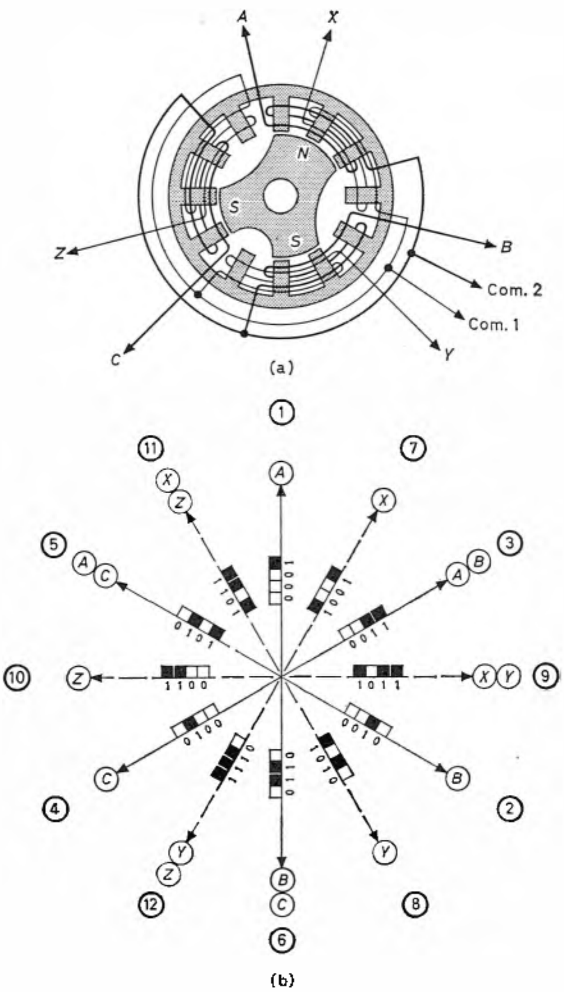


Fig. 6. Prototype of d.c. step-servo/decoder

readout positions? An assembly diagram showing a dual wye-winding assembled on an internal equally spaced twelve tooth lamination is shown in Fig. 5(a); the schematic diagram is shown in Fig. 5(b). The latter also illustrates one method of switching from one wye-winding, say, ABC, to the other wye-winding, XYZ, by means of a simple single pole double throw subminiature relay whose weighted value would be six as designated in Table 1. Of course, the same switching function could be performed by means of semiconductor devices such as a flip-flop which can be described as an electronic version of a s.p.d.t. switch. When this switch is energized, the additional six numerals, 7 to 12, inclusive, are available for read-out in the window.

For the sake of simplicity, the discussion up to this point was mainly on the basis of a two-pole magnetic rotor and having the coil windings embrace single tooth laminations. However, for greater torque and reliability, the practical approach for a production model is shown in Fig. 6(a). Basically this assembly is equivalent to that of Fig. 5(a) except that the coil windings span three teeth for distribution of magnetic flux and thus minimize saturation effects of the stator laminations. The N-pole of the rotor has also been extended to the width of three teeth, while the S-pole has been bifurcated into two sections—both spanning two teeth as shown.

This procedure has the following advantages:

(1) Minimum reluctance
(2) Greater torque
(3) Large stepping angles
(4) Non-ambiguity
(5) Positive magnetic detent
(6) Elimination of stall actions

Open-Loop Stepping Motor Operation

The stepper motor is a natural device for open-loop control systems because of its digital behaviour. With a d.c. programmed input, a step-servo motor will index in given angular increments. Angular displacement in a clockwise or counterclockwise manner is determined by the input to the windings ABC/XYZ as previously described. Basically, there are two types of stepper motors:

(1) PM stepper motor (permanent magnet) which operates on the reaction between an electromagnetic field and a permanent magnet.

(2) VR stepper motor (variable reluctance) which

TABLE 2

30° STEP-SERVO STANDARD SPEED		60° STEP-SERVO DOUBLE SPEED		90° STEP-SERVO TRIPLE SPEED		120° STEP-SERVO QUADRUPLE SPEED	
C.W.	C.C.W.	C.W.	C.C.W.	C.W.	C.C.W.	C.W.	C.C.W.
0001 1	0001 1	0001 1	0001 1	0001 1	0001 1	0001 1	0001 1
1001 7	1101 11	---	---	---	---	---	---
0011 3	0101 5	0011 3	0101 5	---	---	---	---
1011 9	1100 10	---	---	1011 9	1100 10	---	---
0010 2	0100 4	0010 2	0100 4	---	---	0010 2	0100 4
1010 8	1110 12	---	---	---	---	---	---
0110 6	0110 6	0110 6	0110 6	0110 6	0110 6	---	---
1110 12	1010 8	---	---	---	---	---	---
0100 4	0010 2	0100 4	0010 2	---	---	0100 4	0010 2
1100 10	1011 9	---	---	1100 10	1011 9	---	---
0101 5	0011 3	0101 5	0011 3	---	---	---	---
1101 11	1001 7	---	---	---	---	---	---
0001 1	0001 1	0001 1	0001 1	0001 1	0001 1	0001 1	0001 1

NOTE: *Decimal numerals at right of above columns indicate PM rotor position and decimal equivalent as illustrated in Fig. 6(b).*

operates on the reaction between an electromagnetic and a soft iron rotor.

From the aforementioned definitions, the assembly described in this article would fall in the general category of the PM type.

The designed stepping angle of a stepper motor should not be greater than $2\pi/3$ radians or 120°. The assembly shown in Fig. 3 would have six steps for a 360° revolution or 60° per step. Such a unit may be classified as a 60° step-servo. The assemblies shown in Figs. 5 and 6 can then be designated as 30° step-servos (360°/12-steps).

The maximum pulse rate of 300p/s bidirectionally of a PM step-servo is relatively low when compared to a maximum pulse rate of 1 200p/s for a VR step-servo. Upon the application of pertinent logic to the stepper motor, bidirectional rotation can be obtained. In addition, the stepping rate and thus the speed can be doubled, tripled and quadrupled as shown in Table 2.

Conclusion

Logic techniques are universally employed to advantage in almost every part of modern control systems. A new breed of a d.c. brushless motor is described in two parts:

(1) MACHINE TRANSLATOR (Binary-to-Decimal Decoder) An electromechanical device that converts binary coded decimal data directly into the decimal system by means of a unique vectorial principle.

(2) STEPPER MOTOR (DC Step-Servo) Clockwise or counterclockwise angular incremental displacements are obtained when energized by d.c. voltages in a programmed manner. Stepping angles are easily varied to achieve multiple speeds without the need of mechanical gearing.

The design described in this article permits operation in hazardous locations such as radiation and toxic environments as well as in the rarified atmospheres of space. The device will operate under high shock and vibration conditions and also when completely immersed in liquid media.

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Microelectronic Technology Points the Way Ahead

By E. E. Webster*

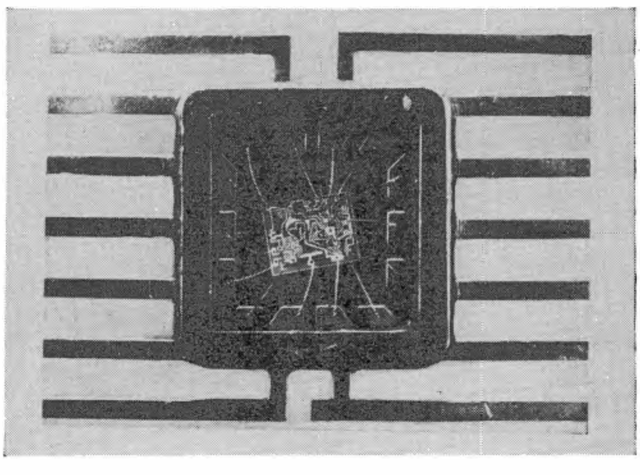
THE level and scope of the technologies which are now being applied to the design, development and manufacture of monolithic integrated circuits represent a major revolution within the components industry. This concentration of technical effort demands a parallel revolution in business policy, on a scale never previously required by any single technological advance. Significant changes in investment are necessary for any company aiming to produce integrated circuits successfully, which means in sufficient quantity, of the right types to suit modern equipment and at the appropriate times.

The sheer volume of investment initially required from a company—either to 'go it alone' or to operate in partnership—is of the order of £½M a £1M a year. This is the sort of cost involved in providing buildings, plant, personnel and staff training. But this is not all; creating new markets, or establishing new applications for integrated circuits in place of assemblies wired from discrete components, will require a fantastic commercial effort unparalleled in the history of the components industry.

Recruiting sufficient competent personnel is a problem of comparable magnitude. Integrated circuit techniques demand a new breed of engineer to handle design, development, manufacture and application. By their very nature, integrated circuits blur the traditional dividing line between components designers and equipment designers so that, at the very least, each must become much more closely involved in the other's function, while in some cases the distinction may disappear altogether. Integrated circuits demand integrated design thinking, such as only a small proportion of present day electronics men have developed. Recruitment and retraining of staff will be a gigantic task fully comparable in scale with the technical, financial and commercial problems posed.

Nevertheless, these problems will be solved, because the economics of equipment manufacture ultimately demand

A monolithic silicon chip mounted in a 14 lead flat-pack as used in the Plessey Automation XL11 microminiature computer. This digital logic circuit, which takes only 5nsec per logic operation, was fabricated at the Allen Clark Research Centre of The Plessey Co. Ltd. The flat pack is ¼in square and the leads are taken out with a 0·050in pitch



* *Director, Plessey Components Group.*

it. Integrated circuits will enable assemblies—which are today available only expensively in small numbers—to be used repetitively, and cheaply, in very large quantity.

Investment on the scale envisaged emphasizes both the heavy capital risk carried by the giants of the U.S. industry and also their competitive ability. It also suggests that no more than an estimated maximum of five companies in Great Britain can engage fully in the design, development and manufacturing technologies needed; indeed, by 1970, there may be only three British companies which can accept the level of investment and utilization —which will certainly be required of them—to stay in this demanding field.

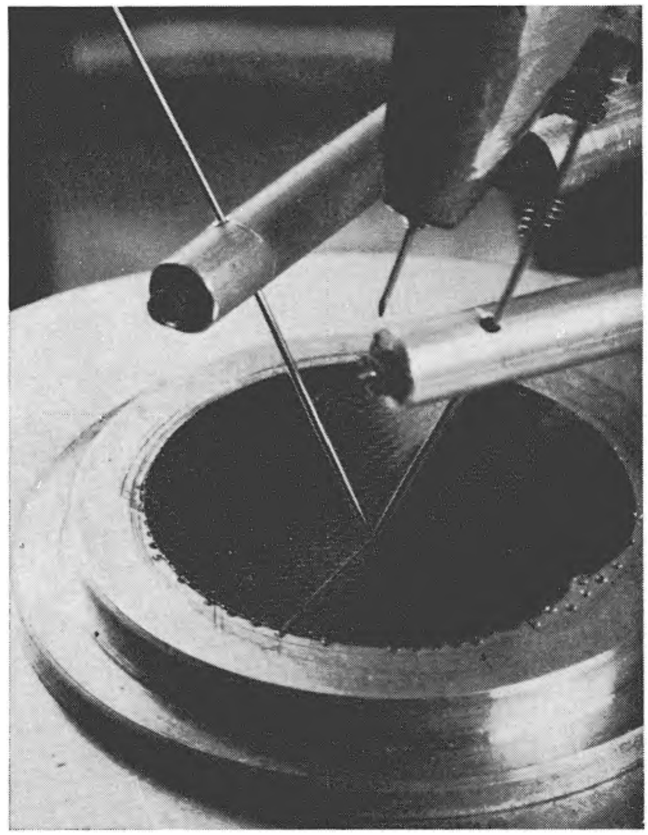
Considerable work has been devoted to the development of thin film and thick film passive networks, which can be made into complete circuits by the addition of active elements. It is considered that this is an interim solution to the problem of micro-miniaturization as film circuits, like all hybrids, must eventually give way to fully integrated circuits for the great majority of applications. Evaporated thin film circuits will almost certainly retain a permanent place in electronics as passive networks for special precision applications, while thick film circuits and elements (due to their notable stability) will survive in special applications demanding ruggedness and the ability to withstand adverse environments. But the elegance, classic simplicity and economy of fully integrated circuits must meet a major and increasing part of the industry's requirements.

Britain cannot afford to be left behind in the race for leadership in an advanced technology which is forced on her by the accelerating use of integrated circuits. However, she can—if desired—compete successfully in all phases of this development on the same technical level as the U.S.A. and U.S.S.R., notwithstanding the fact that Britain has little or no future in space. Many applications of integrated circuits are associated with equipment which has no more exotic housing than an office, a telephone exchange or a ship's instrumentation system. This country has lost out on so many developments in recent years through the absence of design fall-out from defence programmes, that it is encouraging to recognize one new technology with a vast potential home market and which does not require Government Defence spending to sustain it.

Effort and expenditure on semiconductor technologies will inevitably benefit British industry on a very wide front. The design, development and production of integrated circuits extend the frontiers of knowledge in many associated areas; optics, vacuum technology, crystallography, micro-welding and many other manufacturing techniques are all stimulated by the often unprecedented demands of this relatively new field. On this score alone, the effort applied to integrated circuits is very worthwhile.

The short and long term futures of many major electronic and electrical equipments is inevitably tied up with the extended use of integrated circuits. Computers, telephones, radio transmitters, radio receivers and process control systems are notable examples, since each contains circuits which are repetitive in nature and required in large quantities. The economics of integrated circuit manufacture are such that large numbers of identical units must be produced to sustain the enterprise and recover the investment.

Not less than 10 000 units per week of any one particular circuit is, at the present time, the order of production rate needed to ensure survival. Furthermore, to maintain commercial viability and cover the total investment, several different types of circuit must be produced in parallel. Prices initially will be high, but intense competition—and equally intense refinement of production processes—is likely to reduce cost rapidly, from the current level of a few pounds per circuit to a few shillings, by 1970.



Electrical probes testing a slice containing planar epitaxial transistors at the Allen Clark Research Centre of The Plessey Company

Integrated circuits will determine the future for a large range of applications. They will not supersede discrete components for many purposes but they will certainly slow down the growth of demand for discrete components between 1966 and 1970. There are many applications where discrete components are most satisfactory, such as for achieving special characteristics, because of the quantity involved, or to counter problems of vulnerability. There is no doubt where these are required they will be used. In this sense, therefore, integrated circuits will produce an expansion in the total electronics market rather than displace discrete components completely.

To be competitive in performance, assemblies of separate components will have to achieve a reliability similar to that which is inherent in integrated circuits and it is considered that this requirement must force components designers to re-think many aspects of current design. Their aim will have to be to attain a reliability for n components, when assembled together as one circuit, which is equal to or better than that of a corresponding integrated circuit.

The investment in integrated circuits is formidable, but prospects are exciting and the reward is great. As a nation, we must not hesitate but capture for ourselves a leading position in this work, which is essential to this country's economic well-being in the future.

A M.O.S.T. Gate for Pre-Sense Amplifier Strobing

By G. N. Hunter*

The suitability of the m.o.s.t. for low level gating is illustrated. When strobing the input of a sense amplifier, in a computer store, the problem is to reject large noise pulses a short time before allowing low-level signal transmission. A practical gating circuit is examined and its performance, with current samples of m.o.s.t's, shown to be compatible with store cycle times down to 1μsec.

(Voir page 488 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 495)

THE signal output from a ferrite core storage plane requires amplification before reliable discrimination between a '0' and a '1' can take place (the '0' is represented by the absence of a pulse). This amplification is seriously hindered by the presence of large noise pulses appearing at the plane output a very short time before the relatively small signal pulse occurs. (This noise may be ~5V while the signal is ~50mV.) The sense amplifier will be overloaded by the noise, and considerable care must be taken to ensure that it recovers with sufficient speed to amplify the following low level signal.

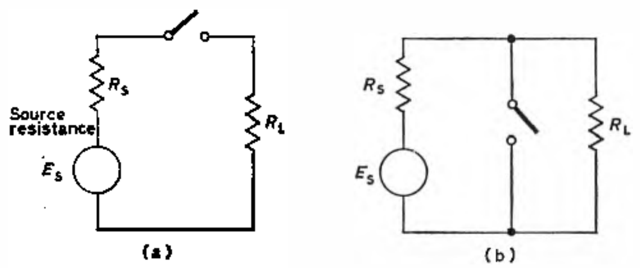
However, the amplifier design may be made much simpler if the amplifier input is gated in such a way that the noise pulses are greatly attenuated but the signal is transmitted with low attenuation.

Possible Gating Circuits

The two simplest well known gating circuits are shown in Fig. 1(a) and (b). Both these arrangements show practical limitations. In the series circuit the 'short-circuit' resistance R_{sc} of the switch provides signal attenuation, while the 'open-circuit' resistance R_{oc} allows noise transmission. The external load resistance R_L must be high compared with R_{sc} and R_s to give small signal-attenuation, but must be low compared with R_s and R_{oc} to give high noise attention. The shunt circuit will introduce less signal attenuation but will also give less attenuation to the noise (depending on how small R_{sc} is compared with R_s). A further point of interest arises if there is parasitic capacitance between the load and the switching source; it is important that the switching transient thus introduced should decay as quickly as possible. The worst case will occur when the switches are being driven into the open-circuit condition. The transients then have to decay through the load resistance which, as seen from consideration of the signal and noise attenuation cannot be made low. Thus, for a high speed gate, both of the simple circuits lack the flexibility required to optimize both the d.c. and the high speed characteristics.

The operation of the series-shunt circuit Fig. 1(c) however, eliminates the disadvantages of the simpler arrangements. The switches A and B are operated in push-pull so that when transmitting the signal A is closed and B open and when attenuating the noise A is open and B closed.

The signal attenuation is determined by the ratio of R_L to $R_s + R_{A(sc)}$, and for this purpose R_L can be made large. The noise attenuation is determined approximately by the ratio of $R_{B(sc)}$ to $R_{A(oc)}$. The high speed operation is now considerably improved, for whenever one switch is open-circuit the other switch will be short-circuit providing a low impedance path to earth. This means that any switching transients introduced across the load resistance will be able to decay rapidly. Furthermore the push-pull operation means that these switching transients will be of opposite polarity and will thus tend to cancel each other out.



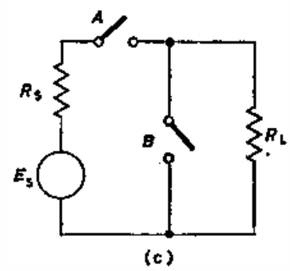


Fig. 1. Basic switching circuits
(*a*) *Gate circuit showing series switch*
(*b*) *Gate circuit showing shunt switches*
(*c*) *Gate circuit showing series and shunt switches*

The transistor switch exhibits leakage currents and voltage drops across the pn junctions, both of which vary with temperature. In this application these effects set a limit to switch performance. If the transistor is either a series or shunt switch at the amplifier input then the leakage currents set a limit to the circuit noise rejection in the series circuit and produce an offset voltage in the shunt circuit, while the junction voltage drop allows noise transmission in the shunt circuit and an offset voltage in the series circuit. The principle problem here is the offset voltage because this appears at the same time as the signal and thus may reduce the signal-to-noise ratio intolerably (i.e. it may reduce the '1' output or increase the '0' output).

This approach is made much more attractive if the switching element is a metal oxide semiconductor transistor (m.o.s.t.). With this device the insulated gate isolates the gating waveform from the signal path, such that the leakage current in the source-drain circuit, caused by the applied drive voltage on the gate, is extremely small (pico amperes). The offset voltage caused by leakage currents flowing through the load is thus reduced to negligible proportions. For this reason the m.o.s.t. is admirably suited to low level gating circuits.

The following sections describe the m.o.s.t. characteristic

* *Mullard Research Laboratories.*

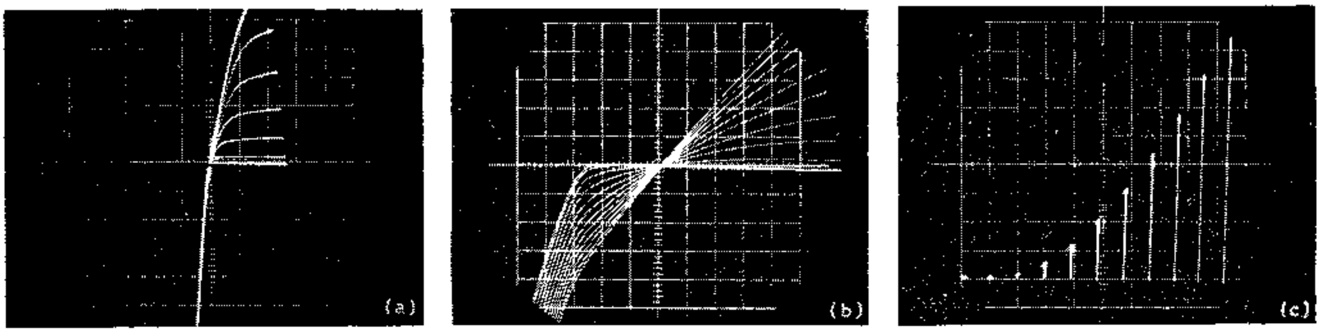


Fig. 2. N-channel m.o.s.t. characteristics

(a) I_{ds} versus V_{ds} with gate voltage as parameter
Horiz : 5V/cm
Vert : 5mA/cm
Gate volts : 2V/step

(b) Linear region of Fig. 2(a)
Horiz V_{ds} : 200mV/cm
Vert I_{ds} : 1mA/cm
Gate volts : 1V/step

(c) I_{ds} versus V_{gs} characteristic
Horiz V_{gs} : 1V/cm
Vert I_{ds} : 1mA/cm
Showing pinch off voltage

in further detail and examine the operation of a circuit suitable for pre-sense amplifier strobing in a computer store.

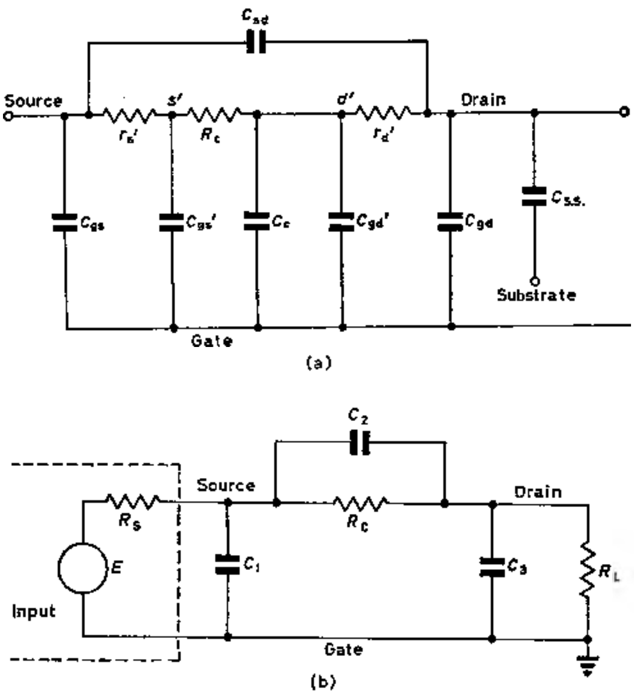
M.O.S.T. Characteristics

A very brief description of the characteristics which make the m.o.s.t. particularly suitable for low level gating circuits will now be given.

The bi-directional nature of the device is illustrated in Fig. 2. Positive and negative signals will be transmitted with equal attenuation, so long as they are limited to the linear range shown in Fig. 2(b), i.e. V_{ds} numerically less than 0·5V. It should also be noted that the linearity improves with increasing gate voltage. This is illustrated in Fig. 2(c) by the curve of I_{ds} versus V_{gs} which becomes more nearly linear as $V_g - V_p$ increases (V_p is the pinch-off voltage).

The slopes of the curves in Fig. 2(b) give the resistance of the paths between source and drain when driven 'on' by a certain gate voltage. This 'on' resistance (R_{sc}) is an important parameter as it determines the attenuation of the gating circuit for signal transmission. The minimum resistance that can be obtained with the particular device shown in Fig. 2(b) is approximately 200Ω.

Fig. 3. The series m.o.s.t. in the 'on' state
(V_{sd} zero and $V_g > V_p$)
(a) Equivalent circuit
(b) Simplified equivalent circuit



The very high gate input impedance (typically $10^{12}\Omega$) means that there will be no current (of practical significance) injected into or drawn from the source-drain circuit by the gating waveform. The resistance between source and drain can be made extremely high if the gate voltage is biased more negatively than the pinch-off voltage (for *N*-channel types). This 'off' resistance can be made greater than 100kΩ with ease.

An equivalent circuit for the series m.o.s.t., when the gate is driven on, is shown in Fig. 3(a) and illustrates how integration can occur. r_s' and r_d' represent the resistance from the channel to the source and drain terminals respectively, (including the contact resistance and bulk material resistance) and will obviously be very small. C_{gs}' and C_{gd}' represent the capacitances from the source and drain region to the gate, due to the small overlap of the electrodes. C_{ds}, C_{gs} and C_{gd} are the encapsulation capacitances. There is also a further quantity C_{ss} representing the capacitance of the conducting path to the substrate. The channel has a distributed *RC* time-constant, which is shown as a lumped *RC* network, R_c and C_c. By assuming r_s' and r_d' are small compared with R_c the circuit may be simplified as shown in Fig. 3(b).

High Speed Effects

The remarks of the previous section have shown that the m.o.s.t. static characteristics make it potentially very suitable for low level gating especially in view of its zero offset voltage. It also has fairly low 'on' resistance and a high 'off' resistance. This section is concerned with the dynamic performance.

At high switching rates transient spikes are caused by capacitive coupling of the driving waveform from the gate to the signal path. Clearly the amplitude of these spikes must not be very much greater than the maximum signal level to be expected, otherwise the sense amplifier will be driven from its linear region, and the whole advantage of gating in this manner will be lost. The length of the spikes must be small (i.e. rapid decay) so that the delay between applying the drive and the gate becoming either fully on or off is a minimum.

The final detection of a '1' or '0' in the presence of

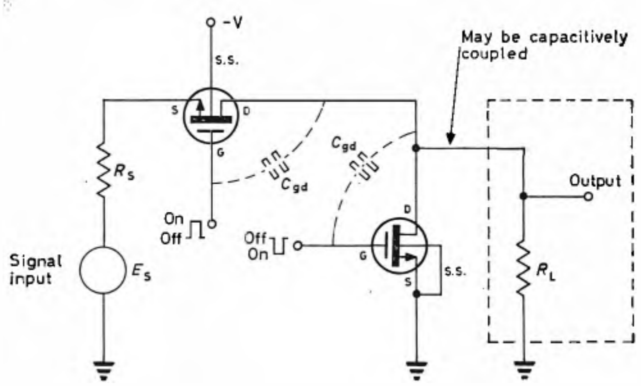


Fig. 4. Series-shunt circuit

noise is usually achieved by comparing the amplitude of the amplified output waveform with a fixed reference level at a specific time. It is necessary that the bandwidth of the gate at the 'front end' of the amplifier is such that there is a minimum loss of detail in the signal waveform. (i.e. no 'rounding off'.)

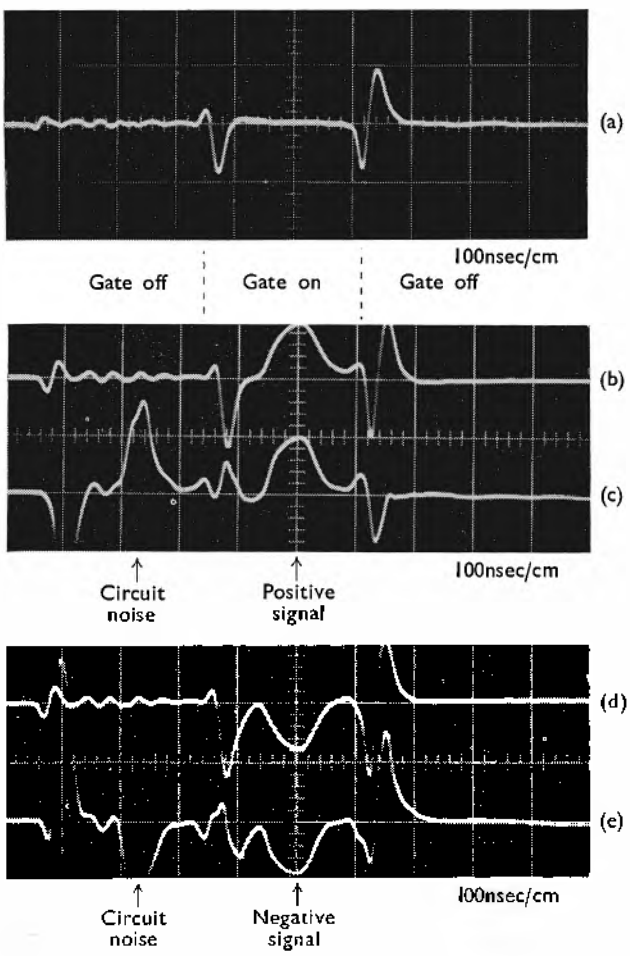
Practical Circuit

As a practical example, a low level gating circuit was designed to strobe the input of a sense amplifier in a computer magnetic core store with a cycle time of 1μsec. The circuit was to transmit a core output of approximately $\pm$50mV and reject noise pulses of up to $\pm$5V.

The circuit configuration chosen was the series-shunt arrangement as shown in Fig. 4 for the reasons stated previously. To ensure that the source/substrate pn junction of the series m.o.s.t. does not become forward biased, when the source is driven negative by the noise pulse, the substrate is fixed at a potential more negative than -5V. This results in the pinch-off voltage becoming more positive. The substrate of the shunt m.o.s.t. is connected to the source, so that its pinch-off voltage remains unchanged. The gates are driven on with V_g (gate voltage)

Fig. 5. Signal transmission of N-channel series shunt circuit

(a) Gate output showing gate transient. Input short-circuited
(b) Gate output with positive input signal
(c) Gate input : positive signal
(d) Gate output with negative input signal
(e) Gate input : negative signal



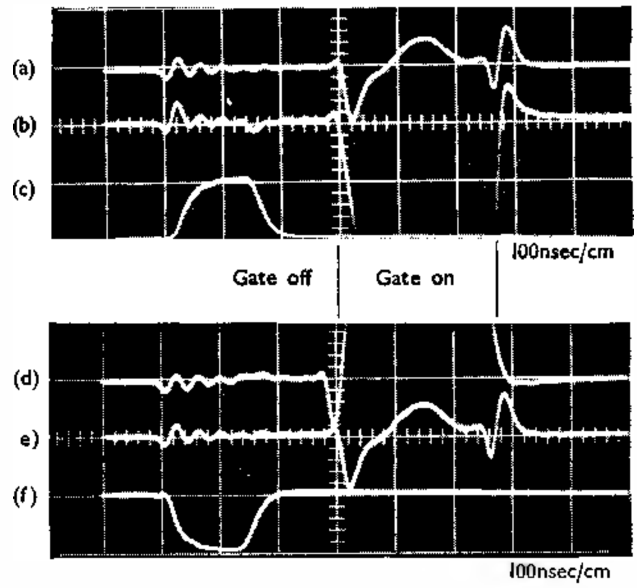


Fig. 6. Noise rejection

(a) Gate output with signal only as an input (for reference) (100mV/cm)
(b) Gate output with noise input (c) showing attenuated noise in gate 'off' period and the d.c. level from (c) during the 'on' period (100mV/cm)
(c) Noise input (with no signal present) with a negative d.c. level shift (5V/cm)
(d) Gate output with noise input (f) showing attenuated noise in the gate 'off' period and the d.c. level from (f) at the 'on' period (100mV/cm)
(e) Gate output with signal only as input (for reference) (100mV/cm)
(f) Noise input (with no signal present) with a positive d.c. level shift (5V/cm)

at least equal to V_p (pinch-off voltage) + 4V to give a small R_{sc}. The gate pulse was chosen to be 250nsec long (20nsec edges) and the signal pulse 100nsec long. The circuit is operated so that the series m.o.s.t. is normally high resistance, and is switched to the conducting state by the application of the gate pulse (shunt m.o.s.t. vice-versa).

Fig. 5(a) shows the output from the switch, across R_L, with no signal present at the input. The transient spikes are clearly visible. The total time required for the transient to die away is 80nsec. During the 'on' period, between the spikes, there is no significant offset voltage.

Fig. 5(b) and (c) shows the input and output for a positive signal; Fig. 5(d) and (e) for a negative signal. The d.c. attenuation to be expected is dependent upon the ratio of R_L to $R_L + R_{on}$ (where R_{on} is the 'on' resistance of the series m.o.s.t.). In this case the ratio is 0·95.

Fig. 6 illustrates the noise rejection capabilities of the circuit. Noise voltages of $\pm$5V at the input are attenuated by at least 1 000 : 1. The resultant noise fed through is thus very small compared with the signal level.

In the particular circuit used, the component values, referring to Fig. 3 and Fig. 5 were $R_s = 200\Omega$, $R_c = 200\Omega$

and $R_L=10k\Omega$. Assuming $C_1=4pF$, $C_2=2pF$ and $C_3=3pF$ a theoretical bandwidth of 40Mc/s from the input (source) to the output (drain) terminal was calculated. Measurements indicate that in practice this is reduced to approximately 35Mc/s by the shunting effect of stray circuit capacitance. Any integration of the signal, due to the series m.o.s.t. itself, will thus be small. This is confirmed by comparing the shapes of the input and output signals in Fig. 5.

Conclusion

It has been shown that the metal oxide semiconductor transistor, when compared with conventional semiconductor devices, is a closer approximation to the ideal low-level switch, because the leakage currents and pn junctions inherent in transistors and diodes are eliminated. This enables low-level signals, such as those from the sense wire of a ferrite core storage plane, to be successfully gated before further amplification. Present samples of N-channel m.o.s.t.'s in the series-shunt circuit enable switching speeds to be compatible with store cycle times as short as 1μsec. The gate bandwidth, in the conducting state, is of the order of 35Mc/s, which will cause no significant loss of detail in the signal waveform at these speeds. The '1' to '0' ratio is not reduced (i.e. no offset voltage) and a noise rejection of better than 1 000 : 1 can be achieved, with a signal gain of 0·95. For medium speed stores where the inhibit noise may be 1V or more, this m.o.s.t. gating technique will eliminate this noise and enable the sense amplifier to be designed for optimum signal handling characteristics alone.

C.R.T. Display Consoles for Digital Computers

Telefunken AG has recently delivered the first screen display console SAP 200-4, which as input/output equipment considerably extends the range of peripheral units for the type TR4 digital computer. Consoles of this description are being employed in growing degree in order to make the control of digital computers easier and yet more effective at the same time. The fields of application range from use in a data processing centre by a large number of remote points through the processing of technical-scientific problems to tasks in process control such as encountered, for example, in air traffic control.

The present edition of the screen display console was developed specifically for the latter function as a principal operating aid in automatic air traffic control. For this purpose it is used as a trial equipment, and it is provided with a number of technical functions which, after their operational testing, will be used to good advantage in modified or simplified form for general application purposes. These applications include operation in conjunction with the Telefunken high-capacity computer TR 440, which is being developed at the present time. Display consoles of this type allow multiple access to the computer from a large number of remote points.

Remote operation is rendered possible by the data transmission system DUE 300, with which the screen display console may be connected via a telephone line over an arbitrary distance to the TR 4 computer. For the transmission of display data as well as for the return transmission of control information, phase modulation and demodulation systems (type PHM 2400/1200/600), which the Company is developing, will be used: these systems allow a maximum transmission rate of 2 400 bits/sec in both directions.

For display the console is fitted with a cathode-ray tube having a useful screen diameter of 50cm. On the screen a maximum of 63 different alpha-numeric characters and symbols as well as elements made up of straight lines may be displayed in arbitrary arrangement. This display is repeated about 50 times a second on a non-persistent screen phosphor to produce a flickerless daylight-bright image, on which dynamic events may be presented without loss of focus. The electronic control, which is installed separate from the screen display console, contains a display repetition storage of 1 024 words of 24 bits each, which relieves the controlling computer of repetitions and renders possible remote operation of the screen display console.

The incorporated joystick control allows manual shifting of an electronically projected marker to any position on the screen and inserting the position co-ordinates of this marker in the computer. In this manner new computer information can be displayed at the point indicated by the marker or existing information may be erased or altered. Near the screen edge press-buttons are arranged with which the data referring to the position of the joystick marker may be transmitted to the computer. Further buttons are used for the input of general data, which are arbitrarily interpreted by the computer depending on the employment of the screen display console.

When the screen display console is employed for air traffic control, it is used primarily for the presentation of synthetic position data for the controller and for the input of a.t.c. data and instructions in the computer. A further display facility, which has great merits particularly for air traffic control, is the presentation of raw radar displays in television form and the insertion of additional computer information during the flyback period.

The data transmission system DUE 300 allows the reproduction at a site remote from the computer of the technical conditions of a TR 4 input/output channel so that at the remote site up to eight TR 4 peripheral units may be operated with their intermediate electronic units for direct TR 4 connexion. In conjunction with the data transmission system DUE 300 the electronic data check unit DSE 300 may be used, which greatly reduces the error rate on transmission.

A display unit specifically designed for automatic air-traffic control use.

Further Developments in the Design of Minimal NOR (and NAND) Combinational Switching Circuits for N-Variables

By D. Zissos*, B.Sc., A.M.I.E.R.E., and G. W. Copperwhite*, B.Sc.

An all-algebraic step-by-step method of designing minimal NOR *(and* NAND*) combinational switching circuits is described. The method essentially is one of optimizing the algebraic expression of the function to be realized in three steps. Theoretical justification of the design steps is given. Several examples are used to illustrate the method, including a full adder/subtractor unit.*

(Voir page 488 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 495)

MINIMAL design at present is essential for economy and reliability, although this may not be so in future with the advent of microminiaturization. A switching circuit is minimal if it uses the minimum number of switching elements and the minimum number of connexions, provided the first condition is satisfied.

In a previous article[2] a semi-mathematical method of designing NOR (and NAND) switching circuits has been described. The method essentially was based on the modification of AND terms in minimal OR-AND expressions, to a form that can be realized more minimally. Further work has resulted in the complete mathematization of the original method. This has been made possible by the evolution of a merging process, which allows two (AND/OR)/AND expressions to merge. The merged function can be realized more minimally than if it were realized in its original form.

The new method allows readily the detection, classification and elimination of static hazards. Due to lack of space, static hazards will not be dealt with in this article.

Several examples are used to illustrate this method.

Design Philosophy

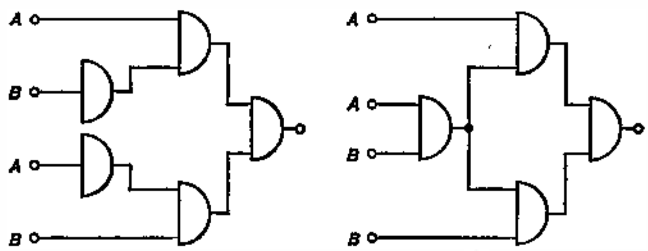
The design philosophy developed deals with the steady-state behaviour of combinational NOR circuits only. Their transient behaviour may be found in practice to be different from the one specified by their equations or excitation tables. This is basically because of the difference in the time delays of a signal reaching the same NOR module through two different paths. The result of this difference in the time delays is the possible appearance of narrow pulses at some point(s) in the circuit immediately following the change of a variable, when the signal at that point should have remained static. These signal spikes are defined as static hazards. The design philosophy followed by the authors is that of first designing the circuit in its absolute minimal form, ignoring the phenomenon of static hazards. Before the circuit is realized, the possible presence of any static hazards is next determined. If the static hazards present are considered to be critical, they are eliminated, otherwise they are allowed to exist in the circuit.

The design philosophy was developed to allow a set of simple rules to be produced that can be used in the design of N-variable minimal NOR circuits. Although this approach has been proved to work for three variables[1] it was not possible to test it for a larger number of variables since no list of minimal NOR circuits of more than three-variables, to the authors' knowledge, has been published.

* *Liverpool College of Technology.*

A minimal NOR/NOR circuit is defined as a circuit that uses the minimum number of modules and the number of connexions is a minimum, provided the first condition is satisfied.

Similarly a minimal NOR/NOR expression is defined as one that contains the minimum number of NOR functions and the number of signals in each NOR function is a minimum, provided condition (1) is satisfied.



(a) (b)

Fig. 1. (a) Non-minimal circuit realized directly from minimal NOR/NOR expression of $f = AB + \overline{AB}$

(b) Minimal circuit realized directly from optimal (non-minimal) expression of the same function

For example:

$$f = \overline{\overline{A+\overline{A}B+C} + \overline{X+Y+WY} + \overline{X+Y+WY}}$$

is not minimal, because the third NOR function $\overline{X+Y+WY}$ is redundant, being a duplicate of the second one, and the signals in the first NOR function can be simplified to $A + B + C$. Similarly the signals of the second NOR function can be reduced to $X + Y$. Both these simplifications are carried out using basic Boolean theorems. See appendix.

Since every NOR function is realized by a NOR module, it follows that the direct realization of a minimal NOR/NOR expression results in a minimal NOR/NOR circuit.

If the complements of the input signals are assumed not available, then the equation has to be optimized, before it is minimally realized using the substitution technique[2]. See also appendix on Substitution Signals.

For example:

$$f = \overline{\overline{\overline{A} + B} + \overline{A + \overline{B}}}$$

is minimal. Realizing this equation results in the non-minimal circuit of Fig. 1(a), whereas realizing its optimized (non-minimal) form

$$f = \overline{\overline{\overline{A + B} + B} + \overline{A + \overline{A + B}}}$$

results in a minimal circuit, as shown in Fig. 1(b).

Before describing the design philosophy it will be useful to define a simple NOR function as being the NOR function

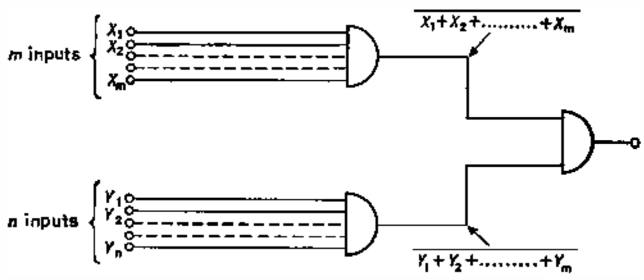


Fig. 2. Circuit derived from $f=\overline{\overline{X_1+X_2+\dots X_m}+\overline{Y_1+Y_2+\dots Y_n}}$

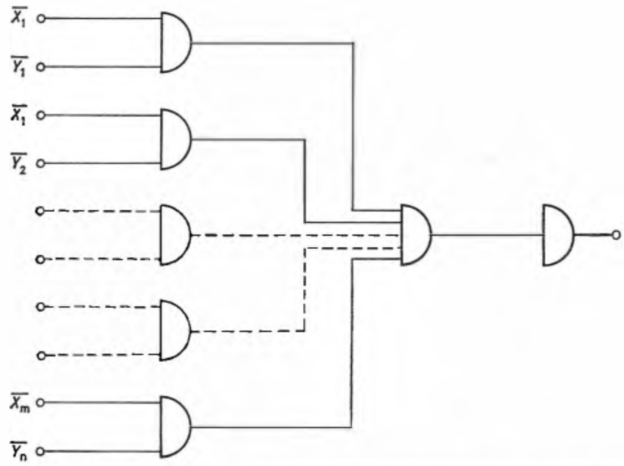


Fig. 3. Circuit realizing $f=\overline{\overline{\overline{X_1}+\overline{Y_1}}+\overline{\overline{X_1}+\overline{Y_2}}+\dots+\overline{\overline{X_1}+\overline{Y_n}}+\overline{\overline{X_m}+\overline{Y_n}}}$

of single primary signals only, such as $\overline{A+B+\overline{C}}$. Since a product of all-inverted-primary-signals is produced by one module, the above definition will be extended to include such products, for example:

$\overline{A+B+\overline{C}\,\overline{D}}$ will be classified as a simple NOR function.

The simple-NOR/NOR form of a given switching function may be derived directly from its simple-OR/AND form by double inverting, for example:

$$f=(A+B)\,(\overline{A}+\overline{B})$$
$$=\overline{\overline{(A+B)\,(\overline{A}+\overline{B})}}$$
$$=\overline{\overline{A+B}+\overline{\overline{A}+\overline{B}}}$$

Since each simple NOR signal in a simple NOR/NOR switching function corresponds to a NOR module in its circuit realization, the next step is to merge as many simple NOR signals in a simple-NOR/NOR function as possible. The following Merging theorem can be used.

Two NOR *functions in a* NOR-NOR *expression are merged, if all but one signal in each of the two* NOR *functions are the same, unless the two different signals are both in their true form.*

Proof: Consider the NOR-NOR equation:

$$f=\overline{\overline{f_1}+\overline{f_2}}\ \dots\dots\dots\dots\dots\dots\ (1)$$

where:

$$f_1=X_1+X_2+\dots\dots+X_m,\text{ and}$$
$$f_2=Y_1+Y_2+\dots\dots+Y_n.$$

f_1 is assumed to contain m terms and f_2 n terms.

Substituting the expressions for f_1 and f_2 in the original expression (1), the following simple-NOR/NOR equation for f is obtained:

$$f=\overline{\overline{X_1+X_2+\dots X_m}+\overline{Y_1+Y_2+\dots Y_n}}\ \dots\ (2)$$

Realizing equation (2) directly, results in the circuit shown in Fig. (2). It is assumed of course that the number of terms in each NOR function does not exceed the maximum fan-in of the NOR module to which they refer.

The two NOR functions, namely $\overline{X_1+X_2+\dots X_m}$ and $\overline{Y_1+Y_2+\dots Y_n}$ are next added. Their sum can be conveniently derived by inverting f into OR/AND form through the process of double-inversion and multiplying out the terms in each OR term. This results in f being expressed in an AND/OR form. The AND/OR expression is then changed into its required NOR/NOR form by converting each product term into a NOR function and double inverting the whole expression. These steps are shown below:

$$f=\overline{\overline{X_1+X_2+\dots X_m}+\overline{Y_1+Y_2+\dots Y_n}}$$
$$f=(X_1+X_2+\dots X_m)\,(Y_1+Y_2+\dots Y_n)$$
$$=X_1Y_1+X_1Y_2+\dots+X_1Y_n+X_2Y_1+X_2Y_2+\dots X_2Y_n$$
$$+\dots X_mY_1+X_mY_2+\dots X_mY_n.$$
$$\overline{\overline{\overline{X_1}+\overline{Y_1}}+\overline{\overline{X_1}+\overline{Y_2}}+\dots\overline{\overline{X_1}+\overline{Y_n}}+\dots\overline{\overline{X_m}+\overline{Y_n}}}$$
$$\dots\dots\dots\dots\ (3)$$

If all the terms of f_1 and f_2 are assumed the same, except X_m and Y_n, equation (3) becomes:

$$f=\overline{\overline{X_1+X_2+X_3+\dots+\overline{\overline{\overline{X_m}+\overline{Y_n}}}}}\ \dots\dots\ (4)$$

Comparing the merged form of the function f with its initial simple-NOR/NOR form in equation (2), it is seen that the merged form contains five NOR signals, namely

$\overline{X_m}$, $\overline{Y_n}$, $(\overline{\overline{X_m}+\overline{Y_n}})$, $(\overline{X_1+X_2+\dots\overline{X_m}+\overline{Y_n}})$ and $(\overline{\overline{X_1+Y_2+\dots+\overline{X_m}+\overline{Y_n}}})$

compared to only three in equation (2). If, however, either X_m or Y_n is replaced by $\overline{X_m}$ or $\overline{Y_n}$ then the merged form will contain four NOR signals, namely

$\overline{Y_n}$ (or $\overline{X_m}$), $(\overline{X_m+\overline{Y_n}})$, and

$\overline{X_1+X_2+X_3\dots+(\overline{X_m+\overline{Y_n}})}$, $\overline{\overline{X_1+X_2+X_3\dots+\overline{X_m+\overline{Y_n}}}}$

and the initial simple-NOR/NOR function four NOR signals.

The circuit realizing directly equation (4), is shown in Fig. 4. It uses five modules and (X_m-1+6) connexions.

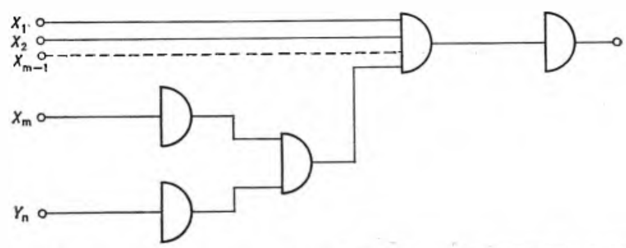
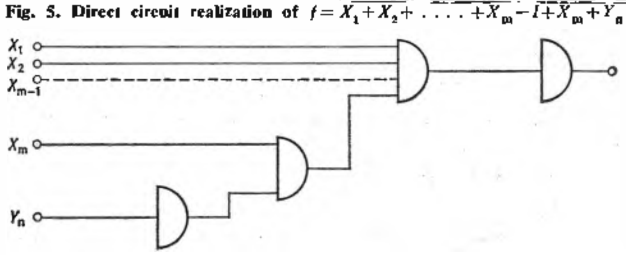


Fig. 4. Direct circuit realization of $f=\overline{\overline{X_1+X_2+\dots+X_m-1+\overline{\overline{X_m}+\overline{Y_n}}}}$

Fig. 5. Direct circuit realization of $f=\overline{\overline{X_1+X_2+\dots+X_m-1+\overline{X_m+\overline{Y_n}}}}$



Compared with the circuit realization of the function before merging, shown in Fig. 2, which uses three modules and $(X_m + Y_n)$ connexions it is not minimal. In other words if the two different signals are both in their true form, the two NOR functions are not merged.

If, however, one of the two different signals is in its inverted form then f becomes:

$$\overline{\overline{X_1 + X_2 + X_3 + \ldots\ldots + \overline{X_m} + Y_n}}$$

Equation (5) is realized directly in Fig. 5. Its circuit realization uses four modules and $(X_m - 1 + 5) = (X_m + 4)$ connexions.

Realizing the same function in its original form, namely equation (2) with X_m present in its inverted form will result in the circuit shown in Fig. 6, which uses five modules and $(X_m + Y_n + 3)$ connexions.

Comparing both circuits it is seen that they both use the same number of modules, namely four, but circuit 5 (corresponding to the merged expression) uses $X_m + 4$ connexions compared to $X_m + Y_n + 3$ connexions that circuit 6 uses. Now $X_m + 4 > X_m + Y_n + 3$ except when $Y_n = 1$ in which case they are equal.

If both X_m and Y_n were inverted, then a direct realization of the simple-NOR/NOR expression would result in a

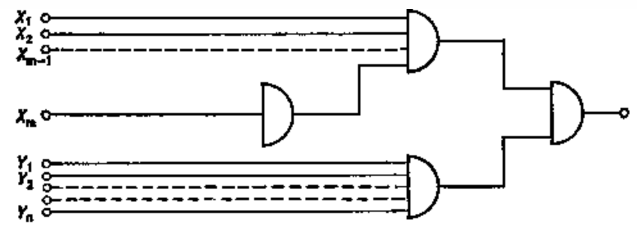


Fig. 6. Direct circuit realization of $f = \overline{X_1 + X_2 + \ldots + \overline{X_m} + Y_1 + Y_2 + \ldots Y_n}$

circuit using five modules, whereas the direct circuit realization of its merged form will result in a circuit using three modules.

It has been the authors' experience that it is easier to think of switching expressions in terms of AND/OR and OR/AND functions rather than NOR/NOR and NAND/NAND signals. Since a minimal NOR/NOR function can be converted to a minimal (AND/OR)/AND expression by the process of double inversion, it has been decided to use the (AND/OR)/AND form of a switching function in preference to its NOR/NOR form, when it is to be realized by a NOR circuit.

For example, the function $f = (A + B)\ (C + D)$ is not converted to its NOR/NOR form, but is realized in its present form directly, as shown in Fig. 7. In other words all the signals of an OR function are fed into a module separately, and the outputs of each of these modules are connected to a final module.

In terms of (AND-OR)/AND expressions the merging theorem may be stated as:

Two AND/OR functions in an (AND/OR)/AND expression are merged, if all but one of the AND signals in each of the two AND/OR functions are the same, unless the two different signals are both in their true form.

For example in

$$f = (A + B\,C + X)\ (A + B\,C + Y)$$

the two AND/OR functions, namely $A + BC + X$ and $A + B\,C + Y$ will not merge because X and Y are both in their true forms.

But in

$$f = (A + B\,C + \overline{X})\ (A + B\,C + Y)$$

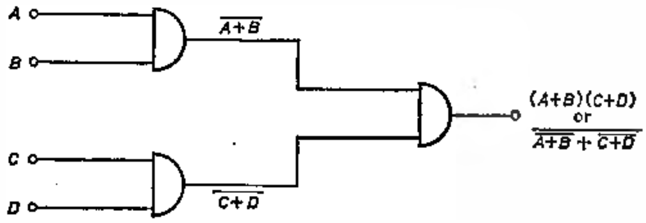


Fig. 7. Basic NOR block realizing the minimal OR/AND expression $(A + B)\ (C + D)$

the two AND/OR functions will merge to $A + B\,C + \overline{X}\,Y$, since X is present in its inverted form.

Similarly in

$$f = (A + B\,C + \overline{X})\ (A + B\,C + \overline{Y})$$

the two OR/AND functions will merge to $A + B\,C + \overline{X}\,\overline{Y}$, since both X and Y are present in their inverted forms.

Design Procedure

The design procedure can now be stated as:

Step (1). Derivation of the (preferred-AND/OR)/AND form of the function to be realized.

Step (2). Merging of the (AND/OR) functions.

Step (3). Introduction of substitution signals.

The design procedure is illustrated by means of six examples. Before introducing the examples, a minimization technique for (AND/OR)/AND expressions using the excitation table techniques as well as a layout for a NOR/NOR circuit is suggested.

The Minimization of AND/OR/AND Expressions Using the Excitation Table Technique

This technique is based on the fact that an excitation table can be split into N product excitation tables, provided, that each '1' entry in the parent table is transferred to the corresponding square in every product table, and each '0' entry is transferred in the corresponding square of at least one product table. This is because by splitting the table into product tables, mathematically, the function becomes equal to the product of AND/OR expressions.

Therefore, for the output to be equal to '1', each AND/OR *expression must be equal to '1', whereas for the output signal to be equal to '0' at least one of the* AND/OR *expressions must be equal to '0'.*

Example Consider the simplification of the AND/OR/AND function

$$f = (\overline{A}\,\overline{B} + B + \overline{C})\ (\overline{A}\,\overline{B} + C + A\,\overline{C})\ (A + \overline{B})$$

The product tables are drawn as shown in Fig. 8. Since there are two '0' entries in squares 2 of tables P_2 and P_3, one of them can change to '1'. Making the '0' entry of map P_2 equal to '1', results in $\overline{A}\,\overline{B} + C + A\,\overline{C}$ becoming equal to '1'. Similarly in table P_1 lift set $\overline{A}\overline{B}$ can be extended

Fig. 8. The product excitation tables of the functions $(\overline{A}\overline{B} + B + \overline{C})\ (\overline{A}\overline{B} + C + A\overline{C})\ (A + \overline{B})$

$\overline{A}\overline{B} + B + \overline{C}$ $[P_1]$

C \ AB	00	01	11	10
0	1_1	1_2	1_3	1_4
1	1_5	1_6	1_7	1_8

×

$\overline{A}\overline{B} + C + A\overline{C}$ $[P_2]$

C \ AB	00	01	11	10
0	1_1	0_2	1_3	1_4
1	1_5	1_6	1_7	1_8

×

$A + \overline{B}$ $[P_3]$

C \ AB	00	01	11	10
0	1_1	0_2	1_3	1_4
1	1_5	0_6	1_7	1_8

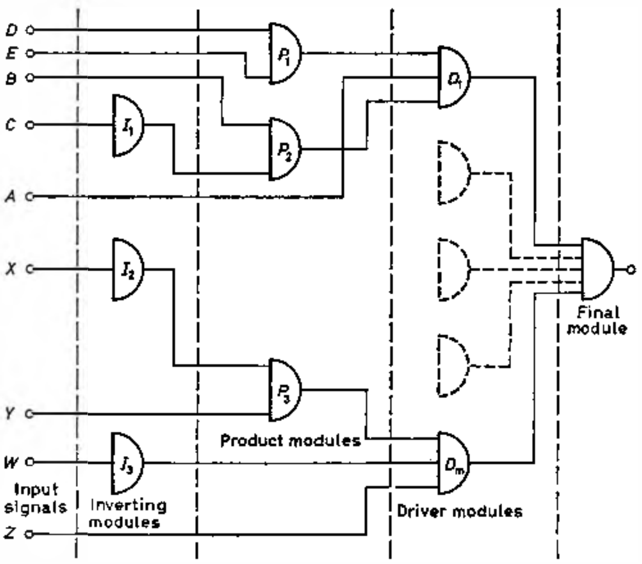


Fig. 9. Suggested layout for NOR/NOR circuits realizing AND/OR/AND expressions

to cover the '1' entries in squares 2 and 6 as well as those in squares 1 and 5, that is it can change to $\overline{A} + B + \overline{C}$. Hence the final minimal AND/OR/AND expression is:

$$(\overline{A} + B + \overline{C})\ (A + \overline{B})$$

Suggested Layout for NOR Circuits

The following layout has been found to assist greatly in realizing AND/OR/AND expressions by NOR circuits.

(1) Draw the *final module*—this is the module generating the actual function.

(2) Draw a module to correspond to each OR term in the AND/OR/AND expressions.

These modules are referred to as *driver modules*. See Fig. 9.

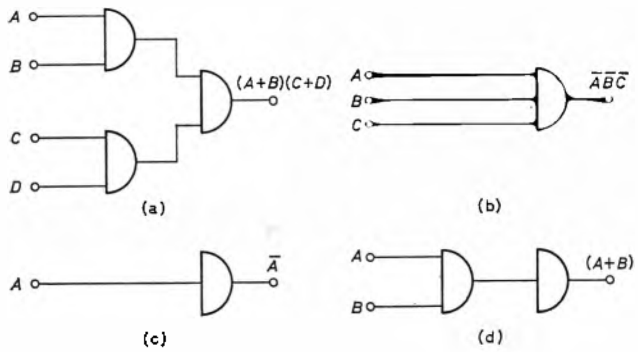
Each driver module is fed by the signals in its corresponding OR expression. For example:

Driver module D_1 is fed by A, $\overline{B}\,C$ and $\overline{D\,E}$; driver module $D\,M$ by $\overline{W}\,X\,\overline{Y}$ and Z, etc.

(3) The product terms are generated by what are referred to as product modules. It is helpful at this stage to think of the NOR module as producing the AND function of the complements of its input signals. For example, to generate $\overline{B}\,C$, B and $\overline{C}$ are fed into the product module P_2, see Fig. 8. To produce $\overline{D\,E}$, D and E are fed into P_1, etc.

Fig. 10. Basic NOR circuit configurations

(a) Basic circuit configuration for simple OR/AND functions
(b) Basic circuit configuration for AND functions
(c) Basic circuit configuration for invertors
(d) Basic circuit configuration for OR functions



Basic NOR Circuit Configurations

Fig. 10 shows the four basic NOR circuit configurations. It must, however, be appreciated that Figs. 10(b), 10(c) and 10(d) are actually derived from the circuit of Fig.. 10(a).

Design Problem (1)

Function to be realized:

$$f = A\,B\,C + \overline{A}\,\overline{B}\,\overline{C} \quad \ldots\ldots\ldots\ldots\ldots \quad (1)$$

Step (1)

Derivation of the (preferred AND/OR)/AND form of f. A preferred AND term is a product of all-inverted-signals, such as $\overline{X}\,\overline{Y}\,\overline{Z}$.

Applying theorem (4) in the appendix, f becomes:

$$(A + \overline{A}\,\overline{B}\,\overline{C})\ (B + \overline{A}\,\overline{B}\,\overline{C})\ (C + \overline{A}\,\overline{B}\,\overline{C}) \quad \ldots\ldots \quad (2)$$

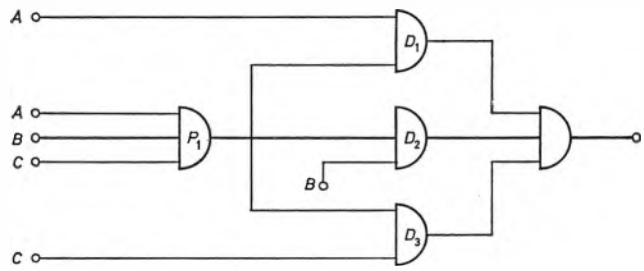


Fig. 11. Minimal NOR circuit realization of $f = \overline{A}\overline{B}\overline{C} + ABC$

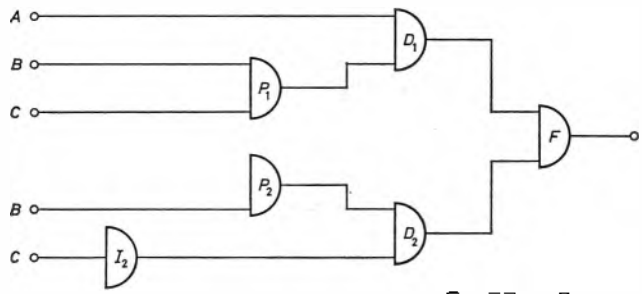


Fig. 12. Minimal circuit realization of $f = A\overline{C} + \overline{B}\overline{C} + A\overline{B}$

Simplifying each (AND/OR) function:

$$f = (A + \overline{B}\,\overline{C})\ (B + \overline{A}\,\overline{C})\ (C + \overline{A}\,\overline{B}) \quad \ldots\ldots \quad (3)$$

Step (2) Merging of (AND/OR) functions

No merging of the AND/OR functions of f can take place, since this will result in an all-true-signal-product term.

Step (3) Introduction of Substitution Signals

Every product term in equation (3) can be modified to $\overline{A}\,\overline{B}\,\overline{C}$. The final optimized form of the equataion, therefore, is:

$$f = (A + \overline{A}\,\overline{B}\,\overline{C})\ (B + \overline{A}\,\overline{B}\,\overline{C})\ (C + \overline{A}\,\overline{B}\,\overline{C}) \quad \ldots\ldots \quad (4)$$

and is realized in Fig. 11.

Design Problem (2)

Function to be realized:

$$f = A\,\overline{B} + A\,\overline{C} + \overline{B}\,\overline{C} \quad \ldots\ldots\ldots\ldots\ldots \quad (1)$$

Step (1)

$$\begin{aligned} f &= (A + A\,\overline{B} + \overline{B}\,\overline{C})\ (\overline{C} + A\,\overline{B} + \overline{B}\,\overline{C}) \\ &= (A(1 + \overline{B}) + \overline{B}\,\overline{C})\ (A\,\overline{B} + \overline{C}\,(1 + \overline{B})) \\ &= (A = \overline{B}\,\overline{C})\ (A\,\overline{B} + \overline{C}) \\ &= (A + \overline{B}\,\overline{C})\ (A + \overline{C})\ (\overline{B} + \overline{C}) \\ &= (A + \overline{B}\,\overline{C})\ (\overline{B} + \overline{C}) \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (2) \end{aligned}$$

Step (2)

No merging is possible.

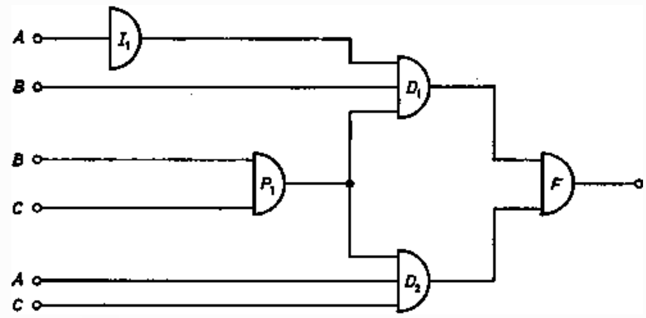


Fig. 13. Minimal circuit realization of $f = \overline{A} + \overline{B}\overline{C} + BC$

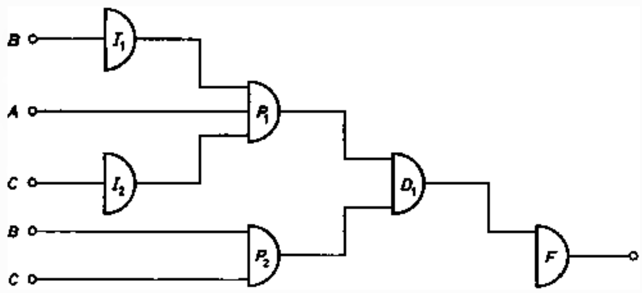


Fig. 14. Minimal circuit realization of $= \overline{A}BC + \overline{B}\overline{C}$

C \ AB	00	01	11	10
0	1: 0	2: 1	3: 1	4: 1
1	5: 0	6: 1	7: 1	8: 1

A+B

C \ AB	00	01	11	10
0	1: 0	2: 0	3: 1	4: 1
1	5: 1	6: 1	7: 1	8: 1

A+C

C \ AB	00	01	11	10
0	1: 0	2: 1	3: 1	4: 0
1	5: 1	6: 1	7: 0	8: 1

$\overline{B}C+B\overline{C}+\overline{A}BC$

Fig. 15. Product tables of $f = (A + B)(A + C)(\overline{B}C + B\overline{C} + \overline{A}BC)$

$$f = (A + \overline{B}\,\overline{C})\ (\overline{C} + \overline{B}) \quad \ldots\ldots\ldots\ldots \quad (3)$$

Step (3)

$\overline{B}\,\overline{C}$ can be modified to $\overline{A}\,\overline{B}\,\overline{C}$

$\overline{B}$,, ,, ,, ,, $\overline{B}C$

$\overline{C}$,, ,, ,, ,, $\overline{C}B$

Since no two substitution signals are the same, equation (3) is in its optimized form. See circuit in Fig. 12.

DESIGN PROBLEM (3)

Function to be realized:

$$f = \overline{A}C + A\overline{C} + BC + \overline{B}\,\overline{C} \quad \ldots\ldots\ldots \quad (1)$$

Step (1)

$$f = (B + \overline{A}C + A\overline{C} + \overline{B}\,\overline{C})\ (C + \overline{A}C + A\overline{C} + \overline{B}\,\overline{C})$$
$$= (B + \overline{A}C + A\overline{C} + \overline{C})\ (C(1 + \overline{A}) + A + \overline{B}))$$
$$= (B + \overline{A} + \overline{C}(1 + A))\ (C + A + \overline{B})$$
$$= (\overline{A} + B + \overline{C})\ (A + \overline{B} + C) \quad \ldots\ldots\ldots \quad (2)$$

Step (2)

No merging of the OR terms is possible.

Step (3)

$\overline{C}$ of the first OR function may be substituted by $\overline{B}\,\overline{C}$ because of B, and $\overline{B}$ also by BC because of C in the second OR function.

Therefore the optimized form of the function is:

$$f = (\overline{A} + B + \overline{B}\,\overline{C})\ (A + \overline{B}\,\overline{C} + C) \quad \ldots\ldots \quad (3)$$

See circuit in Fig. 13.

DESIGN PROBLEM (4)

Function to be realized:

$$f = \overline{A}BC + \overline{B}\,\overline{C} \quad \ldots\ldots\ldots\ldots \quad (1)$$

Step (1)

$$f = (\overline{A} + \overline{B}\,\overline{C})\ (B + \overline{B}\,\overline{C})\ (C + \overline{B}\,\overline{C}) \quad \ldots\ldots \quad (2)$$
$$= (\overline{A} + \overline{B}\,\overline{C})\ (B + \overline{C})\ (\overline{B} + C) \quad \ldots\ldots\ldots \quad (3)$$

Step (2)

Equation (2) indicates that merging is possible between $(\overline{A} + \overline{B}\,\overline{C})$ and $(B + \overline{B}\,\overline{C})$.

The function then becomes:

$$f = (\overline{A}B + \overline{B}\,\overline{C})\ (C + \overline{B}) \quad \ldots\ldots\ldots \quad (4)$$

Further merging is possible between the two AND/OR terms of equation (4), if $\overline{B}$ is replaced by $\overline{B}\,\overline{C}$ in the second bracket. Therefore, the merged form of the function then becomes:

$$f = (\overline{A}BC + \overline{B}\,\overline{C}) \quad \ldots\ldots\ldots \quad (5)$$

Equation (5) is the final optimized form and its circuit realization is shown in Fig. 14.

DESIGN PROBLEM (5)

Function to be realized:

$$f = AB\overline{C} + A\overline{B}C + \overline{A}BC \quad \ldots\ldots\ldots\ldots \quad (1)$$

Step (1)

$$f = (A + \overline{A}BC)\ (\overline{B}C + B\overline{C} + \overline{A}BC)$$
$$= (A + BC)\ (\overline{B}C + B\overline{C} + \overline{A}BC)$$
$$= (A + B)\ (A + C)\ (\overline{B}C + B\overline{C} + \overline{A}BC) \quad \ldots\ldots \quad (2)$$

Because of the relative complexity of the last expression, the map method will be used to simplify the expression of equation (2).

The excitation tables of the three (AND/OR) functions are drawn in Fig. 15. If the '1' entry in square 2 of table 1 is changed to '0', the corresponding '0' entry

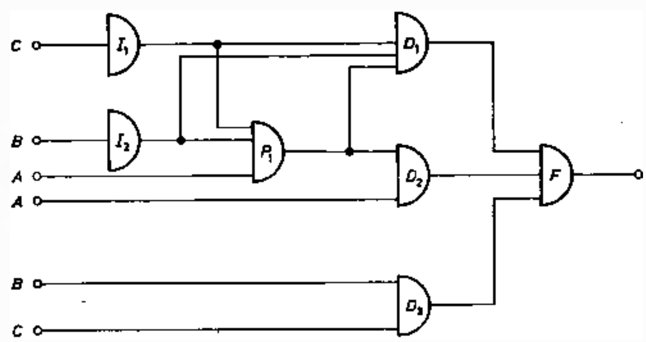


Fig. 16. Minimal circuit realizing $f = AB\overline{C} + A\overline{B}C + \overline{A}BC$

Fig. 17. Truth table for full-adder/full-subtractor unit

A INPUT	B INPUT	C OLD CARRY OLD BORROW	S SUM	K NEW CARRY	D DIFFERENCE	P NEW BORROW
0	0	0	0	0	0	0
1	0	0	1	0	1	0
0	1	0	1	0	1	1
0	0	1	1	0	1	1
1	0	1	0	1	0	0
1	1	0	0	1	0	0
0	1	1	0	1	0	1
1	1	1	1	1	1	1

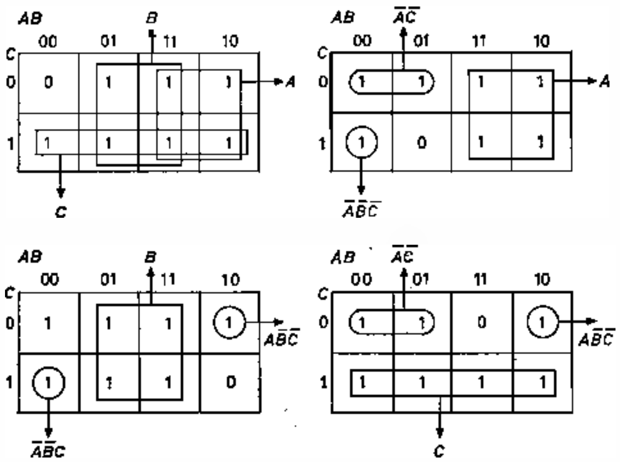


Fig. 18. Tables for OR functions of equations (2) from design problem (6)

in table 2 can be changed to '1'. The '0' entry in square 1 of table 2 can also be changed to '1', because the corresponding entry in table 1 is a '0'. That is the second table contains all '1's after the above two modifications, and hence it is redundant.

Mathematically this step corresponds to $(A + C)$ in equation (2) becoming redundant if $(A + B)$ in the same equation is changed to $A + \overline{A}\,B\,C$.

$$f = (A + \overline{A}\,B\,C)\;(\overline{B}\,C + B\,\overline{C} + \overline{A}\,B\,C)$$

Step (2)

No merging of terms is possible.

Step (3)

$\overline{B}\,C$ and $B\,\overline{C}$ can be replaced by $\overline{B}\,(B+C)$ and $\overline{C}\,(B+C)$.

Therefore the optimized form of the equation is:

$$\begin{aligned} f &= (A+\overline{A}\,B\,C)\;(\overline{B}\,(B+C)+\overline{C}\,(B+C)+\overline{A}\,B\,C) \\ &= (A+\overline{A}\,B\,C)\;(\overline{B}\,(B+C)+\overline{C}\,(B+C)+\overline{A}\,B\,C\,(B+C)) \\ &= (A+\overline{A}\,B\,C)\;(B+C)\;(\overline{B}+\overline{C}+\overline{A}\,B\,C) \quad \ldots\ldots\ldots\ldots (3) \end{aligned}$$

The circuit is realized in Fig. 16.

DESIGN PROBLEM (6)

A minimal full adder/subtractor unit is to be realized. The truth table is shown in Fig. 17.

From the truth table it is noted that both the sum and the difference columns are identical and their AND/OR minimal equation is

$$S = A\,\overline{B}\,\overline{C} + \overline{A}\,B\,\overline{C} + \overline{A}\,\overline{B}\,C + A\,B\,C \quad \ldots\ldots (1)$$

The unit is now designed in the specified three steps.

Step (1)

$$\begin{aligned} S &= A\,\overline{B}\,\overline{C}+\overline{A}\,B\,\overline{C}+\overline{A}\,\overline{B}\,C+A\,B\,C \\ &= (A+A\,\overline{B}\,\overline{C}+\overline{A}\,B\,\overline{C}+\overline{A}\,\overline{B}\,C) \\ &\qquad (B+A\,\overline{B}\,\overline{C}+\overline{A}\,B\,\overline{C}+\overline{A}\,\overline{B}\,C) \\ &\qquad (C+A\,\overline{B}\,\overline{C}+\overline{A}\,B\,\overline{C}+\overline{A}\,\overline{B}\,C) \\ &= (A+B\,\overline{C}+\overline{B}\,C)\;(B+A\overline{C}+\overline{A}C)\;(C+A\overline{B}+\overline{A}B) \\ &= (A+B+\overline{B}\,C)\;(A+\overline{C}+\overline{B}\,C)\;(B+A+\overline{A}\,C) \\ &\qquad (B+\overline{C}+\overline{A}\,C)\;(C+A+\overline{A}B)\;(C+\overline{B}+\overline{A}\,B) \\ &= (A+B+C)\;(A+\overline{B}+\overline{C})\;(\overline{A}+B+\overline{C})\;(\overline{A}+\overline{B}+C) \\ &\qquad \ldots\ldots\ldots (2) \end{aligned}$$

Because of the relative complexity of the equation the map method is used to simplify it. The tables of the OR's in equation (2) are drawn in Fig. 18.

If the '1' entries in the tables are included in the lift sets shown, then the '1'-entry in square 1 of table 3 could be modified to '0', allowing all the entries in table 1 to become '1' entries, that is $(A + B + C)$ in equation (2) is redundant.

The simplified form before merging of the function therefore is,

$$\begin{aligned} S = (A + \overline{A}\,\overline{B}\,C + \overline{A}\,\overline{C})\;(\overline{A}\,\overline{B}\,C + B + A\,\overline{B}\,\overline{C}) \\ (\overline{A}\,\overline{C} + A\,\overline{B}\,\overline{C} + C) \quad \ldots\ldots\ldots (3) \end{aligned}$$

Step (2)

No merging is possible, between any of the AND/OR terms.

Step (3)

No substitution signals can be used to simplify equation (3) further. Hence this equation for S is in its optimized form as realized in Fig. 19.

Generation of the Carry Signal

From the circuit of Fig. 19, the secondary signals available are:

Module	P_1	$S_1 = \overline{A}\,\overline{B}\,C$
,,	P_2	$S_2 = \overline{A}\,\overline{C}$
,,	P_3	$S_3 = A\,\overline{B}\,\overline{C}$
,,	D_1	$S_4 = \overline{A}\,B\,C$
,,	D_2	$S_5 = \overline{A}\,\overline{B}\,\overline{C}\,. + A\,\overline{B}\,C\,.$
,,	D_3	$S_6 = A\,B\,\overline{C}$

It is obvious that if signals S_1, S_2 and S_3 are fed into module C, then the carry signal will be generated.

Generation of the Borrow Signal

Similarly if signals S_3, S_5 and S_6 are fed into module B, the borrow signal will be generated at its output.

NAND/NAND CIRCUIT REALIZATION

To realize a given function using NAND modules interchange the 'pluses' and 'dots' in the original expression. The NOR circuit configuration corresponding to the derived function will realize the initial function, if the NOR modules are replaced by NAND modules. A useful point to bear in mind when transforming Boolean functions is to include all product terms in brackets before the transformation. The brackets are not affected by the transformation.

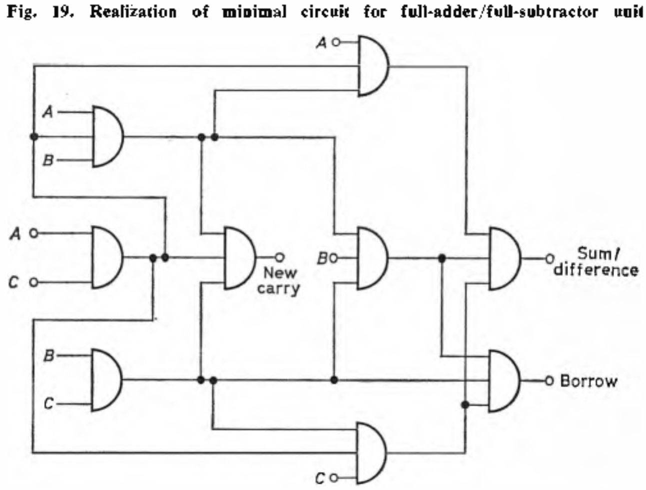
Example

Assume function $f = \overline{A}\,.\,\overline{C} + A\,.\,C$ is to be realized using NAND modules.

The product terms are first included in brackets,

$$f = (\overline{A}\,.\,\overline{C}) + (A\,.\,C)\,.$$

Fig. 19. Realization of minimal circuit for full-adder/full-subtractor unit



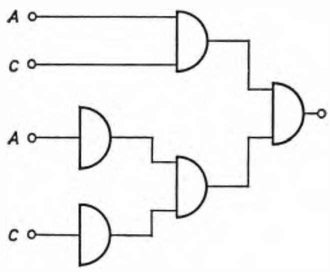


Fig. 20. Minimal circuit realizing $f = A\bar{C} + \bar{A}C$

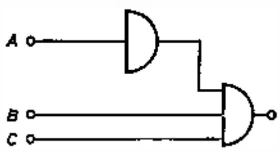


Fig. 21. NOR circuit realizing the function $A\bar{B}\bar{C}$

Interchanging the 'pluses' and 'dots', the modified form of the function becomes:

$$f = (\overline{A} \,.\, \overline{C}) \,.\, (A + C) =$$
$$A \,.\, \overline{A} + A \,.\, \overline{C} + \overline{A} \,.\, C + C \,.\, \overline{C}$$
$$= A \,.\, \overline{C} + \overline{A} \,.\, C \,.$$

The NOR circuit corresponding to this function is shown in Fig. 20. If in this circuit configuration the NOR modules are replaced by NAND modules, the function $A\,C + \overline{A}\,\overline{C}$ will be realized.

Conclusions

A theoretical basis for the design of minimal NOR/NAND combinational circuits has been described, in the form of the merging theorem. Based on the developed theory, three practical design steps have been suggested for NOR/NAND minimal circuits, when the complements of the input signals are assumed not available. If the complements of the input signals are available, then the first two design steps only need be used.

The circuits designed using this method are absolutely minimal, that is they use the minimum number of modules and the minimum number of connexions.

Static hazards, therefore, may or may not be present in a given circuit. Due to lack of space, the detection, classification and elimination of static hazards in such circuits has not been dealt with in this article.

Acknowledgements

The authors wish to acknowledge their gratitude to Mr. E. G. Lamb, Head of the Electrical Engineering Department, for his interest in the project and encouragement, and Mr. R. Haynes for his valuable contribution.

APPENDICES

Theorem (1)

$$A + A\,B = A$$

Proof $A + A\,B = A\,(1 + B) = A \,.\, 1$
$$= A$$

Theorem (2)

$$A + \overline{A}\,B = A + B\,.$$

Proof $A + \overline{A}\,B = A\,(1 + B) + \overline{A}\,B$
$$= A + A\,B + \overline{A}\,B$$
$$= A + B\,(A + \overline{A})$$
$$= A + B$$

Theorem (3)

$$A\,B + \overline{A}\,C + B\,C = A\,B + \overline{A}\,C$$

Proof $A\,B + \overline{A}\,C + B\,C = A\,B + \overline{A}\,C + (A + \overline{A})\,B\,C$
$$= A\,B + \overline{A}\,C + A\,B\,C + \overline{A}\,B\,C$$
$$= A\,B\,(1 + C) + \overline{A}\,C\,(1 + B)$$
$$= A\,B + \overline{A}\,C\,.$$

Theorem (4)

$$A + B\,C = (A + B)\ (A + C)$$

Proof $(A + B)\ (A + C) = A \,.\, A + A\,B + A\,C + B\,C$
$$= A + A\,B + A\,C + B\,C$$
$$= A\,(1 + B + C) + B\,C$$
$$= A + B\,C$$

Substitution Signals

In Fig. 21 signal $\overline{A}$ may be substituted by any of the following signals:

$$\overline{A}\,\overline{B},\ \overline{A}\,\overline{C},\ \overline{A}\,(\overline{B} + C),\ \overline{A}\,(B + \overline{C}),\ \overline{A}\,(\overline{B} + \overline{C}),\ \overline{A}\,\overline{B}\,\overline{C}\,.$$

Proof $P_1 = (A + B)\,\overline{B}\,\overline{C} = A\,\overline{B}\,\overline{C}$

$P2 = (A + C)\,\overline{B}\,\overline{C} = A\,\overline{B}\,\overline{C}$

$P3 = (A + B\,\overline{C})\,\overline{B}\,\overline{C} = A\,\overline{B}\,\overline{C}$

$P4 = (A + \overline{B}\,C)\,\overline{B}\,\overline{C} = A\,\overline{B}\,\overline{C}$

$P5 = (A + B\,C)\,\overline{B}\,\overline{C} = A\,\overline{B}\,\overline{C}$

$P6 = (A + B + C)\,\overline{B}\,\overline{C} = A\,\overline{B}\,\overline{C}$

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Low Cost Digital Equipment

Dawe Instruments Ltd., Western Avenue, Acton, London W3, has introduced a new series of digital modules and modular instruments which is claimed to be cheaper, more versatile and more extensive than any other on the market. Designated the Dawe 900 Series, it has been produced to considerably increase the field of application for digital equipment.

Standard assemblies (type 720) are the basis of the 900 Series. They include power supply units, display counters, clock-pulse generators, double-decade dividers, shaping-and-limiting amplifiers, control units and binary displays. Several variants of each type are available.

Instruments in the 900 Series so far include counters, timers, several digital frequency meters, ratiometers and tachometers. A digital voltmeter, printer, and 10Mc/s divider will follow shortly.

Most standard assemblies are printed-circuit plug-in units with gold-plated contacts, which are wholly compatible with—though not dependent on—one another. Although designed for Dawe's requirements they are equally suitable for building into customers' systems, both on the grounds of versatility and low cost. Solder as well as plug-in connexions are provided for those who wish to wire direct.

Because of the modular construction of the 900 Series and the permutations of standard assemblies this makes possible, the range of standard instruments available—even of one type—is extensive. To take a typical example, the type 900A/4 digital frequency meter comprises a control unit (standard assembly 114) which accepts pulses from a choice of sources, three 400kc/s display counters (SA111/1), one 1Mc/s display counter (SA141/1), a crystal-controlled five-decade clock pulse generator (SA117), and a double-decade divider (SA113). To these must be added a shaping-and limiting amplifier (SA115), a power supply (SA110), a switch unit (SA121), and a case (SA127), which altogether form a complete instrument.

The remarkably low cost of the 900 Series follows from the modular contruction, and from the fact that the same circuit functions are used whatever the overall purpose of the instrument. The digital displays, for instance, represent some 50 to 70 per cent of the cost of a digital instrument, and these can be considered as a large number of one type and manufactured by large-scale methods instead of in the usual small batches.

The modules, or standard assemblies, have been specially designed with symmetrical layout to ensure ease of production, and use a narrow range of component types to boost the buying quantities.

An 'in-phase' Amplifier and its Applications to Voltage Stabilizers

By G. N. Patchett*, B.Sc., Ph.D., M.I.E.E., M.I.E.R.E. M.I.E.E.E.

The device described was developed to reduce the phase angle between the input and output voltages of a precision a.c. voltage stabilizer. The equipment operates by measuring the phase of the output voltage relative to a reference and if there is a phase error a quadrature voltage is fed into the stabilizer to cancel this quadrature component of output voltage. With this device the phase error is reduced to 15′ under all conditions of load and power factor.

(Voir page 489 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 496)

THE device to be described was developed for use with an a.c. voltage stabilizer but has other possible uses. A small phase shift between the input and the output voltages of an a.c. voltage stabilizer is often not important. However, when testing wattmeters on a phantom load it is important that any stabilizer used to feed the current or voltage circuit should not introduce a phase shift or, if it does, this must be constant, or otherwise changes in the readings of the wattmeters will occur. The magnitude of the change will depend on the power factor at which the wattmeters are being tested. For example, if the test power factor is 0·8 then a change of 1° corresponds to a change of reading of 1·8 per cent. Similarly, if two stabilizers are used in an open delta circuit to obtain a three-phase supply then any phase changes caused by the stabilizers will result in changes of the three-phase supply voltage and, for accurate results, the phase shift of the stabilizers must be extremely small.

The circuit was designed specifically to operate with the precision voltage stabilizer previously described by the author[1,2]. In this stabilizer the measuring unit is a thermistor bridge and this operates a valve regulating unit (through the main amplifier) which bucks or boosts the supply so maintaining a constant voltage output. It has been shown[1] that all voltage sensitive bridges which use temperature sensitive devices produce unwanted voltage outputs due to the cyclical temperature variations of the devices. One such output is of three times the supply frequency and can be removed by a parallel-T or other filter. The other component is of fundamental frequency but 90° out of phase with the bridge supply voltage. There is also a quadrature component from a practical bridge circuit due to stray capacitances. To a large extent these two quadrature components can be eliminated in the output by a capacitor of suitable value across the thermistor. Unfortunately the compensation is only approximate and the required value of compensating capacitor varies with the ambient temperature. Provided that this quadrature bridge output is not large enough to overload the regulating stage of the stabilizer, the effect is only to produce a small phase shift between input and output voltages. For many purposes this is not important but for the applications mentioned earlier this variable phase shift is not permissible. Attempts by the author to compensate for variations in ambient temperature failed due to the fact that the compensation is trying to eliminate a large unwanted component (greater than the wanted component) and this component appears to depend on the thermistor ambient temperature, bridge transformer temperature, etc.

Therefore, in order to eliminate this phase shift it is required to build an amplifier which will amplify the in-phase component but reject the quadrature component,

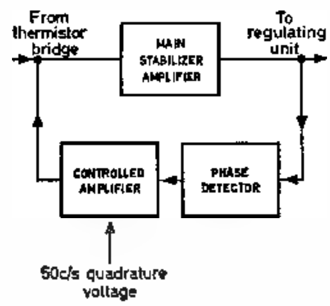

Fig. 1. Basic arrangement of 'in-phase' amplifier

i.e. an 'in-phase' amplifier. The author has not seen a similar device previously described. The basic principle of the device to perform this function is shown in Fig. 1. The phase detector gives an output approximately proportional to the magnitude of the quadrature component fed to it but no output when an in-phase component is fed to it. A 50c/s quadrature voltage (the stabilizer being designed to operate on 50c/s) is fed to the controlled amplifier and the output from this amplifier is fed to the input of the main amplifier together with the output from the thermistor bridge. The output from the controlled amplifier is arranged to be in such a direction that it opposes the normal quadrature output from the voltage sensitive bridge. The output of the controlled amplifier is varied by the output of the phase detector and its output is approximately proportional to the output of the phase detector. Thus, if there is a large quadrature voltage present in the output of the main amplifier the output of the phase detector will be large and there will be a large output from the controlled amplifier and this opposes the quadrature component from the voltage sensitive bridge (which was causing the large quadrature output from the main amplifier). Hence there is a large amount of negative feedback and, if the loop gain is high, little resultant output is obtained from the main amplifier. There is, of course, no negative feedback as regards the

* *Bradford Institute of Technology.*

in-phase component and the gain to this signal is unaltered. Any phase shift which occurs in the main amplifier which would give a quadrature component is also corrected for by this device. Expressions for the magnitude of the corrections are derived in the appendix.

The phase detector and controlled amplifier will now be described:

Phase Detector

This is really a gating circuit, the output of the main amplifier being sampled at a brief instant corresponding to zero voltage for an in-phase component. If there is a quadrature voltage present then the resultant voltage at this instant is not zero and an output is obtained from the phase detector. In the actual circuit the operation is reversed in the sense that there is normally an output from the phase detector and this is reduced by the presence of the quadrature component. By also reversing the operation of the controlled amplifier the overall result is the same as that already described. The circuit is shown in

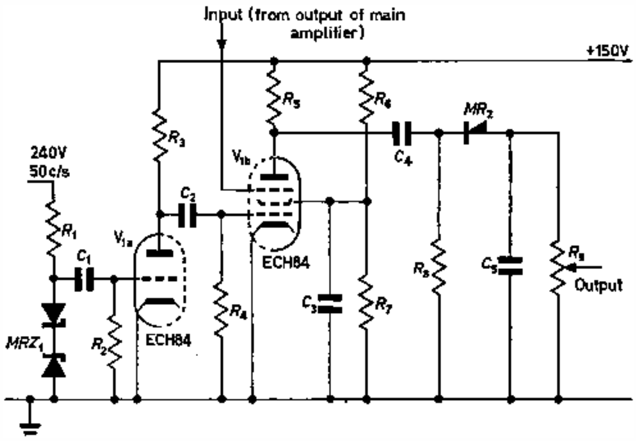

Fig. 2. Circuit of phase detector

Fig. 2. The double anode Zener diode MRZ_1 (symmetrical Zener diode or two Zener diodes in reverse series or back to back) is fed through R_1 from a suitable in-phase voltage. This produces a square wave voltage across MRZ_1 of about 30V peak-to-peak (with the particular Zener diode used) with a rise time of approximately 250μsec. The square wave is differentiated by C_1 and R_2 producing positive and negative pulses corresponding to zero voltage for an in-phase voltage. The positive pulse is largely removed by grid current of V_{1a} and the negative pulse is amplified producing a positive pulse of about 25V and duration 250μsec across R_3. This is fed to the grid of V_{1b}. The positive pulse drives V_{1b} into grid current which charges C_2 (self bias action) so that V_{1b} is cut off except during the period of the pulse. The output from the main stabilizer amplifier is fed to g_3 of V_{1b}. With no voltage fed to g_3 a negative pulse is produced across R_5 corresponding to the positive sampling pulse on G_1. If the voltage fed to g_3 does not contain a quadrature component, the voltage is substantially zero at the instant of sampling and the output pulse is of the same amplitude as when no signal is fed to g_3. If there is a quadrature voltage present then the voltage on g_3, at the instant of sampling, is either positive or negative during the period of the sampling pulse. If the voltage is negative the magnitude of the pulse across R_5 is reduced. If positive the pulse magnitude is increased to some extent but this direction of operation is not used. In the particular circuit used by the author the pulse is reduced to zero with a quadrature voltage of about 3V on g_3. With a voltage of 35V on g_3 the magnitude of the pulse is changed from 33V to zero if the phase of the voltage is changed from zero to about 6°. The pulses from V_{1b} are rectified by MR_2 to produce a negative voltage, C_5 acting as a smoothing capacitor.

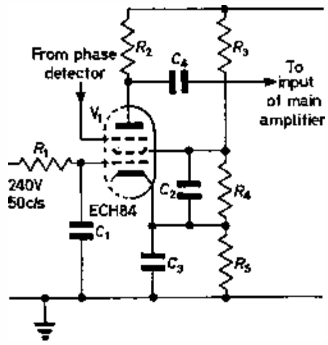

Fig. 3. Circuit of controlled amplifier

Controlled Amplifier

The circuit is shown in Fig. 3. An in-phase 50c/s voltage is fed to R_1-C_1. The reactance of C_1 is arranged to be low compared with the resistance R_1 so that the voltage across C_1, which is fed to g_1 of V_1, is 90° lagging the reference voltage. This is amplified by the valve, producing an output across R_2 which is fed to the input of the main amplifier, through an isolating resistor to prevent loading the thermistor bridge. The magnitude of the output is controlled by the output of the phase detector fed to g_3. Thus if there is no quadrature voltage in the output of the main amplifier the output of the phase detector is large and the voltage on g_3 of V_1 is such (by adjustment of R_9 of Fig. 2) that there is just no output from V_1. If there is a quadrature component from the main amplifier the output from the phase detector is

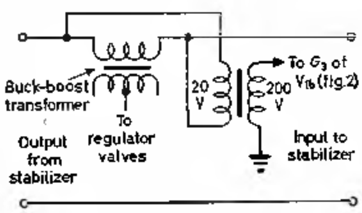

Fig. 4. Connexion of phase detector to buck-boost transformer

reduced and the output from V_1 increases to cancel the original quadrature component output of the main amplifier.

Although this arrangement effectively reduced the phase shift of the stabilizer due to the quadrature voltage output of the thermistor bridge and phase shift in the main amplifier, it did not correct for phase changes in the correcting voltage caused by changes in the phase of the load current. These are caused by the impedance of the output transformer and regulator valves. To overcome this the phase detector was connected as shown in Fig. 4 so that it was fed with the actual buck and boost correcting voltage. By this means the device was able to correct for phase shift produced by any cause in the stabilizer. As the device only works in one direction (it can be modified to work in both directions but it is not necessary in this application) it is arranged that the unwanted component of the thermistor bridge is always in one direction (by arranging that, over the normal ambient temperature range, the bridge is under compensated by the use of a smaller than normal compensating capacitor).

With the phase corrector in circuit the following results were obtained:

	Phase error
No load	5′ lag
Full load unity power factor	5′ lag
Full load 0·15 power factor lagging	10′ lag
Full load zero power factor leading	5′ lead
Total variation 15′	

The actual angle must be taken as approximate due to the difficulty of measuring such small angles. Variation of the compensating capacitor of the thermistor bridge from its normal value (0·002μF) to zero produced no measurable phase change in the output voltage of the stabilizer. Without the phase corrector this caused a phase change of about 2° with some overloading of the regulating stage.

Although the device has been described for use with a particular stabilizer it can be used to correct the phase angle of any circuit or device provided an in-phase reference voltage is available. It would probably have applications in voltage and current amplifiers designed to feed wattmeters where the phase error must be small.

Appendix

(1) Elimination of Quadrature Input Signal

See Fig. 5.

Let the input signal (v_i) be $A + j.B$ where $j.B$ is the component to be rejected.

Let the resultant output voltage (v_o) be $C + j.D$.

Let the output voltage of the controlled amplifier (v_f) be $-j.k.D$.

(Note that v_f is taken as being proportional to the quadrature output component of the main amplifier)

k is a constant relating the output v_f of the controlled amplifier per 1 volt of quadrature signal from the main amplifier.

Let gain of main amplifier be M

$$v_a = v_i + v_f \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (1)$$

and $v_o = M.v_a$

Substituting from equation (1):

$$v_o = M(v_i + v_f)$$

Substituting the values of v_i and v_f:

$$v_o = M(A + j.B) + M(-j.k.D) =$$
$$M.A + j.M.B. - j.M.k.D \quad \ldots\ldots \quad (2)$$

but:

$$v_o = C + j.D \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (3)$$

Equating equations (2) and (3):

$$C + j.D = M.A + j.M.B. - j.M.k.D$$

or $C = M.A$

$$D = M.B - M.k.D$$

or $D(1 + M.k) = M.B$

$$D = \frac{M.B}{1 + M.k}$$

or $v_o = M.A + j\dfrac{M.B}{1 + M.k}$

with no corrector (i.e. $k = 0$)

$v_o = M.A + M.B$ as expected.

With the corrector in circuit the quadrature component is reduced by the factor $1 + M.k$.

Example

Suppose $M = 1000$ and $k = 1$

Let $v_i = (1 + j.1)10^{-3}$ that is, a phase angle of 45°

$$v_o = M.A + j\frac{M.B}{1 + M.k}$$
$$= 1000 \times 10^{-3} + j\frac{1000}{1 + 1000}10^{-3}$$
$$= 1 + j.10^{-3} \text{ compared with } 1 + j.1 \text{ with no corrector.}$$

That is the quadrature component is reduced by a factor of approximately 1000 and the phase angle reduced from 45° to

$$10^{-3}/1 \times 180/\pi = 0{\cdot}06°$$

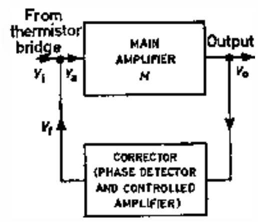

Fig. 5. Arrangement of device

(2) Correction for Phase Error in Amplifier

Let gain $= M + j.N$ and the input voltage $= A$. Other values as previously.

$$v_o = (M + j.N)v_a = (M + j.N)(v_i + v_f)$$
$$= (M + j.N)(A - j.k.D)$$
$$= M.A + j.k.M.D + j.N.A + N.k.D$$

but this equals $C + j.D$

Equating terms:

$$C = M.A + N.k\frac{N.A}{1 + M.k}$$
$$D = \frac{N.A}{1 + M.k}$$

Hence $v_o = M.A + \dfrac{N^2.k.A}{1 + M.k} + j\dfrac{N.A}{1 + M.k}$

With no corrector $v_o = M.A + N.A$

Example

Let $M = 1000$ and $N = 176$, i.e. phase angle is 10°

$k = 1$

$A = 10^{-3}$

$$v_o = 1000 \times 10^{-3} + \frac{176^2 \times 1 \times 10^{-3}}{1 + 1000} + j\frac{176 \times 10^{-3}}{1 + 1000}$$
$$= 1{\cdot}031 + j.0{\cdot}000176$$

Thus a small in-phase component is added (approximately 3 per cent) and the quadrature component becomes negligible. The phase angle is reduced to 10^{-2} degree.

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Impedance Correction in the Vertical Distribution Feeder System of a Null-filled High Gain Aerial

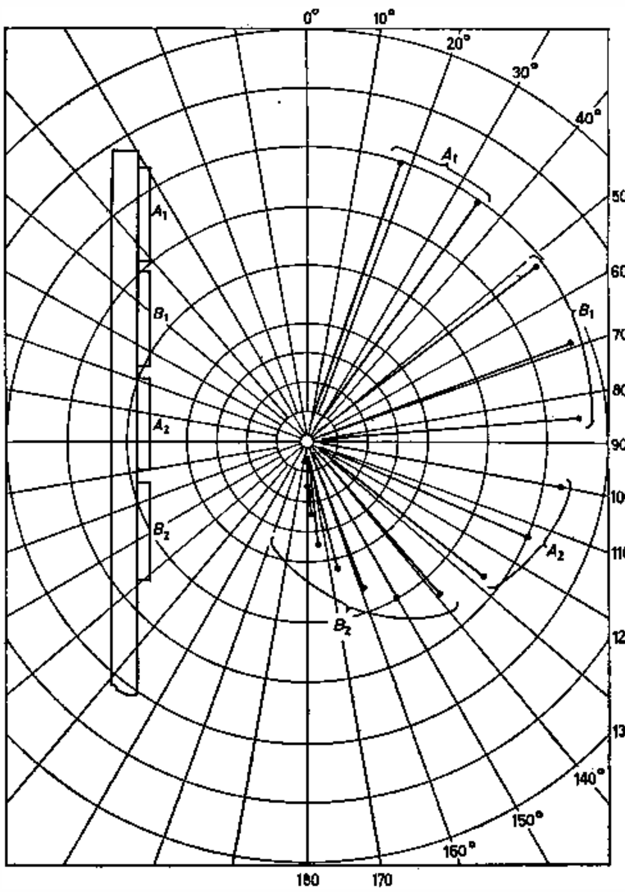
By A. B. Shone*, B.Eng., A.M.I.E.E.

It is the national policy in the United Kingdom to co-site transmissions in the u.h.f. band and in many cases it becomes economic to use the whole of the available aerial aperture for one common aerial carrying up to four u.h.f. channels. This arrangement poses the problems of obtaining the required impedance match over a wide bandwidth and also, with colour transmissions particularly in mind, obtaining a smooth vertical pattern which should, if possible, approach a cosecant law. The article suggests a possible way of achieving both these results which would be applicable providing the beam tilt inherent in this type of solution is suitable.

(Voir page 489 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 496)

IN the past the general policy of aerial manufacturers has been to use phase rotation in the horizontal plane to improve the input impedance over a broad frequency band. This introduces several complications: firstly, with a phase rotating system it is necessary to mount the radiating elements at specific physical spacings to obtain the best radiation patterns and these spacings are not always convenient or readily adjustable; secondly, the radiation patterns of the individual faces of the aerial become asymmetrical and in general the overall radiation patterns obtained are not so good as can be obtained with a co-phased system: thirdly, considerable development time is wasted in trying to find a suitable compromise between the acceptable horizontal radiation pattern and an acceptable impedance bandwidth: fourthly, directional horizontal radiation patterns are not in general practical except for a limited number of patterns having a mirror image type of symmetry which lends itself to impedance correction. If therefore the correction of the impedance can be done in the distribution harness to the vertical tiers the above difficulties are eliminated and it should be practicable to adjust the horizontal tiers for optimum radiation pattern (which may be directional if desired) quite independently of the adjustments of the distribution harness for optimum correction of the impedance bandwidth. A method for doing this is described in the following paragraphs. The procedure proposed involves the use of both amplitude and phase variation over the aperture to minimize the vector sum of the reflected waves from individual tiers of panels. With this arrangement there is an inherent tilt of the main lobe of the vertical radiation pattern varying from about 1° for a 32 wavelength aerial to 2·5° for a 12 wavelength aerial, but these values are likely to be acceptable in practice and may even be desirable as these beam tilts are always coupled with a drop in radiation of about 3dB in the horizontal direction which helps to reduce the interference to adjacent stations operating on the same channels. The null-filling of the vertical radiation pattern using this system is very much better than the patterns that can be obtained with most other null-filling arrangements and does in fact result in an almost perfect distribution if the vertical aperture can be fed in a sufficient number of discrete sections. The loss due to null-filling by this method is a little over 1dB for the best arrangements and the maximum power fed to any tier is approximately twice the

Fig. 1. Theoretical amplitude and phase of 32 wavelength aerial



* *British Broadcasting Corporation.*

power which would be fed to the tier if null-filling by phase perturbation and constant amplitude were employed.

Basic Principle

The basic arrangement is based upon a sinusoidal amplitude and progressive phase distribution over the aperture to obtain good null-filling coupled with a quadrature feeding arrangement to one or more pairs of phase centres in the vertical aperture. It should be noted that a linear phase progression of 180° over the whole aperture will give substantially the same beam tilt as is obtained by dividing the aerial into two halves fed in quadrature or into four quarters fed with currents having a 0°, 45°, 90° and 135° phase relationship or six sections having a 0°, 30°, 60°, 90°, 120° and 150° phase relationship. If it is desired to feed the aerial by two main feeders the second arrangement is attractive and the quadrature feed by one feeder is between the first and third vertical sections of the aerial and the quadrature feed from the other feeder is between the second and fourth sections of the aerial. Due to the quadrature feed from each main feeder the reflected waves from the two sections of the aerial to which it is connected will return to the distribution transformer in anti-phase and can be absorbed in a power equalizing load providing the reflected voltages are equal.

Since however a sinusoidal amplitude distribution is to be employed the reflected waves will not be equal in amplitude unless the aperture is divided unequally between the four sections of the aerial in such a way that the vector sum of the reflected voltages from all the elements in each section (assuming every element has the same reflection coefficient) is the same. This can be achieved by making the phase increments over the aperture approximately proportional to the sinusoidal amplitude distribution over the aperture and at the same time keeping the total phase change over the aperture constant at

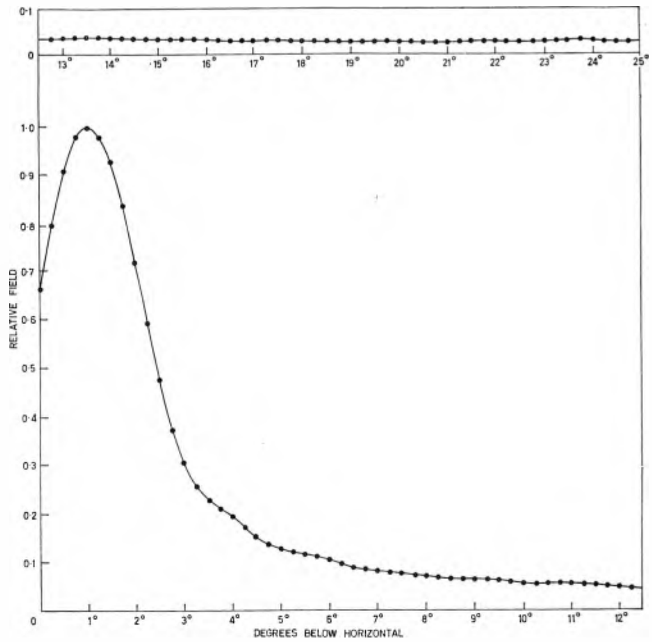


Fig. 3. Computed vertical radiation pattern

Fig. 2. Relative amplitude and phase of reflected waves

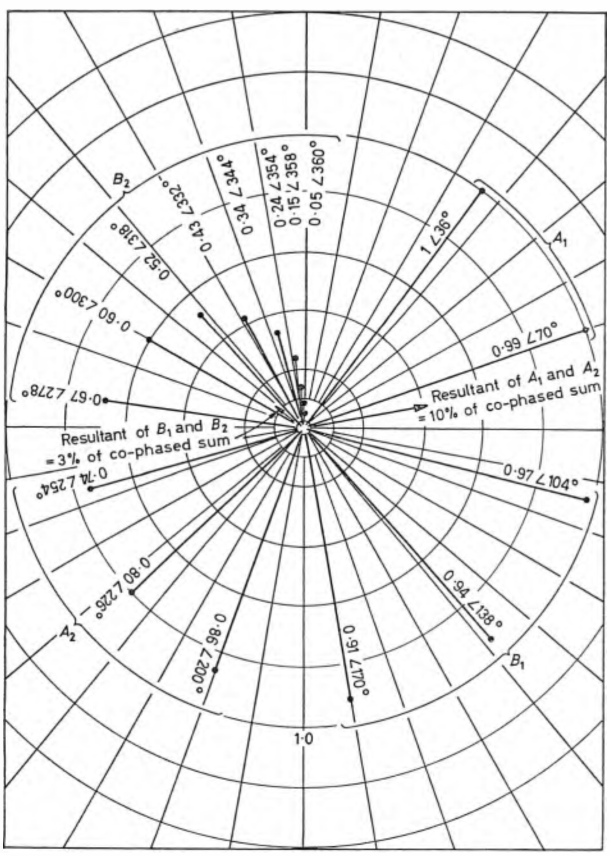
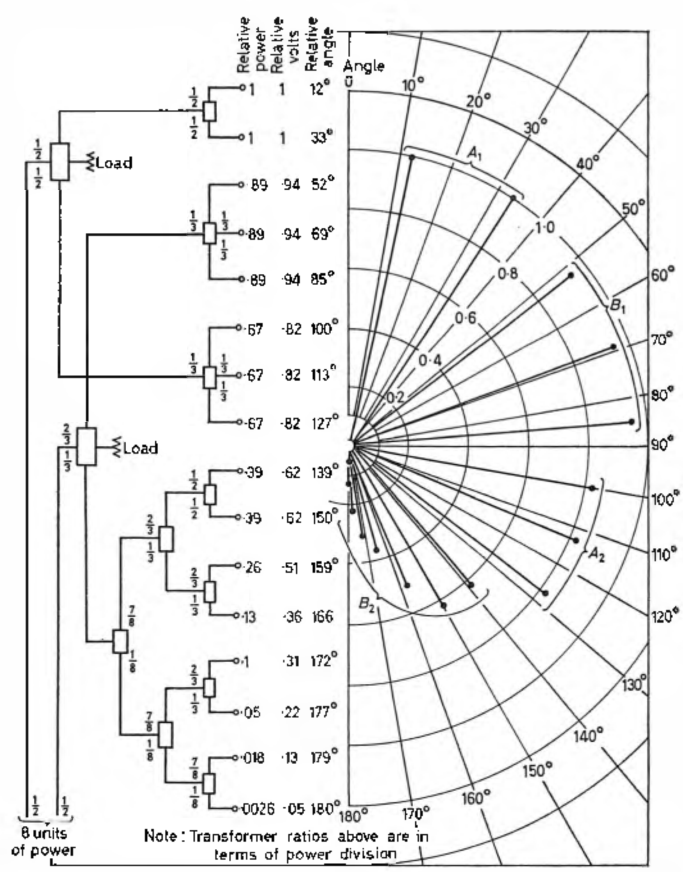


Fig. 4. Suitable distribution harness



180°. It will be appreciated that this is practicable where the number of feed points is large but becomes increasingly difficult as the number of feed points is reduced. Due to the mutual impedance between vertical tiers the reflection coefficients at the feed point to each vertical tier will not be constant but, if two wavelength panels are employed, the variation between tiers will be small.

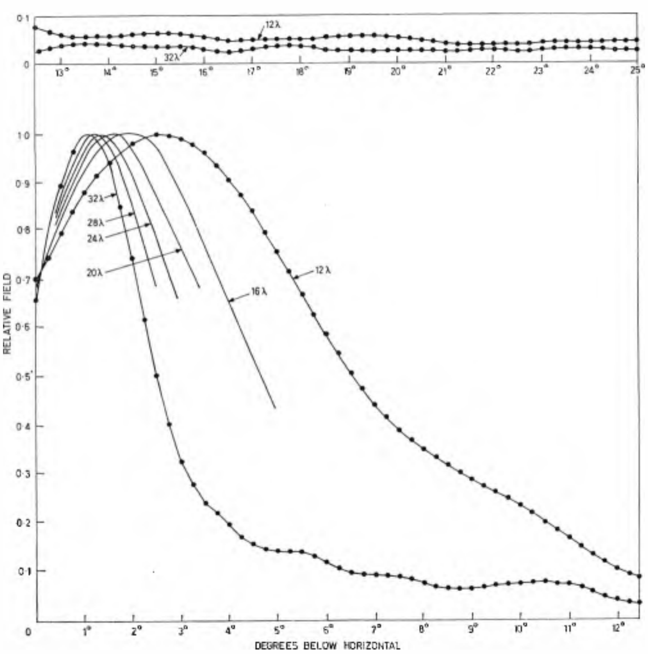
By way of example the theoretical and practical distribution over a 32 wavelength aerial will be illustrated. This is a typical case for which high gain u.h.f. aerials are currently required.

Fig. 1 shows the theoretical relative amplitude and phase for a 32 wavelength aerial split into 16 two wavelength panels divided up into 4 sections having respectively 2 panels, 3 panels, 3 panels, and 8 panels. On the drawing the sections are labelled A_1, B_1, A_2 and B_2 respectively. Assuming all the panels have the same reflection coefficient (in phase and amplitude), and also assuming perfect wideband transformers, the relative amplitude and phase of the reflected waves returning to the distribution transformer will be as in Fig. 2 and it will be seen that the vector sum of the reflected waves from the A_1 section of the aerial are in anti-phase and approximately equal to the vector sum of the reflected waves from the A_2 section. The same applies to the reflected waves from the B_1 and B_2 sections. The computed vertical radiation pattern of such an aerial is shown in Fig. 3 and the computed null-fill loss is 1·04dB.

Practical Arrangements for a 32 Wavelength Aerial

A suitable distribution harness (which derives currents in each tier approximately equivalent in amplitude and phase to those indicated in Fig. 1) is shown in Fig. 4. The arrangement is based upon the use of a limited number of standard transformer types which can be used for a variety of aerial sizes varying from 12 wavelengths up to 40 wavelengths. The computed vertical radiation pattern for a 32 wavelength aerial fed in the above way is shown in Fig. 5. On the same figure the computed patterns for other aerial apertures (fed in a similar manner) are shown

Fig. 5. Computed vertical distributed pattern for aerial fed as in Fig. 4



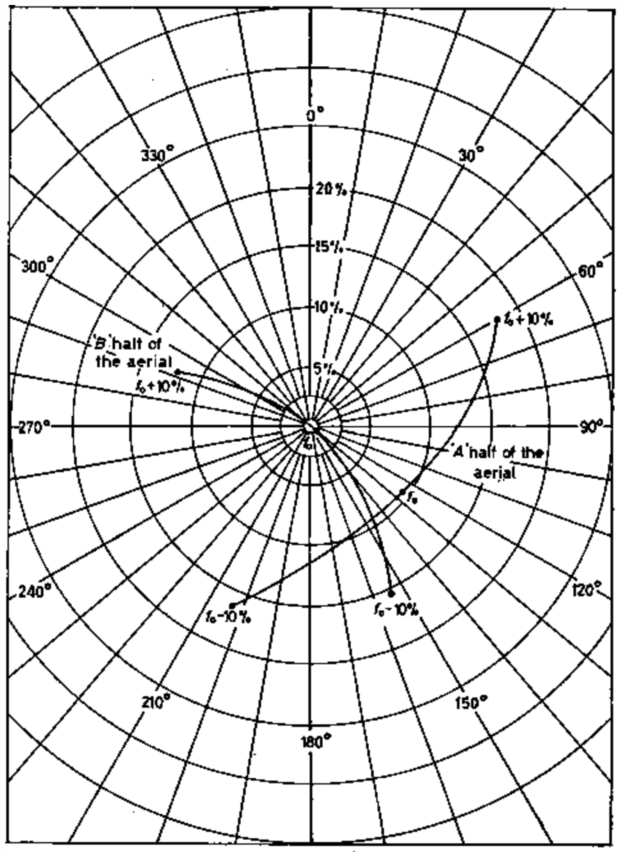


Fig. 6. Computed reflection coefficients

to indicate the order of the beam tilt. The computed null-fill loss in this case was 1·1dB. The computed reflection coefficients at the input to the power equalizing transformers at the top of each main feeder for a 100 per cent reflection at each level of the aerial are shown in Fig. 6. The improvements in impedance bandwidth obtainable in this case is 6:1 over ±10 per cent frequency band. This is typical of the results obtained for the other apertures considered.

Conclusions

The method described for obtaining a broad impedance bandwidth has a particularly good vertical pattern and is generally applicable if the beam tilt associated with the aperture to be employed is suitable and if the radiating elements with the highest power loading can handle approximately twice the mean power loading. It avoids the necessity for phase rotating the drives to the horizontal tiers which increases the chance of obtaining good directional and omnidirectional horizontal radiation patterns.

A 16 wavelength aerial with this type of distribution has been built for a low power relay station and proposals have been formulated for a group of high power 32 wavelength aerials with this arrangement.

Acknowledgments

The author wishes to thank the Director of Engineering of the British Broadcasting Corporation for permission to publish this article and his colleagues, Dr. R. G. Manton and J. A. C. Hugh, also C. H. Millard of the BBC Research Department for their valued co-operation in the work described above.

Finite Input Resistance Amplifier

By J. B. Matley, B.Sc., A.M.I.E.E.

This article presents and analyses a two stage amplifier with unity voltage gain employing a field effect transistor in the first stage capable of providing input resistance well in excess of 100MΩ. A modification to this circuit is discussed which enables the input resistance to be adjusted to infinity and an analysis for the modified circuit is presented.

(Voir page 489 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 496)

SINCE the advent of the field effect transistor to the semiconductor field, the inability of the simple transistor circuit to provide a high input resistance has been overcome. Input resistance greater than 100MΩ can readily be obtained by the field effect transistor without any circuit complexity.

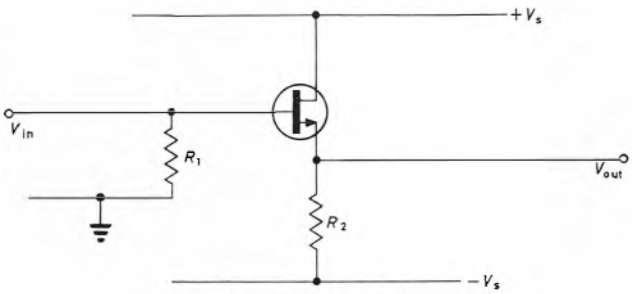


Fig. 1. Source follower

This article describes and analyses a two stage field effect transistor amplifier with input resistance in excess of 100MΩ and then discusses and analyses a simple modification to the circuit which enables the input resistance to be set to infinity. Both circuits have unity voltage gain and are suitable for use in sample and store circuits where the amplifier may be required to measure small voltages stored on a capacitor without discharging the capacitor. The advantages of the two stage amplifier over the single stage source follower as shown in Fig. 1 are: more linear operation, lower output impedance, less susceptibility to supply voltage variations and output voltage is equal to input voltage (with no offset proportional to the source gate voltage of the source follower circuit). The advantages of the two stage amplifier are specified but not proven since the purpose of this article is to present and analyse two circuits for a very high input impedance amplifier.

Finite Input Resistance Circuit

Fig. 2 shows the circuit of a two stage feedback amplifier of unity gain and capable of input resistance greater than 200MΩ. This circuit is shown a.c. coupled with capacitor C but is also suitable as a d.c. amplifier with C short-circuited. The circuit comprises a field transistor VT_1 to provide the high input resistance. The second transistor VT_2 is a conventional pnp transistor. The 5·6V Zener diode MRZ_1 provides reference voltage for the base of VT_2. The temperature coefficient of the Zener diode tends to compensate for the negative temperature coefficient of VT_2.

The operation of the circuit is as follows. The input voltage is applied to the gate of VT_1 creating a change in the source current through R_2 and R_3. This current is amplified in VT_2 and fed back to the junction of R_2 and R_3 in phase with the input voltage change thus tending to maintain the voltage across R_1 constant. Potentiometer R_2 sets the bias conditions in the field effect transistor so that for 0 volts input at the gate, the output voltage is also 0 volts and nominally zero current flows in R_1.

The above statement neglects the leakage current in the gate circuit of VT_1. By suitable choice of transistor a gate leakage current as low as 10^{-12}A can be obtained and the assumption of negligible leakage current is then valid. Such small leakage currents can be obtained with metal oxide semiconductor transistors.

In the circuit shown in Fig. 2 an input resistance of approximately 500MΩ was obtained when R_1 was approximately 3·3MΩ.

Analysis of the Finite Input Resistance Circuit

Neglecting the leakage current in the gate circuit of VT_1 the equivalent circuit can be expressed as shown in Fig. 3 where the symbols are as designated in the List of Symbols.

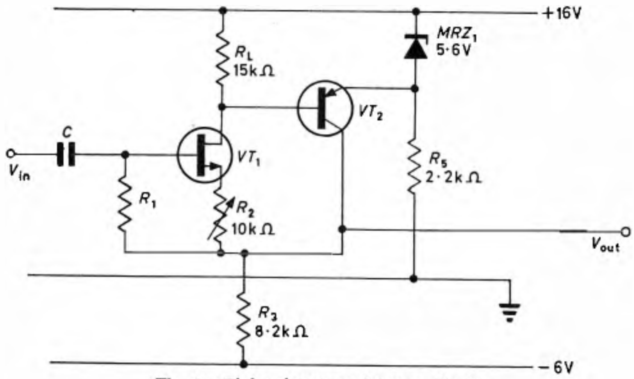
The analysis may be simplified by the following assumptions:

(1) h_{12} is of the order of $5 \cdot 10^{-4}$ hence $h_{12}e_o$ may be neglected.

Symbols

C	=	coupling capacitor
e_g	=	source to gate voltage of VT_1
e_i	=	input voltage to amplifier
e_o	=	output voltage from amplifier
g	=	mutual conductance of VT_1
h_{12}	=	reverse voltage amplification factor of VT_2
h_{21}	=	forward current amplification factor of VT_2
i_b	=	base current of VT_2
i_1	=	input current to amplifier
i_3	=	drain current in VT_1 ($= ge_g$)
R_E	=	additional resistance in emitter circuit of VT_2
R_I	=	input resistance of amplifier
R_L	=	drain load resistance of VT_1
R_o	=	output resistance of VT_1
R_1	=	circuit resistance at gate of VT_1
R_2	=	circuit resistance at source of VT_1
R_3	=	circuit output resistance
R_4	=	circuit feedback resistance
R_5	=	circuit resistance at anode of MRZ_1
R_{11}	=	input resistance of VT_2 with output short-circuited
R_{22}	=	output resistance of VT_2 with input open-circuit

* *University of Oxford.*



Fi. 2. Finite input resistance circuit

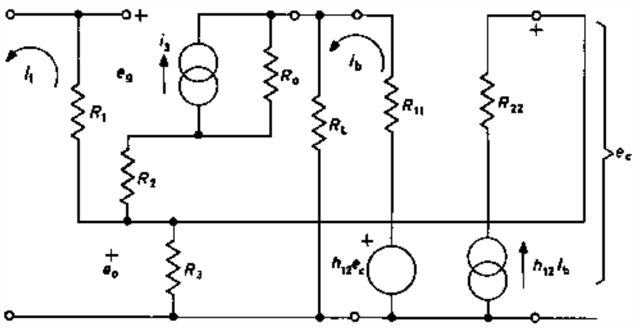


Fig. 3. Equivalent circuit of Fig. 2

(2) R_{11} is generally much smaller than R_L and hence R_L may be neglected.

(3) The sum of R_{11}, R_2 and R_3 in series is generally much less than R_o and hence R_o may be neglected.

The equivalent circuit thus simplifies to that shown in Fig. 4.

The equations for the above circuit are:

$$i_1 = \frac{e_1 - e_o}{R_1} \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (1)$$

$$i_2 + i_3 = h_{21} i_b = -h_{21} i_3$$

therefore:

$$i_2 = -(h_{21} + 1) i_3 \quad \ldots\ldots\ldots\ldots\ldots \quad (2)$$

as $h_{21} \gg 1$ equation (2) can be reduced to:

$$i_2 \simeq -h_{21} i_3 \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (3)$$

$$i_3 = g \,.\, e_g$$

$$e_g = i_1 R_1 - i_3 R_2$$

therefore:

$$i_3 = g i_1 R_1 - g\,(i_3 R_2)$$

or:

$$i_3 = \frac{gR_1}{1 + gR_2} \,.\, i_1 \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (4)$$

also:

$$e_o = (i_1 - i_2)\, R_3 \quad \ldots\ldots\ldots\ldots\ldots \quad (5)$$

from equations (3) and (4):

$$i_2 = \frac{-h_{21} g R_1 i_1}{1 + gR_2}$$

and from equation (1):

$$i_2 = \frac{-h_{21} g\,(e_o - e_1)}{1 + gR_2}$$

Hence equation (5) becomes:

$$e_o = (e_1 - e_o)\left[1/R_1 + \frac{h_{21} g}{1 + gR_2}\right] R_3$$

$$= \frac{e_1 \left[\dfrac{1 + gR_2 + R_1 h_{21} g}{R_1\,(1 + gR_2)}\right] R_3}{1 + \dfrac{(1 + gR_2 + R_1 h_{21} g)\, R_3}{R_1(1 + gR_2)}}$$

$$e_o/e_1 = \frac{(1 + gR_2 + R_1 h_{21} g)\, R_3}{R_1(1 + gR_2) + (1 + gR_2 + h_{21} R_1)\, R_3}$$

The input resistance R_I may be determined as follows:

$$R_I = e_1/i_1 = e_1 \Big/ \left(\frac{e_1 - e_o}{R_1}\right) = \frac{e_1 R_1}{e_1 - e_o}$$

$$R_I = \frac{R_1}{1 - e_o/e_1}$$

i.e.

$$R_I = R_1 \Big/ \left[\frac{R_1\,(1 + gR_2)}{R_1\,(1 + gR_2) + (1 + gR_2 + h_{21} R_1)\, R_3}\right]$$

$$= R_1 + R_3 + R_1 \frac{h_{21} g R_3}{1 + gR_3} \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (6)$$

This result shows that the input resistance is always finite and positive.

If $gR_2 \gg 1$ then:

$$R_I = R_1 + R_3 + R_1 h_{21}\,(R_3/R_2) \quad \ldots\ldots\ldots \quad (7)$$

which is independent of mutual conductance (g) of the field effect transistor VT_1.

Infinite Input Resistance Circuit

The circuit already discussed is adequate for most high input resistance circuits. By suitable choice of R_1 input resistance greater than 1 000MΩ can be obtained. Should a higher input resistance be required a simple modification to the circuit of Fig. 2 enables the input resistance to be adjusted through infinity.

This modification is shown in Fig. 5 where potentiometer R_4 is added between the source of VT_1 and earth. The action of this circuit is approximately as follows.

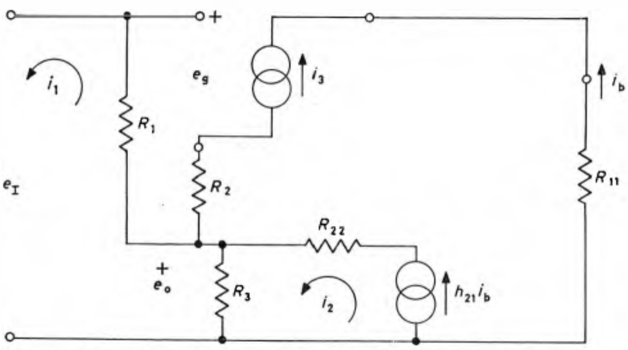
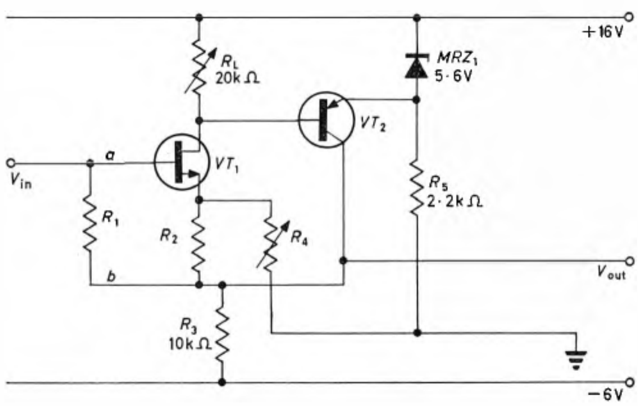


Fig. 4 (above). Simplified equivalent circuit of Fig. 2

Fig. 5 (below). Infinite input resistance circuit



For infinite input resistance point b must follow exactly the potential at point a. This requires the current in VT_1 to increase as the gate voltage increases. In order for more current to pass through VT_1 as the voltage input goes positive, the gate to source voltage of VT_1 must reduce, i.e. voltage across R_2 must reduce. The value of R_4 is set so that the current increase therein less the current fall in R_3 as the input goes positive, provides the required net increase of current in VT_1 for feedback through VT_2 to permit point b to follow exactly point a.

The input resistance of this circuit is set to the desired value by adjusting R_4. The quiescent current through VT_1 is set by adjusting R_L so that the potentials at a and b are equal and no current flows in R_1. R_L is preferable to R_2 for adjusting the quiescent conditions because the input resistance of the circuit is dependent upon the value of R_2 and substantially independent of the value of R_L.

As in the previous circuit this description neglects the gate to source leakage current of the field effect transistor.

Analysis of Infinite Input Resistance Circuit

Neglecting gate leakage current in VT_1 the equivalent circuit may be as shown in Fig. 6.

As the assumptions made in the previous analysis are still valid the equivalent circuit may be simplified to Fig. 7.

The equations for the circuit of Fig. 7 are:

$$i_1 = \frac{e_1 - e_o}{R_1} \quad \ldots\ldots\ldots\ldots (8)$$

$$e_o = i_1R_3 - i_2R_3 \quad \ldots\ldots\ldots\ldots (9)$$

$$h_{21}i_b = i_2 - i_4 \quad \ldots\ldots\ldots\ldots (10)$$

$$i_b = -i_3 = -g\,e_g$$

$$e_g = i_4R_2 + i_1R_1$$

$$i_b = -g\,(i_4R_2 + i_1R_1) \quad \ldots\ldots\ldots\ldots (11)$$

also:

$$\begin{aligned} e_o &= i_4(R_2 + R_4) + ge_gR_4 \\ &= i_4(R_2 + R_4) + R_4g\,(i_4R_2 + i_1R_1) \\ &= i_1(gR_1R_4) + i_4(R_2 + R_4 + gR_2R_4) \quad \ldots\ldots (12) \end{aligned}$$

Now substituting for equations (10) and (11) in equation (12),

$$\begin{aligned} e_o &= i_1R_3 - R_3(h_{21}i_b + i_4) \\ &= i_1R_3 - R_3(i_4 - h_{21}g\,(i_4R_2 + i_1R_1)) \\ &= i_1(R_3 + h_{21}gR_1R_3) + i_4(h_{21}gR_2R_3 - R_3) \quad \ldots\ldots (13) \end{aligned}$$

From equation (13) i_4 may be written:

$$i_4 = [e_o - i_1(R_3 + h_{21}gR_1R_3)]/h_{21}gR_2R_3 - R_3$$

or:

$$i_4 = e_o\frac{(R_1 + R_3 + h_{21}gR_1R_3) - e_1(R_3 + h_{31}gR_1R_3)}{R_1(h_{21}gR_2R_3 - R_3)} \quad .. (14)$$

Substituting equations (8) and (14) in equation (5):

$$e_o = gR_4(e_1 - e_o) + \left[\frac{R_2 + R_4 + gR_2R_4}{R_1(h_{21}gR_2R_3 - R_3)}\right]$$
$$e_o(R_1 + R_3 + h_{21}gR_1R_3) - e_1(R_3 + h_{21}gR_1R_3)$$

Collecting terms this becomes:

$$e_o/e_1 = \frac{gR_1R_4(h_{21}gR_2R_3 - R_3) - (R_2 + R_4 + R_2R_4g)(R_3 + h_{21}gR_1R_3)}{R_1(h_{31}gR_2R_3 - R_3) + gR_1R_4(h_{21}gR_2R_3 - R_3) - (R_2 + R_4 + gR_2R_4)(R_1 + R_3 + h_{21}gR_1R_3)} \quad \ldots\ldots (15)$$

For infinite input resistance $e_1 = e_o$

Hence $1 - (e_o/e_1) = 0$.

$$\text{But } 1 - (e_o/e_1) = \frac{R_1(h_{21}gR_2R_3 - R_3) - R_1(R_2 + R_4 + gR_2R_4)}{\Delta}$$

where Δ denotes the denominator of equation (15).

Hence the input resistance $R_I = \infty$ when:

$$h_{21}gR_2R_3 - R_3 = R_2 + R_4 + gR_2R_4$$

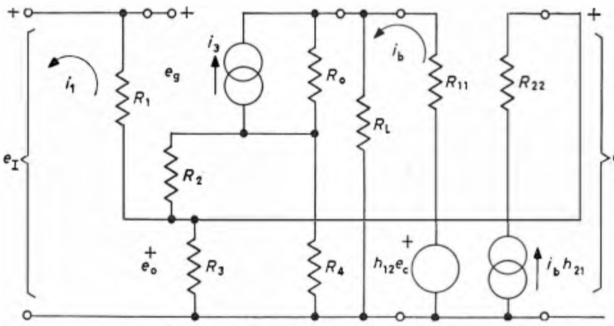


Fig. 6. Equivalent circuit of Fig. 5

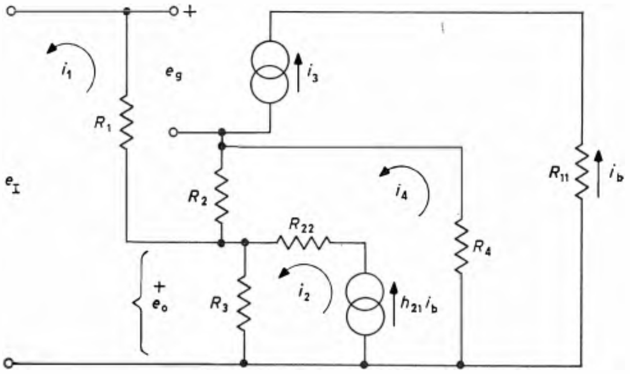


Fig. 7. Simplified equivalent circuit of Fig. 5

Now in the practical circuit $R_4 \gg R_2$ hence

$$R_I \to \infty \text{ when } h_{21}gR_2R_3 = R_3 + R_4 + gR_2R_4 \quad \ldots\ldots (16)$$

Conclusions

The infinite input resistance amplifier of Fig. 5 is most suitable in a sample and store circuit where small voltages stored on a capacitor are to be read without discharging the capacitor. Although the circuit of Fig. 5 can be adjusted to infinite input resistance the temperature dependence of the gate to source voltage of the field effect transistor creates a resultant current in R_1 as the temperature varies, and this current may charge or discharge the capacitor being measured. In such applications where a d.c. connected input circuit is used it is preferable to make R_1 open-circuit so that any small change in output voltage with temperature does not cause a current flow through R_1 and thus change the steady state input resistance of the amplifier.

Equation (16) relating the parameters of the circuit of Fig. 5 for infinite input resistance includes h_{21} the current gain of transistor VT_2 and g the mutual conductance of the field effect transistor. As both of these parameters are temperature dependent the circuit can only be adjusted to infinity at one temperature. However, for temperature independent input resistance h_{21} can be replaced by inserting a resistance R_E say in the emitter of VT_2. h_{21} is then replaced by the ratio R_L/R_E in equation (16). Furthermore, if $gR_2R_4 \gg R_3 + R_4$ the equation reduces to:

$$h_{21}R_3 \simeq R_4$$

or $(R_L/R_E)R_3 \simeq R_4$ which is independent of transistor parameters.

For a.c. coupled high input impedance amplifiers the circuit configuration of Fig. 2 is preferred with the value of R_1 chosen to give the desired input resistance. The input resistance of this circuit is finite and positive with stable quiescent voltage input set by R_2. As with the circuit of Fig. 5 the input resistance (equation (6)) of this circuit is dependent on h_{21} and g. Here again if temperature independent input is required, R_E may be placed in the emitter of VT_2, thus h_{21} becomes R_L/R_E and by making $gR_3 \gg 1$ equation (7) reduces to:

$$R_I \simeq R_1 + R_3 + R_1 ((R_L/R_E) + 1) R_3/R_2$$
$$\simeq R_1 [(R_L/R_E) + 1] R_3/R_2$$

The disadvantage of these modifications is that the input resistance is reduced.

Acknowledgments

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A Stable 10kV Reversible Power Supply

By M. C. Jackson*

This safe ±10kV power supply has solid state circuits, overload protection, and conforms to the Harwell 2000 series pattern. The unit was originally designed to provide the acceleration and focusing potential for a mass spectrometer ion source. Stability of the output voltage is better than 1 part in 10^4.

(Voir page 489 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 496)

THE supply was originally designed to provide the acceleration and focusing potentials for a mass spectrometer ion source. The requirements are as follows:

(1) Voltage ±10kV with a fixed 20V step adjustment.
(2) Stability, 1 part in 10^4 per hour.
(3) Ripple better than 1 part in 10^4.
(4) Output current greater than 100μA.
(5) Stored energy less than 1 joule.
(6) Output power less than 5W.
(7) Auxiliary supply isolated at (1) to provide 4mA at 200V.
(8) The unit to be compatible with the Harwell 2000 series[1].
(9) Transistor circuits to be used where possible.

Items (5) and (6) relate to the local safety regulations and allow the equipment to be used without screens and locks.

Item (7) is required because the ion source may take current between the first two electrodes of the lens.

A triple width unit size was chosen approximately 10¼in by 6in by 12in deep, and since the unit system permits easy connexion of external power supplies, a commercial −30V, 1A supply was chosen to power the high voltage unit.

The output capacitance and resistance are limited to 0·02μF and 20MΩ respectively by requirements (5) and (6). The external current is made 250μA maximum, allowing an internal resistance of 40MΩ for stabilization feedback.

In order to make full use of solid state circuits, a chopper amplifier system is used, and the a.c. produced is amplified and fed via a power output stage to a transformer and cascade rectifier. The chopping frequency is chosen at 16kc/s, which gives good response speed and reduces the smoothing problem. The cascade rectifier allows a low transformer ratio which avoids secondary resonance difficulties and eases the high voltage insulation requirement.

Circuit Design

Chopper and Amplifier

The 16kc/s oscillator ($VT_{10}VT_{11}$), with its output squared by a Zener diode, provides the drive for the balanced chopper (VT_1VT_2). Two silicon transistors (2S005) are used for this and initial selection is required to find a pair giving 'low zero' output at the 'no signal' conditions. The chopped output is amplified via VT_3VT_7 and drives the power stage (VT_8VT_9).

Reference Unit

The high voltage chosen gives the advantages of reduced loop gain requirement and chopper zero stability. The chosen value of 33V is below the V_{cb} of the 2S005 transistors which cannot therefore be damaged by external overload conditions. The five diodes used have a temperature coefficient of 5 in 10^5/°C and are immersed in an oil bath so that the effects of rapid external temperature change are slowed down, but a controlled oven could be used with advantage. The current supply is regulated via $VT_{16}VT_{17}$ which gives a high effective series resistance with low power dissipation. The reference voltage changes by less than 1mV for a 5 per cent mains change. If a larger input change has to be accommodated, the smoothing capacitor can be replaced by a Zener diode (IS 5082) as shown in the circuit diagram.

Cascade Generator

A ten stage voltage multiplier is used which is designed mechanically for easy reversal, so giving positive or negative output. To obtain the nominal voltage multiplication, $\omega CR = 13{\cdot}23n$ where n = number of stages, R = load resistance, $\omega = 2\pi \times$ drive frequency.

This gives C, the output capacitance across the generator, as 66pF, hence each capacitor should be a minimum of 330pF. The units chosen were 1 000pF since these were of a convenient form from the mechanical standpoint.

The power efficiency of this generator is about 80 per cent. The efficiency of the supply reaches approximately 60 per cent without the ancillary circuits such as overload protection etc.

The output ripple is given by $V_r = I_m/fC$ volts peak-to-peak, where I_m = mean current f = drive frequency and C = output capacitance of generator. This gives V_r = 156V and a two stage filter is used to reduce this below half a volt. It is important to screen the filter thoroughly as pick-up can give rise to several volts of ripple appearing at the output.

Protection Circuits

Two separate circuits are used which give protection for both short-circuit and over voltage conditions. The former frequently occurs in ion source work and the latter can occur due to a fault inside the unit, particularly if the

* *United Kingdom Atomic Energy Authority.*

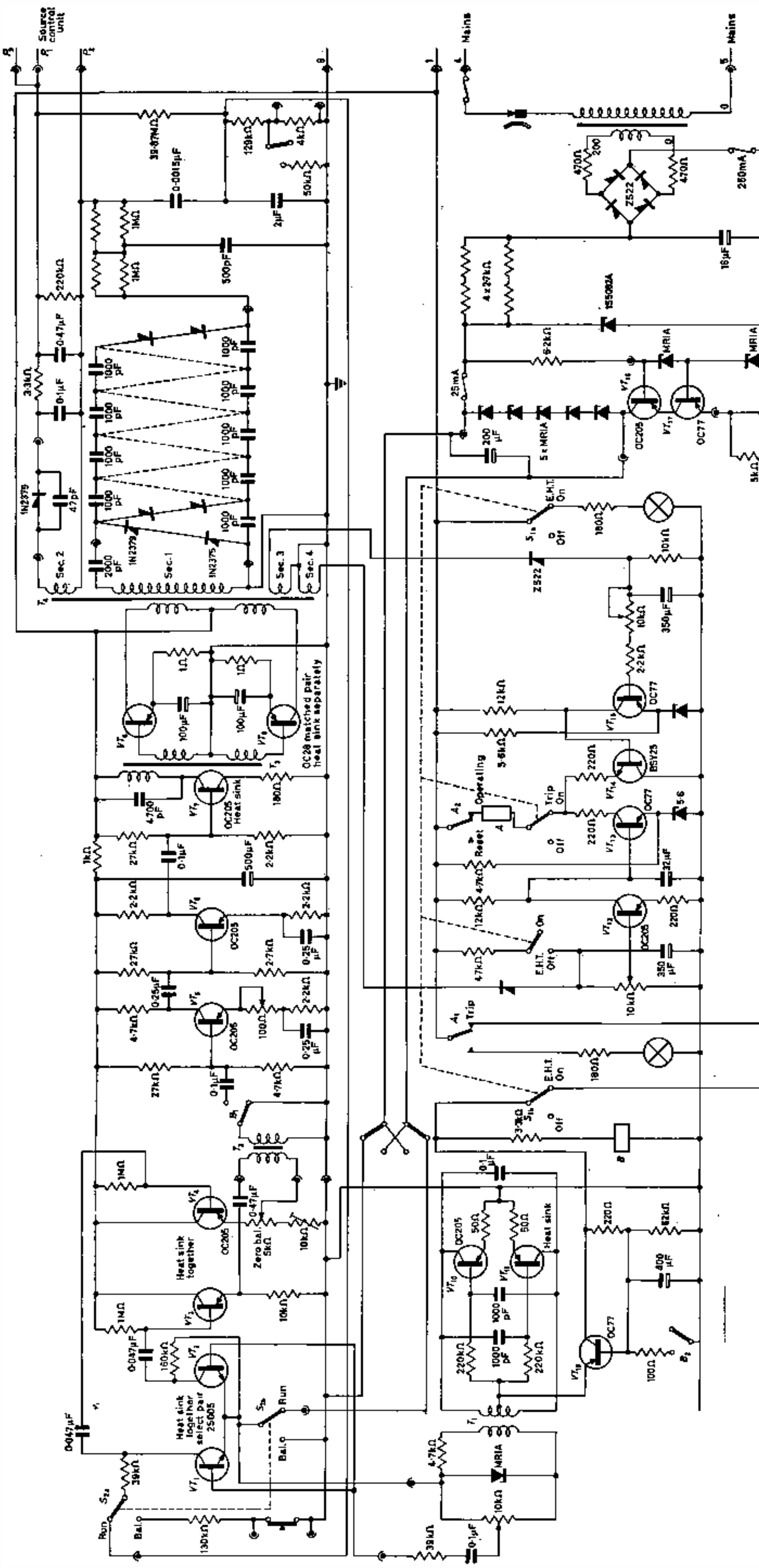

Fig. 1. The complete circuit

reference supply fails. The secondary windings Sec. 1 and Sec. 2, on the drive transformer, provide bias voltages and the short-circuit trip ($VT_{12}VT_{13}$) has its speed of operation controlled by varying the time-constant provided at the base of VT_{12}. This allows a number of spark discharges to occur before the supply switches off. The high voltage side ($VT_{14}VT_{15}$) fires when the input voltage rises above a pre-set value.

Test Results

In use the stabilizer will be run for long periods at a time into a constant load. The results below were obtained under these conditions, starting from cold.

Positive Output:

1st hour 4 parts in 10^4
2nd hour 2 parts in 10^4
3rd hour 5 parts in 10^5

Ripple Output:

400mV peak-to-peak.

Conclusions

The supply is compact, simple and gives adequate performance. Improvements could be made by use of more recently developed components. Reference diodes are now available with lower temperature coefficients and packaged choppers would ease selection difficulties.

Acknowledgments

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Relay Switching, Analogue Logic Circuits

By C. D. Lewis*, B.Sc., M.Sc.

A series of analogue computer circuits involving one or two operational amplifiers are described which will simulate the binary functions NOT, AND, OR, XOR, IF . . . .THEN, EQUIVALENT, NOR, NAND *and* BUT . . . . NOT.

(Voir page 489 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 496)

OFTEN when programming an analogue computer, one requires logic circuits to perform switching operations when certain conditions within the problem being considered are satisfied. With more modern analogue computers, logical switching can be done by means of discriminators and the binary logic circuits made available on such equipment. The majority of analogue installations, however, are not provided with these facilities and this article describes analogue circuits which will perform relay switching to the basic logical decisions such as NOT, AND, OR, XOR, IF . . . . THEN, EQUIVALENT, NOR, NAND and BUT . . . . NOT.

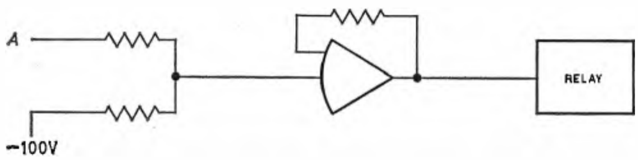


Fig. 1. Analogue NOT circuit
(All resistors 1MΩ unless otherwise stated)

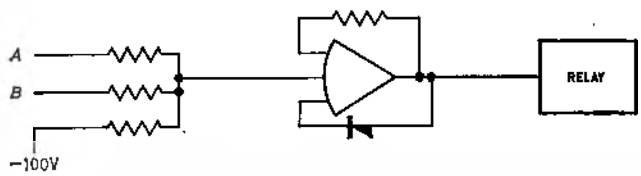


Fig. 2. Analogue AND circuit

Two State Analogue Logic

If the two states of an input parameter to an analogue logic circuit are, say, 0V and +100V it is apparent that with inversion through an amplifier, this binary state will soon became a ternary state with 0V, +100V and −100V representing the three states. Fortunately, however, the relay is a bipolar device and responds similarly to either +100V or −100V so that although three analogue states exist, these are interpreted by the relay as only two states; on and off.

In the circuits described here those that operate in a true binary fashion, in so far that both the input voltages and output voltages are only either 0V or +100V, are marked (*Tr*).

Analogue NOT Circuit (*Tr*)

	A	$\bar{A}$
Binary Truth Table		
	0	1
	1	0

In this circuit (Fig. 1) when A is +100V, the resultant input voltage to the amplifier, programmed as a summer is 0V, hence the output is 0V and the relay does not operate. When A is 0V, the resultant input to the summer is −100V, hence the output is +100V and the relay operates. Because the output voltage is +100V this circuit can be considered as true binary with respect to A.

	A	$\bar{A}$	Relay operates
Analogue Truth Table			
	0V	+100V	Yes
	+100V	0V	No

Analogue AND Circuit

	A	B	$A\&B$
Binary Truth Table			
	0	0	0
	0	1	0
	1	0	0
	1	1	1

In this circuit (Fig. 2) the amplifier is programmed as a limited summer, whose output voltage can only take negative values. Hence, when both A and B are 0V, although the resultant input voltage is −100V, the output voltage is 0V due to the feedback diode, and the relay does not operate. When A is +100V and B 0V or B 100V and A 0V, the resultant input voltage is 0V and again the relay does not operate. When both A and B are +100V, the resultant input voltage is +100V, hence the output voltage is −100V and the relay operates. Because, in this final condition, the output voltage is −100V this circuit cannot be considered as true binary with respect to A and B.

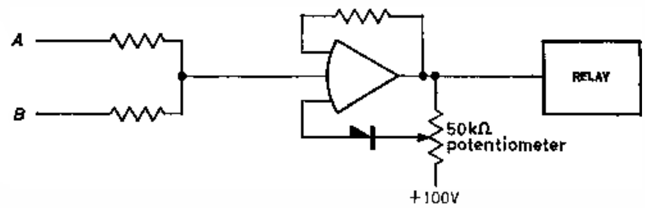


Fig. 3. Analogue OR circuit

	A	B	$A\&B$	Relay operates
Analogue Truth Table				
	0V	0V	0V	No
	0V	+100V	0V	No
	+100V	0V	0V	No
	+100V	+100V	−100V	Yes

Analogue OR Circuit

	A	B	$A \vee B$
Binary Truth Table			
	0	0	0
	0	1	1
	1	0	1
	1	1	1

In this circuit (Fig. 3), the amplifier is again programmed as a limited summer, in this case the output voltage when negative is limited to −100V. When both A and B are 0V the resultant input voltage is 0V and the relay does

* *University of Birmingham.*

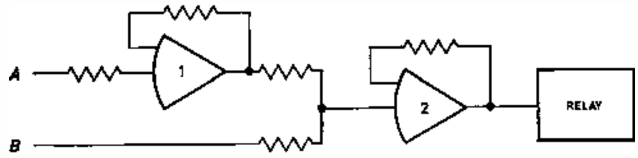


Fig. 4. Analogue XOR circuit

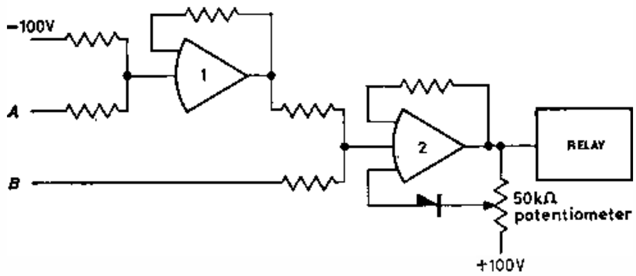


Fig. 5. Analogue IF . . . THEN circuit

not operate. For all other input conditions of A and B the output voltage is -100V and the relay operates.

Analogue Truth Table

A	B	$A \vee B$	Relay operates
0V	0V	0V	No
0V	+100V	−100V	Yes
+100V	0V	−100V	Yes
+100V	+100V	−100V	Yes

Analogue XOR Circuit

Binary Truth Table

A	B	$A \vee B$
0	0	0
0	1	1
1	0	1
1	1	0

The analogue XOR circuit (Fig. 4) requires two amplifiers, one programmed as a simple invertor (amplifier 1) and the other as a summer (amplifier 2). When both A and B are 0V or both +100V the resultant input voltage to the summer (amplifier 2) is 0V and the relay does not operate. When A is 0V and B +100V the resultant input to the summer is +100V and the relay is operated by the output voltage of -100V. When A is +100V and B 0V the resultant input voltage to the summer is -100V and the relay is operated by the output voltage of +100V.

Analogue Truth Table

A	B	$A \vee B$	Relay operates
0V	0V	0V	No
0V	+100V	−100V	Yes
+100V	0V	+100V	Yes
+100V	+100V	0V	No

Analogue IF . . . . THEN Circuit

Binary Truth Table

A	B	$A \rightarrow B$
0	0	1
0	1	1
1	0	0
1	1	1

The analogue IF . . . . THEN circuit (Fig. 5) also requires two amplifiers, both programmed as summers, the output voltage of amplifier 2 being limited at -100V. When both A and B are 0V, the output of amplifier 1 is +100V and hence the output voltage of amplifier 2 -100V and the relay operates. When A is 0V and B +100V the resultant input voltage to amplifier 2 is +200V and the output voltage hence -100V, due to the limiter, and again the relay operates. When A is +100V and B 0V the output voltage of amplifier 1 is 0V and the resultant input voltage to amplifier 2 therefore 0V and the relay does not operate. When both A and B are +100V the resultant input voltage to amplifier 2 is +100V and hence the relay is again operated by the output voltage of -100V.

Analogue Truth Table

A	B	$A \rightarrow B$	Relay operates
0V	0V	−100V	Yes
0V	+100V	−100V	Yes
0V	+100V	0V	No
+100V	+100V	−100V	Yes

Analogue EQUIVALENT Circuit

Binary Truth Table

A	B	$A \leftrightarrow B$
0	0	1
0	1	0
1	0	0
1	1	1

The analogue EQUIVALENT circuit (Fig. 6) consists of a single amplifier programmed as a summer. When A and B are both 0V the resultant input voltage is -100V and hence the relay is operated by the +100V output voltage. When A is 0V and B +100V or A +100V and B 0V the input voltage is 0V and the relay remains unoperated. When both A and B are +100V, the resultant input voltage is +100V and the relay is operated by the -100V output voltage.

Analogue Truth Table

A	B	$A \leftrightarrow B$	Relay operates
0V	0V	+100V	Yes
0V	+100V	0V	No
+100V	0V	0V	No
+100V	+100V	−100V	Yes

Analogue NOR Circuit (*Tr*)

Binary Truth Table

A	B	$\overline{A \vee B}$
0	0	1
0	1	0
1	0	0
1	1	0

The analogue NOR circuit (Fig. 7) is exactly the same as the EQUIVALENT circuit with the exception that a diode feedback prevents the output voltage of the amplifier from becoming negative. Hence for the condition where both A and B are +100V the output voltage is limited at 0V (would have been -100V without diode) and the relay remains un-operated.

Analogue Truth Table

A	B	$\overline{A \vee B}$	Relay operates
0V	0V	+100V	Yes
0V	+100V	0V	No
+100V	0V	0V	No
+100V	+100V	0V	No

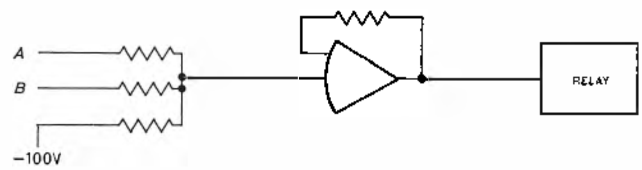
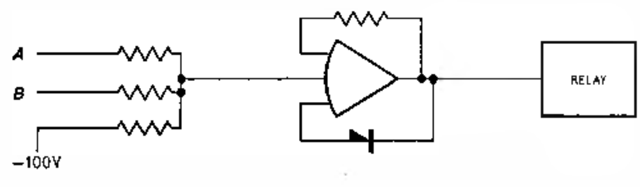


Fig. 6. Analogue EQUIVALENT circuit

Fig. 7. Analogue NOR circuit



Analogue NAND Circuit (*Tr*)

Binary Truth Table

A	B	$\overline{A \& B}$
0	0	1
0	1	1
1	0	1
1	1	0

The analogue NAND circuit (Fig. 8) is comprised of a single amplifier programmed as a summer with the output limited at +100V. When both A and B are 0V the input to the summer is equivalent to −200V (due to the 500kΩ input resistor) but because of the limiter the output voltage is +100V, and the relay operates. When A is 0V and B +100V or A +100V and B 0V the resultant input voltage is equivalent to −100V and hence the relay is operated again by an output voltage of +100V. When both A and B are +100V the resultant input voltage is equivalent to 0V and hence the relay is unoperated.

Analogue Truth Table

A	B	$\overline{A \& B}$	Relay operates
0V	0V	+100V	Yes
0V	+100V	+100V	Yes
+100V	0V	+100V	Yes
+100V	+100V	0V	No

Analogue BUT .... NOT Circuit (*Tr*)

Binary Truth Table

A	B	$\overline{A \to B}$
0	0	0
0	1	0
1	0	1
1	1	0

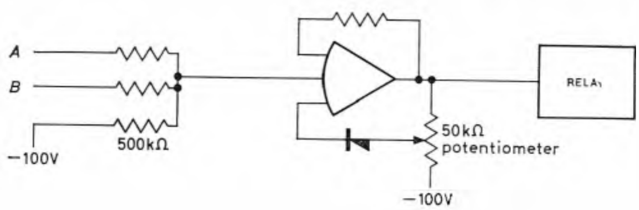


Fig. 8. Analogue NAND circuit

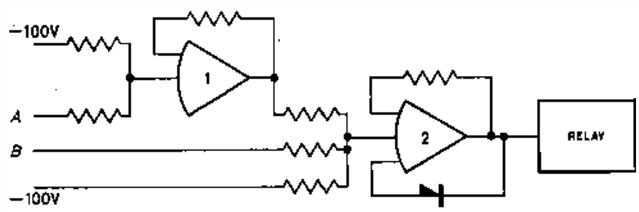


Fig. 9. Analogue BUT . . . NOT circuit

The analogue BUT .... NOT circuit (Fig. 9) is comprised of two amplifiers, the first of which acts as a summer and the second as a limited summer whose output voltage can only take positive values. When both A and B are 0V the output voltage of amplifier 1 is +100V, hence the resultant input voltage to amplifier 2 is 0V and the relay is unoperated. When A is 0V and B +100V the output voltage of amplifier 1 is again +100V so that the resultant input voltage to amplifier 2 is +100V, but due to the limiting diode the output voltage is 0V and the relay remains unoperated. When A is +100V and B 0V, the output voltage of amplifier 1 is 0V, so that the resultant input voltage to amplifier 2 is −100V and the relay is operated by the +100V output voltage. When both A and B are +100V the output voltage of amplifier 1 is again 0V and the resultant input voltage to amplifier 2 0V, hence the relay does not operate.

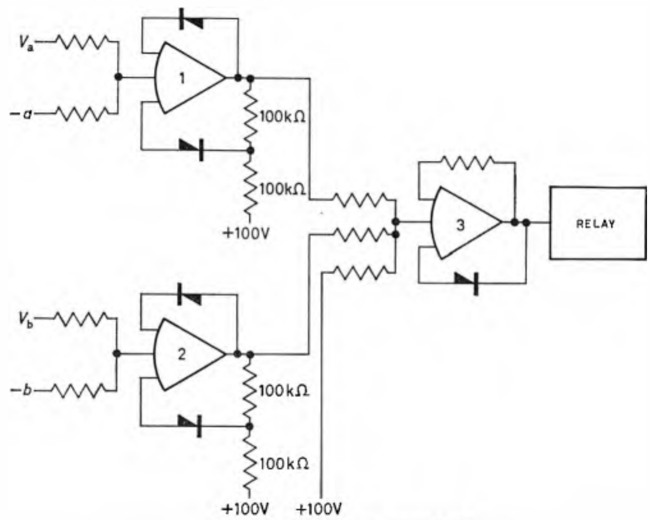


Fig. 10. Circuit for example 1

Analogue Truth Table

A	B	$\overline{A \to B}$	Relay operates
0V	0V	0V	No
0V	+100V	0V	No
+100V	0V	+100V	Yes
+100V	+100V	0V	No

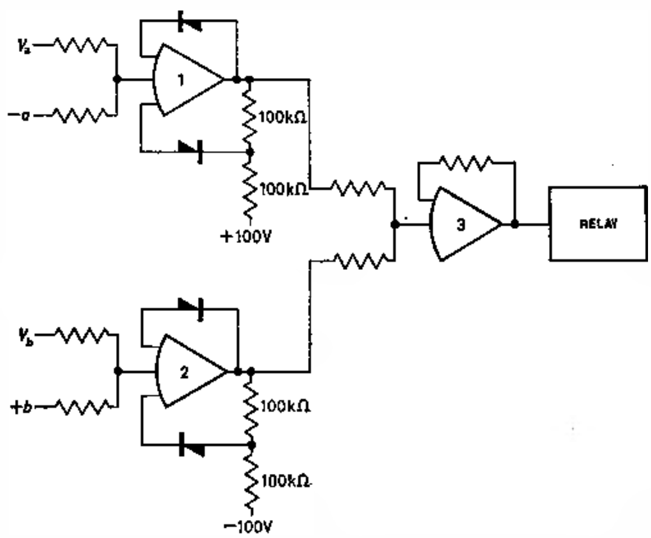
Example 1

In an analogue simulation it is required to operate a relay when two voltages V_a and V_b exceed $+a$ and $+b$ respectively, the relay to be operated only under this condition. The AND circuit (amplifier 3) and the two discriminating amplifiers (amplifiers 1 and 2) for this circuit are shown in Fig. 10. In this circuit because the two discriminators (amplifiers 1 and 2) have an output voltage of −100V when $V_a > +a$ and $V_b > +b$ the polarity of the diode feedback and the fixed input voltage are reversed in the AND circuit (amplifier 3) to that in Fig. 2.

Analogue Truth Table

V_a	V_b	Output amplifier 1	Output amplifier 2	Relay operates
$V_a < +a$	$V_b < +b$	0V	0V	No
$V_a < +a$	$V_b > +b$	0V	−100V	No
$V_a > +a$	$V_b < +b$	−100V	0V	No
$V_a > +a$	$V_b > +b$	−100V	−100V	Yes

Fig. 11. Circuit for example 2



Example 2

In an analogue simulation it is required to operate a relay only when a positive voltage V_a exceeds $+a$ and a negative voltage $-V_b$ does not exceed $-b$, or when the positive voltage V_a does not exceed $+a$ but the negative voltage $-V_b$ does exceed $-b$. The XOR circuit (amplifier 3) and the two discriminators for this circuit are shown in Fig. 11. Because the polarities of the two output voltages from the discriminators are different (output of amplifier 1 is $-100V$ when $V_a > +a$ and output of amplifier 2 is $+100V$ when $-V_b > -b$) when the restrictions are satisfied, the XOR circuit requiring two operational amplifiers, Fig. 4, can be replaced by a single amplifier circuit, amplifier 3.

Analogue Truth Table

V_a	V_b	Output amplifier 1	Output amplifier 2	Relay operates
$V_a < +a$	$-V_b < -b$	0V	0V	No
$V_a < +a$	$-V_b > -b$	0V	+100V	Yes
$V_a > +a$	$-V_b < -b$	−100V	0V	Yes
$V_a > +a$	$-V_b > -b$	−100V	+100V	No

Conclusion

Although it is acknowledged that logical switching is the prerogative of binary logic circuits, in analogue computing when such circuits are not available those described in this article will be a useful alternative.

A New Radio Controlled Tractor

A remote controlled tractor capable of working under conditions too dangerous for human beings is the end product of ten years' intensive development by the Ford Tractor Company Ltd.

Ford have provided the tractor, a Ford 3000 with Select-O-Speed transmission (Ford's 'change-on-the-go' transmission), and Fenlow Products Ltd, in conjunction with C. & L. Developments Ltd, and Noble and Dowden Ltd, have provided the 'driver.' The 'driver' is a multi-channel, simultaneous radio-link between a human 'switchboard' operator, and the tractor.

The radio-controlled Ford 3000 is the fourth and by far the most sophisticated remote-control tractor yet produced. Whereas the earlier three, with standard transmission, were of little more than 'gimmick' value, the Mark IV is believed to possess genuine commercial potential.

For instance, the robot tractor, controlled by cable or radio, can work in areas contaminated by radio-activity, or in hazardous terrain without any risk to the operator. There may also be agricultural applications.

The radio controlled functions available are as follows: engine start and stop; steering; gear change (10 forward, 2 reverse speeds and park position (wheels locked)), inching pedal; throttle; brakes (differential left and right); implement raise and lower; external hydraulics service (3 positions); differential lock; power take off; lights—side and tail; light—head; horn; bonnet opening and operation of battery to swung-out position for accessibility to engine.

The transmitter is energized by 8 internal DEAC batteries with provision for charging from 110 to 240V 50c/s a.c. Alternatively 8 standard U2 dry batteries can be used and a two-pin plug is also fitted for connexion to an external 12V battery. A panel mounted voltmeter is supplied to show the state of charge of the batteries.

Gear selection is by 13 self-cancelling push-buttons giving 10 forward and 2 reverse ratios and park position.

Other controls are by centre biased or locking toggle switches with push-buttons for horn and starter controls. The latter is inter-locked with the park button to prevent starting the engine when in gear.

The receiver is carried on the tractor and is powered by the 12V tractor battery. A row of signal lamps housed under a hinged cover indicate if the transmitted channel is being received, and a row of associated push-buttons enable the secondary circuits and tractor functions to be checked, if required, without the use of the radio link. It is therefore possible to drive the tractor from these buttons for checking purposes. The receiver supply is fed from the tractor start key, isolating the feed if the key is in the 'off' position or removed. A further changeover switch on the tractor enables the engine start solenoid to be energized independently for manual driving.

The secondary circuits operate from output transistors or secondary relays to the load requirement, the final operation of the tractor controls being carried out by solenoid operated valves in the case of hydraulic functions such as steering, battery swing-out, and by electric motor actuators in the case of throttle, raise lower, external hydraulics and differential lock. The lights and horn are operated from secondary relays.

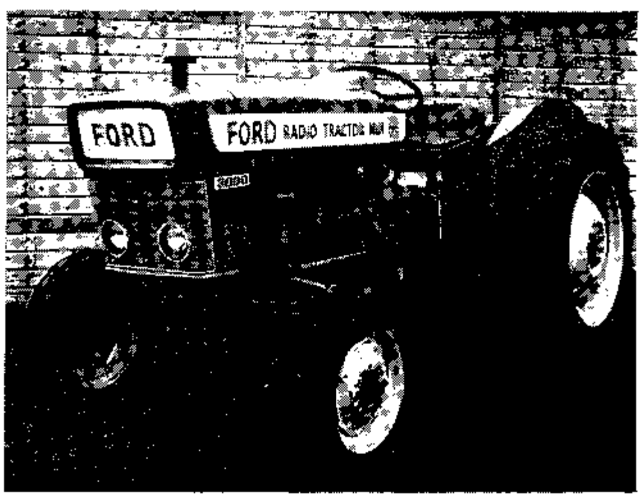
In order to give 'fail safe' operation the engine stop control (which interrupts the fuel supply of the diesel injector pump) is spring loaded to the stop position. A solenoid energized from the battery and controlled by the carrier signal holds this control in the 'on' position when operating the tractor. Thus failure of the transmitter, receiver, or local battery supply will cause the engine to stop by the action of the spring, and the transmission will assume the 'park' or 'locked' position. Similarly when the limiting range of the transmitter is reached the engine will stop.

It is necessary for each operator to negotiate with the GPO for frequency allocations to use a radio link, but two-wire direct links can be used as required, without any prior approval.

For certain very restricted range requirements where interference or screening problems might be severe, for example operation in the hold of a ship, a multi-core direct cable could be used.

All the controls are arranged so that they can be operated manually for driving the tractor in the normal manner. In addition to operating the radio manual switch a tap is provided to release the hydraulic fluid in the steering cylinders to enable manual steering to be performed.

The Ford mark IV radio controlled tractor



Integrated Electronics — Its Place in Education

By K. J. Dean*

In this article some recent developments in electronic engineering are discussed, together with their implications in education. Apparatus is described with which the applications of integrated circuits can be investigated in the laboratory.

(Voir page 489 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 496)

ELECTRONICS is a fast-moving and rapidly changing subject. The crystal triodes of 1948 seemed then no match for the Goliath of thermionics. In turn they have been replaced by silicon planar devices whose development has lead directly to monolithic silicon chip integrated circuits.

The circuits now available include linear amplifiers and logic circuits fabricated on silicon chips which are frequently less than 4mm² in area. These are produced by multiple diffusion into an epitaxial layer and by selective oxide evaporation. Resistors, capacitors, transistors and diodes are all produced in this way, but the circuit designer must also take into account parasitic elements between components. These are much more important than with circuits constructed with discrete components.

In addition, there is a second, and larger, form of integrated circuit, the thin fiim circuit. Here passive components are produced on a ceramic substrate by vacuum evaporation or by some similar method. Active components, such as flip-chip transistors, are subsequently added so that the final hybrid circuit is built on a substrate the area of which is of the order of 5cm².

The Developing Technology

In education one cannot afford to teach principles in such a way that the practice of electronics is unrelated to current industrial practice. For instance, thermionics is taught because it still has its place in high power systems and in the entertainment market. Transistor theory is taught because transistors are important wherever mobility is at a premium; in instrumentation, in computers, and in a host of other applications.

Now integrated electronics is showing the same growth pattern as was experienced years ago with transistors. First, integrated circuits were accepted for military applications, then as price fell, for industrial purposes, and finally with increased production, they will be used in the entertainment market.

One cannot expect students to support an educational establishment which provides them with chiefly obsolescent material. Hence, integrated circuits must be introduced in the courses. This is not to say that one must be blown about with every wind of change, but consolidated development must be reflected in college work.

As technology advances, so the ability to use the new developments improves also. Early transistor circuits imitated the valve circuits with which we were then so familiar. A few years passed before we were all alive to the implications and added freedom of the new techniques, such as complementary circuits and the use of pnpn diodes.

A similar period of change and development is now taking place with the design of integrated circuits. At first the early circuits had a great deal in common with the former discrete circuits. The next stage—the evolution of an integrated circuit design philosophy—is now proceeding. For example, the design of metal-oxide-semiconductor integrated circuits is at a stage where the functions of resistors and charge storage diodes is taken over by m.o.s. devices.

Practical Work in Basic Technology

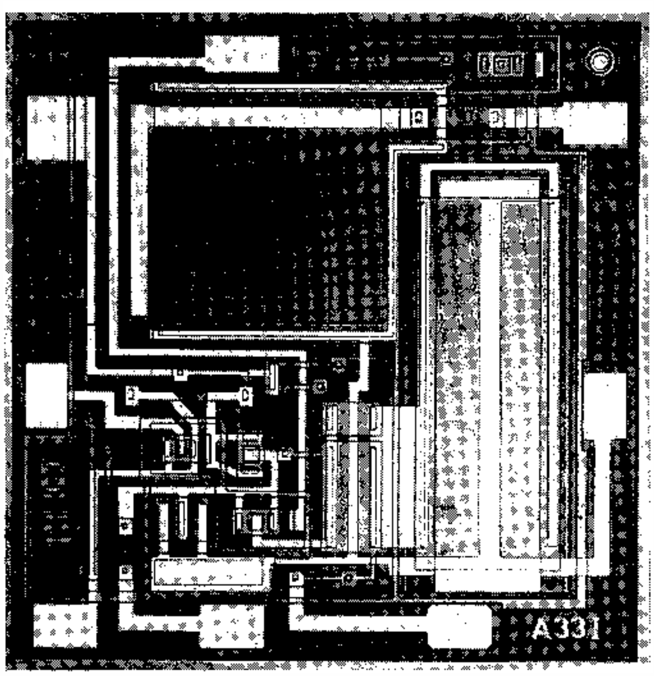
What part can all this play in undergraduate courses at Universities, and Higher National Diploma and Certificate courses in Colleges of Technology? The construction of monolithic circuits is virtually impossible, and even the construction of thin film circuits involves a large capital outlay. The majority of Colleges are thus restricted to the applications of integrated circuits. In basic technology students should be given the opportunity of examining silicon chips, perhaps of photographing them and of identifying individual components, to supplement theoretical lectures. It is of help if chips are available which show typical defects and failures, such as the effects of overheating and of purple plague. Fig. 1 shows a typical photograph of an integrated circuit made on a projection microscope. Similar results could probably be achieved with care using a less sophisticated instrument.

Fig. 2 shows the circuit diagram which corresponds to Fig. 1. It is drawn in this way because it was originally produced as an overlay for Fig. 1. The production of drawings of this kind gives students practice at identifying components on the chip.

Linear Amplifiers

There can surely be no course in electronics in which at some time the characteristics of an amplifier are not

Fig. 1. Projection microscope photograph of integrated circuit



* College of Technology, Letchworth.

examined. The input and output impedances, gain and frequency response are all of importance. Examination of these can take place just as easily with integrated circuits as with discrete amplifiers. The circuit shown in Fig. 3 is an example of a typical amplifier in the range produced by Plessey Components Ltd. It is useful in this connexion because it is not intended to be a complete amplifier but its characteristics can be modified by the use of external components.

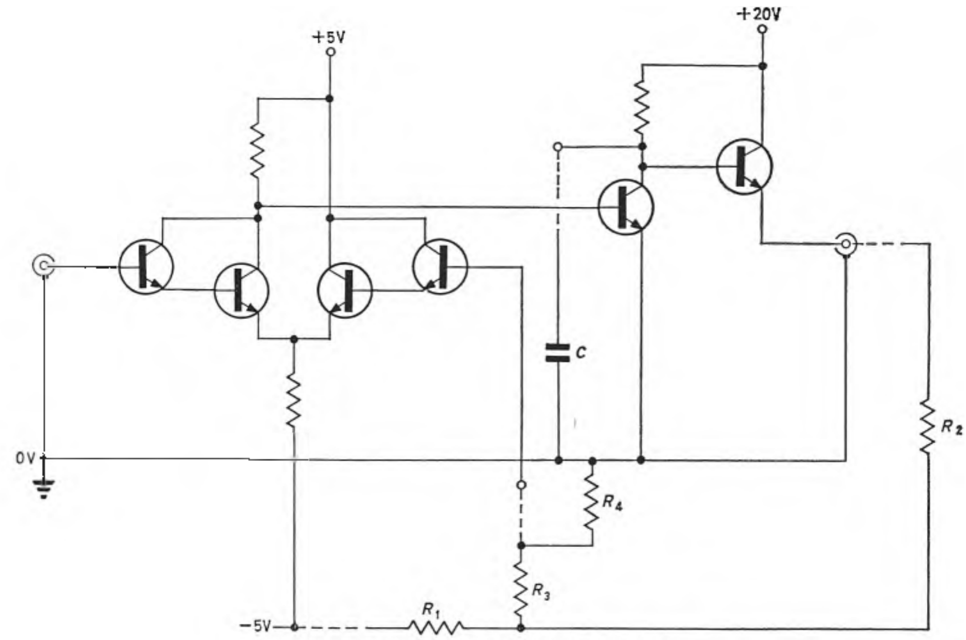


Fig. 3. Typical circuit of integrated amplifier

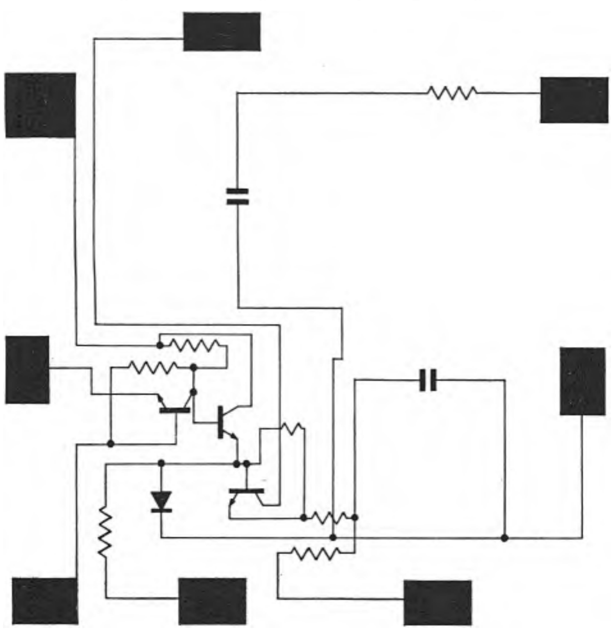
With this amplifier it is possible to alter its working point and the amount of directly coupled feedback used. Feedback networks may, in addition, be capacitively coupled so that the upper frequency response, in this case extending to about 3Mc/s, may be modified. The obvious technique here is to have the amplifier and any possible feedback components mounted on a 'breadboard'. However, care must be taken to ensure that the amplifier does not become unstable, either through common impedance in the power supply lines, or through operation at too high gain. Similar circuits can also be connected as crystal controlled oscillators, or oscillation can be sustained by selective feedback.

Thus, many of the demonstrations which are met with discrete amplifiers can be reproduced with highly reliable integrated circuits, which take up less storage space and which produce enthusiastic co-operation from students.

Logic Circuits

The widest use of integrated circuits is at present in the field of logical circuits and if they are to be included in College courses dealing with this subject they must be suitably mounted. This is because practical work with logical circuits involves the frequent rearrangement of basic 'building bricks' to form useful functions. For this reason it will be appreciated that unless they are 'student-proof' they will not be used at their present price.

Fig. 2. Circuit diagram corresponding to Fig. 1



One of the simplest pieces of equipment, intended as a teaching aid, is Lan Electronics' Lan-Dec 20*. This uses SGS-Fairchild industrial micrologic elements† and enables interconnexion to be made between a number of two-input and four-input gates. These are NOR gates in positive logic. The complete device includes lamp units, suitable power supplies and switches to set the input logic levels.

These SGS-Fairchild elements are examples of transistor resistor logic (t.r.l.), but since the Lan-Dec 20 only uses gates, and bistable elements are made by interconnexion, its use is in demonstrating Boolean algebra rather than in the development of complex counters‡ SGS-Fairchild also manufacture JK flip-flops and other more complex gates and to use these, some more comprehensive equipment is required. In addition, reasonably priced current mode logic (c.m.l.) elements are available from Motorola (marketed by Celdis Ltd, Reading) and slightly more expensive resistor capacitor transistor logic (r.c.t.l.) elements from Texas Instruments, Ltd, Bedford. Students should be given experience in a variety of logic systems such as those described. The equipment used for this purpose at Letchworth is illustrated in Figs. 4 and 6.

The logic tutor shown in Fig. 4 uses a patchboard (No. F00 31911) from Sealectro Ltd, Portsmouth, and includes four transistor amplifiers which operate the lamps at the top of the board. One of these amplifiers is shown in Fig. 5. Push-buttons are also provided to give the appropriate logic levels. Because this kind of circuit requires the '1' logic level to light a lamp, and the '0' level to extinguish it, each board is built to operate with one logic system. For example, the levels are 0·2V and 1·2V for the SGS-Fairchild industrial range and −0·75V and −1·55V for the Motorola elements, etc. A selection of

* *Electronic Engng. 38, 49 (Jan. 1966).*
† *Or Motorola equivalents.*
‡ *There is also an attachment using five JK flip-flops.*

these tutors is therefore required. The complete tutor (with external power supplies) can be assembled in the college for under £80.

Fig. 6 shows two of the boards which use Sealectro 'Press-Fit' terminals (types BAO 533 and BAO 557). These boards carry a small printed circuit to which the integrated circuit is either welded or soldered. These boards can be produced with quite simple printed circuit facilities, or could be manufactured in Veroboard. As many as 40 integrated circuits can be interconnected at one time on a single tutor.

One of the advantages of working with integrated circuits is that students can acquire experience in the use of JK flip-flops. When using the older RS and RST flip-flops, some combinations of logic levels must be forbidden, since they give rise to unpredictable output states. JK flip-flops are bistable elements which change state without ambiguity, whatever the states of their inputs. This kind of flip-flop is only readily available in integrated form.

Fig. 7 shows a typical counter which uses JK flip-flops and which can easily be assembled and tested using this tutor. It is only one of a large number of different types of counter which have been investigated with the tutor, some of which do not require extra gates. This is a decade counter whose output is weighted in 8421 b.c.d., and its ability of omitting six possible states to give a count of ten, depends on the special properties of JK flip-flops.

A more sophisticated apparatus which also uses Sealectroboard and also accepts the small printed circuit boards already described is manufactured by Texas Instruments Ltd, primarily for use with their own integrated circuits. Its features include a variable speed oscillator and DM 160 triode read-out which imposes negligible load on the logic circuits. This instrument was originally introduced for logical designers in industry. However, staff of some colleges may consider that a simpler arrangement, as described here, would be better suited to their needs.

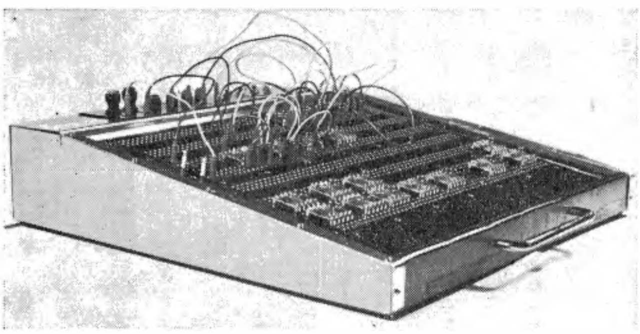
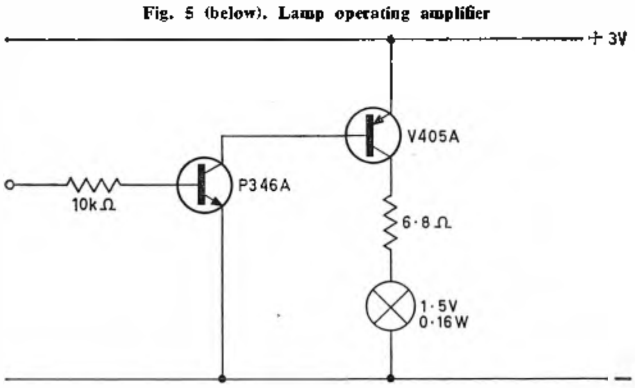


Fig. 4 (above). The logic tutor

Fig. 5 (below). Lamp operating amplifier



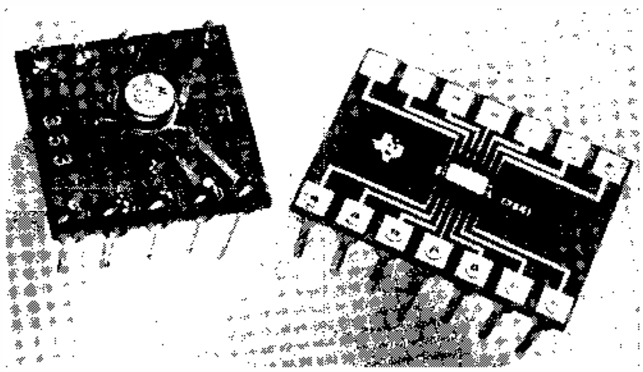


Fig. 6 (above). Two of the circuit boards

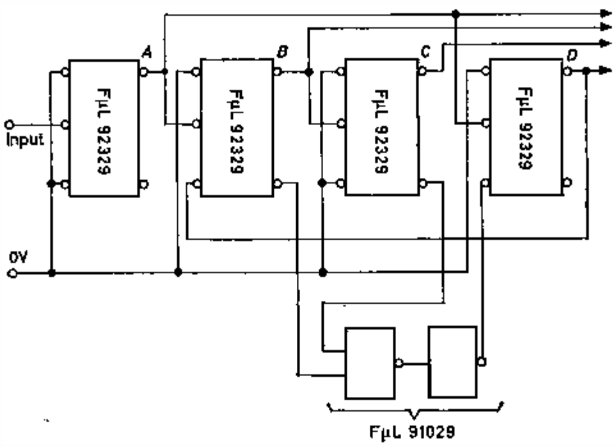


Fig. 7. Counter using JK-flip-flops

The Role of Integrated Electronics in Education

It would be perhaps premature to advocate completely 'integrated' electronics in our colleges. There is still a place, though a shrinking one, for thermionic valves and alloy junction transistors. These are used in vast numbers but there is a danger that if integrated electronics is neglected, our education system will lag too far behind industry to play its correct part in providing the technologists and technicians which industry needs.

Integrated amplifiers can be introduced in Higher National Certificate and Diploma courses and used to illustrate negative feedback, frequency response characteristics, and the other features of operational amplifiers.

A number of colleges, of which Letchworth is one, include logical circuits in the H.N.D. course in Electrical and Electronic Engineering or offer an endorsement to H.N.C., perhaps in Counting Techniques, Digital Methods, or Transistor Logic. Integrated circuits should now be playing a large part in such courses. Computer manufacturers are now using these elements, and the rest of instrumentation cannot be far behind. It is to be hoped that this survey will accelerate the involvement of education in integrated electronics.

Acknowledgment

The author acknowledges the help of members of staff of the Department of Science and Electrical Engineering in preparing this article and wishes to thank the Principal of the Letchworth College of Technology for permission to reproduce the accompanying photographs, and the Directors of Plessey Ltd. for permission to reproduce Fig. 1.

Rectification Using Devices with Symmetrical Characteristics

By J. F. Young, C.G.I.A., A.M.I.E.E., A.M.I.E.R.E.

It is shown that a device having a symmetrical voltage-current characteristic can act as a rectifier of pulse or even-harmonic-distorted waveforms. The characteristics of such a pulse symmetrical rectifier are derived and the dependence of the rectified output on mark-to-space ratio rather than on input pulse amplitude is shown. Zener diode pairs have been used as symmetrical rectifiers, though other devices have suitable characteristics. Some possible applications of the symmetrical rectifier are discussed.

(Voir page 489 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 496)

IT is usual to assume that in order to obtain rectification of an alternating current waveform a non-linear device having an asymmetrical current-voltage characteristic is required. This is quite true when the alternating current has a sinusoidal waveform or when it contains only odd harmonics of the fundamental frequency so that positive and negative half cycles of the waveform have the same shape. However, it is possible to obtain d.c. output from a device having a symmetrical current-voltage characteristic provided that the input voltage waveform is asymmetrical. An obvious form of waveform asymmetry is provided by some d.c. content in the input voltage[1]. It is not so obvious that even where there is no d.c. content in the input waveform, a d.c. output is obtainable from a device having a symmetrical current-voltage characteristic provided that waveform asymmetry due to harmonics is present. There seems to have been little application of this fact.

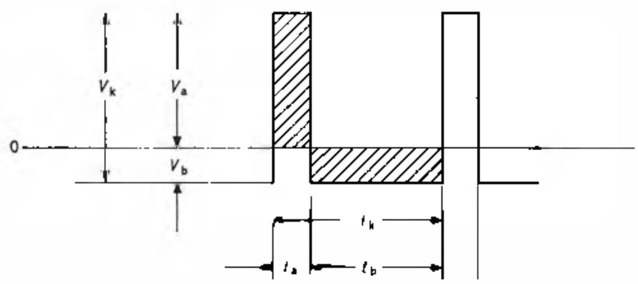


Fig. 1. Pulse waveform

The phenomenon is of particular interest in pulse applications. For example, in the pulse voltage waveform of Fig. 1 there is no d.c. content and the integral of the waveform over the period $t_k = t_a + t_b$ is zero. Consequently the shaded area $V_a \times t_a$ must be equal to the shaded area $V_b \times t_b$. If $t_a < t_b$ then $V_a > V_b$, and the peak positive value exceeds the peak negative value. If this voltage waveform is applied to a device having the symmetrical, though non-linear, current-voltage characteristic of Fig. 2, then the current flowing when the voltage is positive (V_a) will exceed that when the voltage is negative (V_b). In fact if V_b is less than the threshold value V_x, then no current will flow in the negative direction at all.

The sloping parts of the characteristic of Fig. 2 can be considered to be due to a pure ohmic resistor, and the non-linear part of the characteristic as formed from three straight lines at right-angles. The characteristic would be given by that part of the circuit of Fig. 3 which is to the right of the points M and N. If the input voltage V_1 has the waveform of Fig. 1 then a voltage V_2 appears across

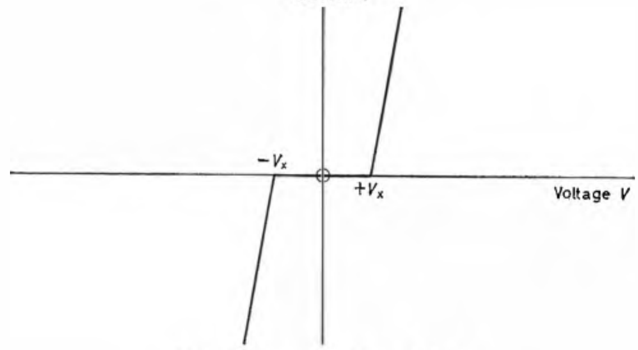


Fig. 2. Symmetrical characteristic

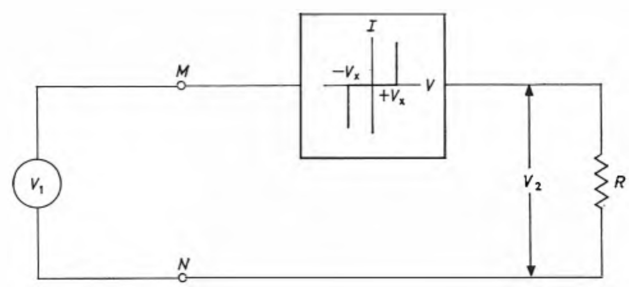


Fig. 3. Basic circuit

the resistor R only when the input voltage V_1 exceeds the value V_x, either in the positive or in the negative direction. Thus the voltage V_2 across the resistor R has the form $(V_1 - V_x)$ and the mean value of the voltage V_2 can be found by integration.

It is of interest to find how the mean value of the output voltage V_2 varies with the mark-to-space ratio t_a/t_b of the waveform, or with the ratio $t_a/(t_a + t_b) = t_a/t_k$, where t_k is the period of the waveform. It is assumed that the period t_k and the peak-to-peak amplitude $V_a + V_b = V_k$ are both fixed. There are two conditions which must be considered:

(1) V_b greater than V_x so that current flows in both positive and negative directions;

(2) V_b less than V_x so that current flows only in the positive direction.

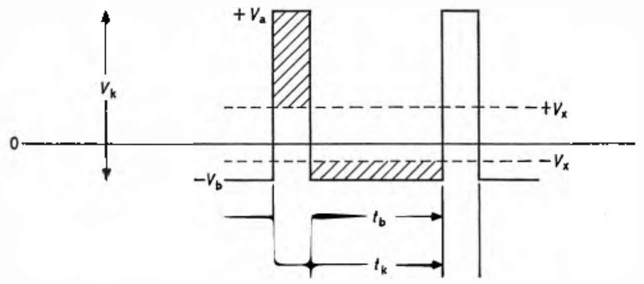


Fig. 4. Pulse waveform

In the first case, the positive shaded area in Fig. 4 is $(V_a - V_x)t_a$ volt-seconds, while the negative shaded area is $(V_b - V_x)t_b$ volt-seconds.

Consequently:

(positive area) − (negative area) = $V_a t_a - V_b t_b + V_x(t_b - t_a)$ and the mean value of the output voltage V_2 is:

$$V_2 = \frac{V_a t_a - V_b t_b + V_x(t_b - t_a)}{t_a + t_b}$$

Now $V_a + V_b = V_k$ is a constant,

and $t_a + t_b = t_k$ is a constant.

Also, since it is assumed that there is no d.c. content in the waveform of the input voltage V_1 in Fig. 4,

$$V_a t_a = V_b t_b$$

Hence $V_a = V_k\,(t_b/t_k)$

and $\quad V_b = V_k\,(t_a/t_k)$

Substituting in these values, re-arranging and eliminating gives:

$$\text{Mean output } V_2 = V_x\,(1 - 2(t_a/t_k))$$

Thus under these conditions the mean value of the output voltage V_2 applied to the resistor R in Fig. 3 is linearly proportional to t_a/t_k. When $t_a = t_k/2$, i.e. when the mark-to-space ratio is unity, the mean output is zero since the positive shaded area in Fig. 4 equals the negative shaded area. Note that the mean output is quite independent of the peak-to-peak input amplitude V_k in this condition where $V_b > V_x$.

Now if the second case, with V_b less than V_x, is considered, there is no conduction at all in the negative direction and:

$$\text{mean output } V_2 = \frac{(V_a - V_x)t_a}{t_k}$$

However, as before $V_a = \dfrac{V_k(t_k - t_a)}{t_k}$

so that $V_a - V_x = \dfrac{V_k(t_k - t_a) - V_x t_k}{t_k}$

and mean output $V_2 = (V_k - V_x)\,(t_a/t_k) - V_k(t_a/t_k)^2$.

Thus in this case the mean value of the output is quadratically dependent on the value of t_a/t_k. Differentiation with respect to t_a/t_k shows that the mean value of V_2 is a maximum (provided that $V_b < V_x$) when:

$$t_a/t_k = \frac{V_k - V_x}{2V_k}$$

$$= \frac{V_k/V_x - 1}{2V_k/V_x}$$

The mean value of the output can be written as:

mean output $V_2 = V_x[((V_k/V_x) - 1)t_a/t_k - V_k/V_x(t_a/t_k)^2]$

Note that the value of the mean output V_2 depends on the peak-to-peak input amplitude in this case where $V_b < V_x$.

The borderline between the two conditions considered occurs when $V_b = V_x$,

But:

$$V_b = V_k(t_a/t_k)$$

so that the borderline occurs where:

$$t_a/t_k = V_x/V_k$$

If $t_a/t_k > V_x/V_k$, then the mean output voltage V_2 across resistor R depends linearly on t_a/t_k.

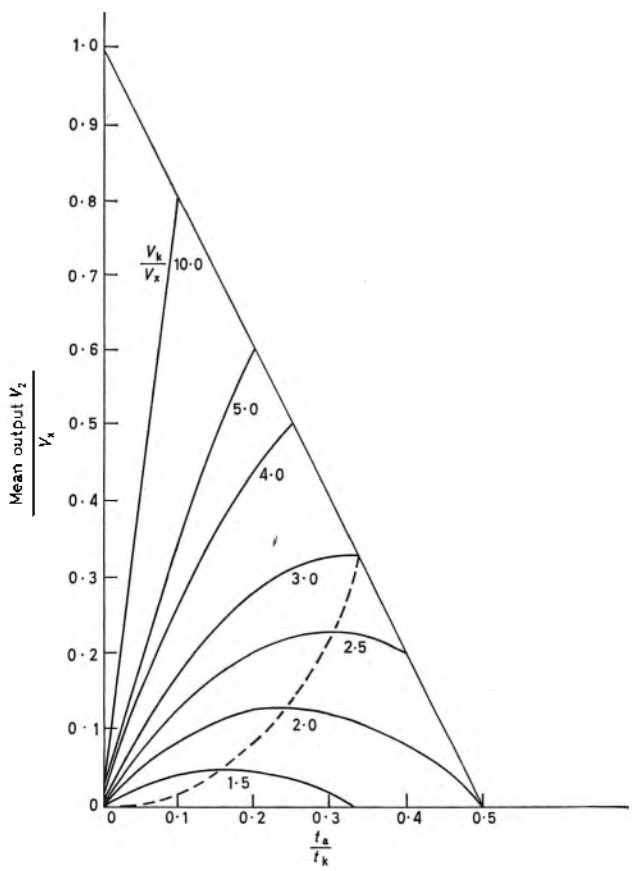
If $t_a/t_k < V_x/V_k$, then the mean output voltage V_2 across resistor R depends quadratically on t_a/t_k.

The borderline occurs at the point where:

mean output $V_2 = V_x(1 - 2(V_x/V_k))$
$\qquad = V_x(1 - 2(t_a/t_k))$

Curves showing how $\dfrac{\text{mean output } V_2}{V_x}$ varies with t_a/t_k are given in Fig. 5. If t_a/t_k exceeds 0·5, the shape of the curves is repeated symmetrically in the negative direction. The curves are plotted for various fixed values of relative input amplitude V_k/V_x. Various points of interest will be noted from these curves. Firstly if V_k/V_x is less than two (i.e. if the peak-to-peak amplitude of the input waveform is less than the 'clipping range' or 'window' of the non-linear device), there can be no linear portion of the curves. Instead, there is a range of mark-to-space ratios over which the mean output voltage is zero, as seen for example on the curve for $V_k/V_x = 1·5$. If, however, V_k/V_x exceeds two, there is a linear portion on the

Fig. 5. Characteristics of symmetrical rectifier



the output curve. As V_k/V_x is increased, the maximum point on the output curve follows a circular locus until, when V_k/V_x equals three, the maximum point falls on the linear part of the curve, and the maximum point on the quadratic curve is never reached.

If the peak-to-peak input amplitude V_k is greater than about four times the 'window' V_x, then the mean output voltage V_2 is linearly proportional to the mark-to-period ratio t_a/t_k over quite a wide range, and it is completely independent of the actual value of V_k. This is advantageous compared with mark-to-space measuring arrangements which are sometimes used. The maximum possible mean output, obtainable when V_k/V_x is very large, is equal to V_x. Practical curves taken using Zener diodes connected back-to-back as the non-linear element approximate closely to the predicted proportionality, but the peak height of the curves is reduced.

Some Zener diodes give the required symmetrical characteristic inherently, though it can easily be produced by two normal asymmetrical Zener diodes connected as mentioned above, back-to-back in series so that one conducts in a forward direction when the other conducts in a reverse direction. Various other devices can give the symmetrical current voltage characteristic of Fig. 2. Two conventional rectifier diodes in reverse parallel connexion give a degraded form of the characteristic, with the sharp corners replaced by more gradual curves. A voltage limiting device such as a single silicon carbide disk gives a similar curved characteristic. Gaseous breakdown devices such as cold-cathode diodes give a similar form of characteristic with a higher voltage, though in this case a voltage peak must be surmounted before the flat constant voltage region is reached. A similar peaked characteristic, though with a lower voltage, was obtained with one of the earliest electronic devices, the coherer. Any of these devices is suitable for use in pulse rectification arrangements such as that of Fig. 3.

The fact that pulse rectification is possible with devices having symmetrical characteristics naturally leads to the question of the possibility of rectification using the symmetrical saturation characteristics of iron cored components. Unfortunately, the saturation of flux density in such devices depends on the voltage integral applied to a winding rather than on the peak value of voltage. In pulse waveforms of the type discussed here, the positive voltage integral always equals the negative voltage integral since it has been assumed that the mean value (i.e. the net integral over one complete period) is zero. Consequently changes of the mark-to-space ratio cannot produce saturation in one direction only. However, if it can be assumed that the peak-to-peak voltage amplitude is maintained constant, it should be possible to avoid saturation when a pulse is short and the mark-to-space ratio is small, but to produce saturation, alternately positive and negative, as the mark-to-period ratio approaches 0·5 (i.e. the mark-to-space ratio approaches unity). This possibility has not been investigated. The asymmetrical non-linear characteristics of magnetically biased iron cored devices have been used in demodulation arrangements[2,3]. The possibility of rectification without non-linear resistance was considered by Gewartowski[4].

The symmetrical rectifier has possible applications in feedback clipping circuits[5,6,7,8] such as that shown in the block diagram of Fig. 6. Here, an input waveform is clipped to produce a pulse waveform. The mark-to-space ratio of the clipped waveform is measured by a symmetrical rectifier and the rectified output is filtered and added to the input. The addition is carried out in such a

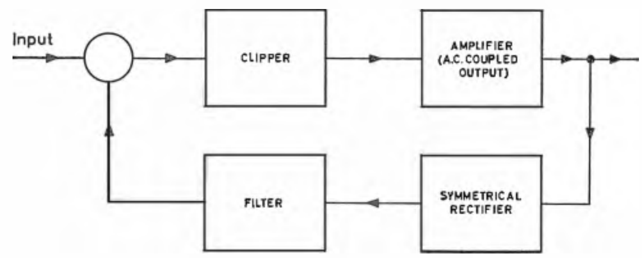


Fig. 6. Arrangement of feedback clipper

sense as to hold the mark-to-space ratio at any desired value despite changes of input waveform or circuit parameters. Such an arrangement could be extended by the addition of an input to a comparison circuit after the filter to produce a pulse-width modulator arrangement.

There are other possible applications of a device having symmetrical characteristics. For example, a carrier wave having an amplitude just less than the threshold 'window' amplitude V_x would give no carrier frequency output. However, an output would appear when a modulating voltage was added to the carrier wave. With such a circuit it would be desirable to make the threshold level adjust itself automatically to any variations in the carrier level (or vice-versa). Such an arrangement might be used as a form of suppressed carrier modulation system. Another possible application depends on the fact that odd-harmonic distortion in the input waveform produces no output from the symmetrical rectifier, while even-harmonic distortion does produce an output. When an iron-cored transformer is switched on to an a.c. supply, a heavy 'inrush' current can flow initially in the primary winding. On a power transformer, this inrush can cause mal-operation of differential protective devices which compare the primary and secondary currents, and it is necessary to restrain the operation of the protective gear at switch-on. The inrush current has a large second harmonic content, whereas normal fault currents have not, and the symmetrical rectifier could be used in a restraint circuit. A short search of the literature reveals that symmetrical rectifiers have in fact been used with asymmetrical arrangements, in particular to detect second harmonics[12]. With the growing interest in pulse width modulation techniques, the symmetrical rectifier should find an increasing range of applications.

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A Simple Automatic Photographic Enlarger Timer

By J. L. Linsley Hood

In the circuit described a photoconductive cell is used to measure the brightness of the projected image and, by means of a simple integrating circuit, to control the exposure time. Exposure times between 1sec and 40sec are provided.

(Voir page 489 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 496)

ONE of the most time-consuming tasks in the average small darkroom is the determination of the correct printing time for the enlargement of negatives. Fortunately, there is a certain measure of latitude in the actual time of exposure during printing of any given negative for which a satisfactory print can be made, and an experienced worker can usually judge, from the appearance of the negative, the order of exposure which will produce an acceptable print.

However, appearances are sometimes misleading, and the estimated exposure may not be within the usable limits. For these reasons, the author has speculated for some years on the possibility of constructing an instrument which would measure the brightness of the projected image on the enlarger baseboard, and control the printing time so that the printing exposure was always within acceptable limits.

The measurement of the brightness of the projected image can be done in several ways. Of the possible alternatives the most suitable appeared to be the use of a photocell mounted above the baseboard, so as to measure the light scattered from the surface of the printing paper; and employed in a simple integrating circuit; although a very sensitive photocell arrangement would be needed. Although adequate sensitivity could be obtained by the use of a photomultiplier, or even a normal photoemissive cell, neither of these arrangements recommended themselves as sufficiently simple or inexpensive.

However, recent developments in photoconductive cells have enabled these components to be used in applications involving very low levels of light intensity, and, in particular, the sintered cadmium selenide photoconductive cell combines the necessary qualities of high sensitivity to blue light, and fast response at low light levels.

An experimental automatic enlarger timer has been built using a cell of this type in a simple *CR* integrating circuit of the type shown in Fig. 1, and experience with this has shown that it is capable of providing reasonably appropriate exposures for the printing of negatives of a wide range of densities, with a variety of enlarging lens apertures and degrees of enlargement. The equipment itself is compact and relatively inexpensive to construct.

Description of Circuit (Fig. 2)

The cell employed is an acrylic resin encapsulated sintered cadmium selenide photoconductive detector, type TP.MD-6 FR., marketed in the U.K. by Photain Controls Ltd. This has a dark resistance exceeding 500MΩ, and a linear relationship between conductivity and illumination over the range of light intensities of interest. In this application the cell resistance lies between 0·5 to 20MΩ, giving printing exposure times in the range 1 to 40sec, with a 4μF polyester foil timing capacitor.

At the start of the timing cycle, the capacitor is discharged through S_2 and R_1, and the potential applied to the base of transistor VT_4 falls below the Zener voltage developed across the Zener diode MRZ_2. When S_2 is released the circuit to the enlarger lamp is made through S_2 and the normally closed contact of the switching relay *A*, and the timing capacitor charges through the photoconductive cell, at a rate determined by the light incident

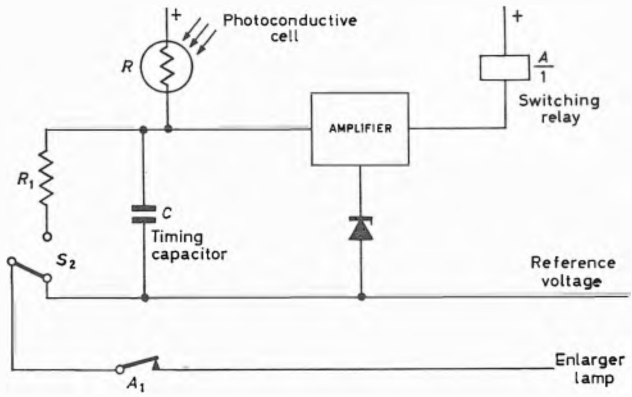


Fig. 1. General arrangement of timer

on the photocell, and the setting of the 'Paper speed' potentiometer RV_1. A field effect transistor could be used with advantage in the input stage of the circuit, as shown in Fig. 3. However, in the present circuit the necessary high input impedance has been obtained by the use of an augmented emitter-follower circuit ($VT_2\,VT_3$) in which the input circuit ($VT_1\,VT_2$) is a conventional Darlington pair.

Unfortunately, transistors are current amplifying devices, in whichever circuit configuration they are used, and since the input current to the timing capacitor may be as low as 1μA, at low light levels, it is undesirable that the input current to the subsequent amplifier should exceed 0·1μA. A current amplification of 400 000 is, therefore, necessary to provide the 40mA operating current required by the switching relay in the collector circuit of VT_4. Although this only requires a mean current gain of somewhat less than 27 in each stage, transistor current gains fall drastically at low collector currents, and even with a high gain transistor in the input stage, the current gain to be expected from this stage at a collector current of only a few microamperes, is only of the order[1,2] of 10.

This consideration imposes a more serious limitation on the input impedance obtainable with transistor d.c. amplifiers than the collector-base leakage current, which can be less than 10nA at room temperature with modern silicon transistors. High gain transistors are used, therefore, where this is possible without incurring the penalty of high leakage currents. A silicon Zener diode is used to stabilize the emitter potential of the output transistor, in order to obtain a more precise 'turn-on' potential for the charging circuit than would be given by the 'pull-in' characteristics of the relay.

POWER SUPPLY

In order to reduce the size and cost of the unit, a 'transformerless' design is used. This employs two medium-power Zener diodes in a modified Cockcroft-Walton rectifier circuit, and provides a unidirectional square-wave input to the two half-wave rectifiers MR_1 and MR_2, of amplitude equal to the Zener voltage of the diodes, which gives effective stabilization against changes in mains input voltage.

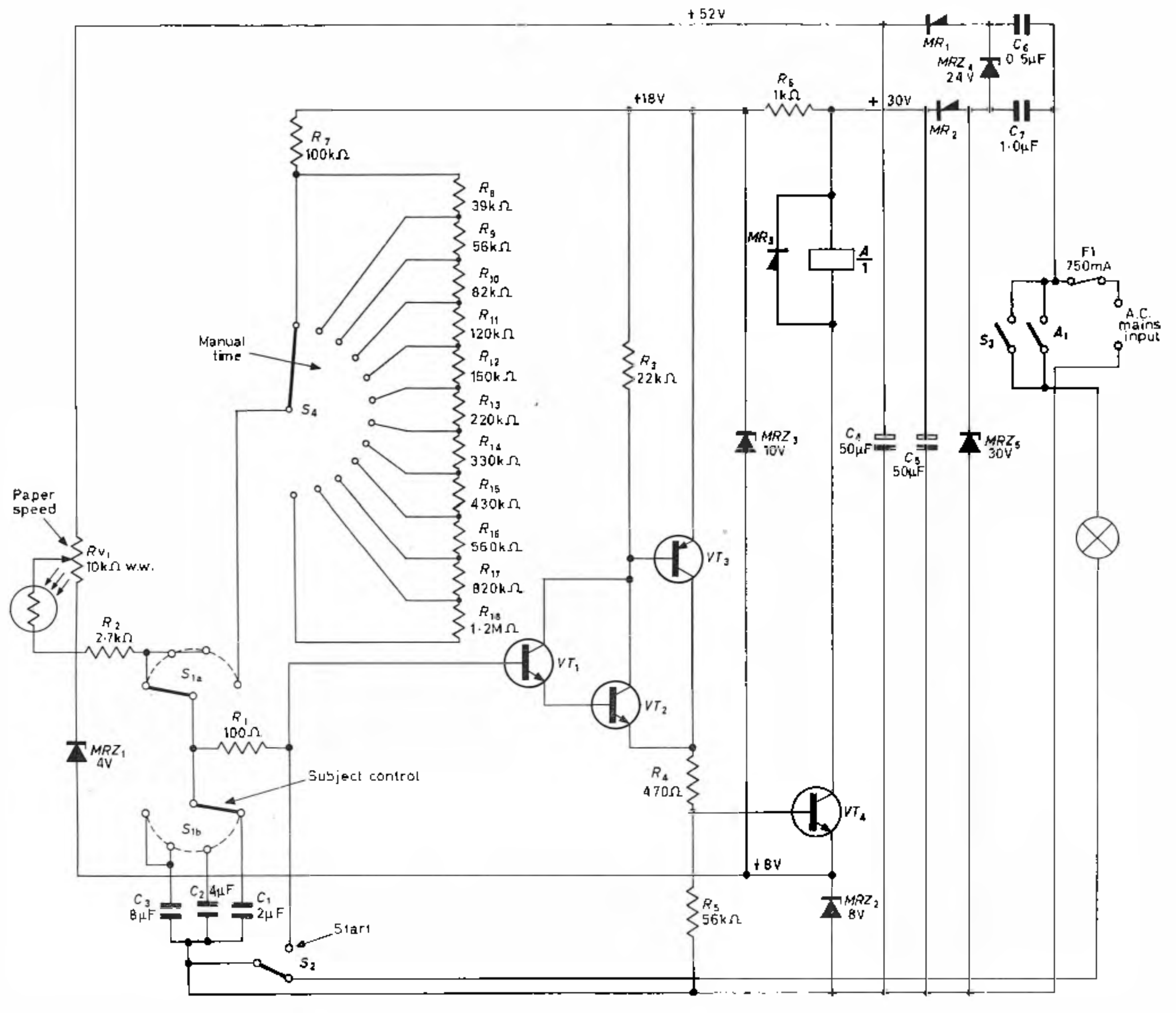
Method of Use

The photoconductive cell is mounted in a suitable housing, on an arm attached to the base of the enlarger column, and is arranged to receive light from an area in the centre of the baseboard approximately equal to the size of the smallest print normally made. A simple lens of large aperture and suitable focal length is employed to focus this area of the enlarged image on to the sensitive area of the photocell. Although a simple double convex lens is adequate, a large aperture 8mm cine lens, of some 20mm focal length, would be ideal, and would allow the photocell to be mounted more conveniently out of the way of the projected image, which would avoid difficulty when large prints are made. The aperture of the lens is adjusted, if necessary, so that suitable exposure times are given, with a fast bromide paper, when the 'Subject' control switch is in the normal position (position 2) and the 'Paper speed' potentiometer is set near the maximum positive position.

Ths 'Subject' control switch is necessary since the operation of the circuit is based on the assumption that the tonal range of the print can be integrated to grey, and while this is frequently approximately true, there are obvious occasions on which the subject will be predominantly of a lighter or darker tone, and some allowances for this fact must be made, so that the mean density of the print can be made lighter or darker as necessary.

In addition to the three mean-density positions on the switch, a fourth position is employed to provide a

Fig. 2. A simple automatic enlarger timer



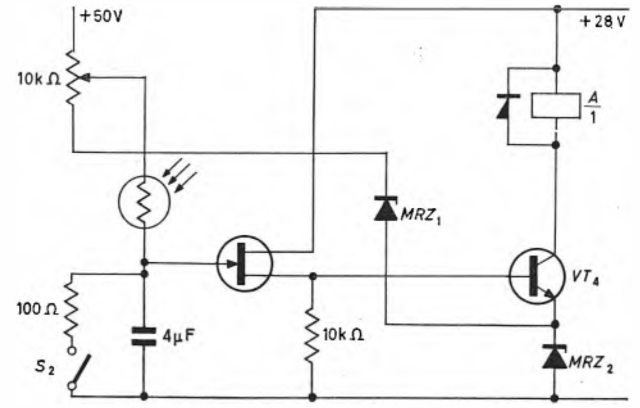


Fig. 3. Use of field-effect transistor

manually pre-set range of time intervals through the resistor chain on switch S_4. These cover the range 1 to 40sec in 1:1·4 ratio steps, and are intended for use when it is necessary to hold back, or 'print-in' particular areas of the print during the enlarging exposure.

It is desirable that the light from the darkroom safelight should not influence the operation of the timer circuit excessively, and the position of the photocell head and the illumination of the enlarger baseboard by the safelight should be adjusted so that the timing period provided by the apparatus when the enlarger lamp is disconnected is at least five times greater than the longest exposure times normally required. If this order of 'dark' times cannot be obtained conveniently, the safelight may be connected in series with the unused changeover contact on the relay *A*, so that it is switched off during the printing exposure. The contacts of this relay are connected in parallel to minimize contact erosion.

Once the effective speeds of the printing paper—developer combinations have been determined, a sequence of satisfactory prints can be made using a wide variety of negatives and degrees of enlargement, without further adjustment, except as required by change of subject type or deterioration, through use, of the developer.

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Recording Underwater Currents

A recording current meter having an 80-day in-sea cycle has been produced by Marine Systems Division of the Plessey Electronics Group.

Designated MO 20/21, the recording current meter is one of the latest aids to underwater investigations. Primarily designed to measure the direction and speed of ocean or river currents, the Plessey recording current meter will record this information and other parameters at varying sample rates at periods of up to 80 days in the water. It will operate down to 2000 metres and its performance can be monitored after installation, ensuring correct functioning without having to surface the unit for examination.

Special construction of non-corrosive materials, including stainless steel, naval brass and anti-foulant coated plastics, ensure that the meter is fully protected against the effects of attack by sea. Standard ¼in magnetic tape is used for recording the information which is recorded in digital form. This is easily translated for computer in-feed, greatly simplifying data processing.

The meter is normally fitted with sensors to measure current speed and current direction, but other sensors may be added. There are two versions, one fitted with a propeller rotor and the other fitted with a Savonius-type rotor.

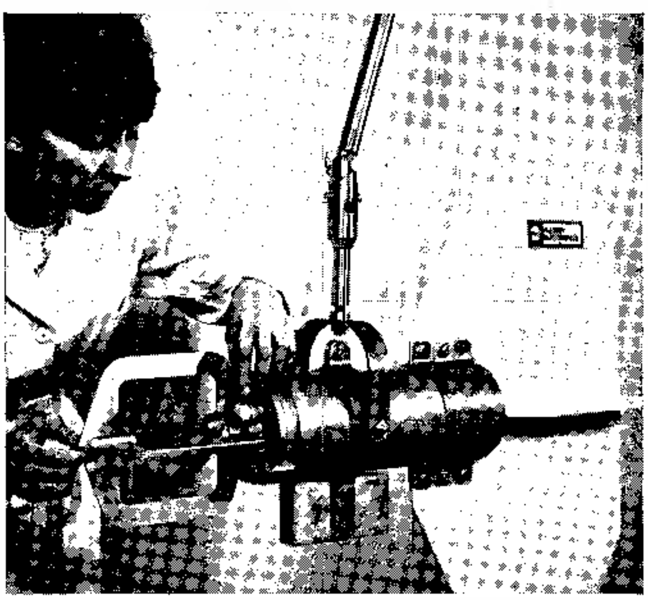
The electronic circuits are fully transistorized and power can be supplied from normal U2 cells for short periods or mercury cells for longer immersions. Any physical parameter measured must be converted into a variable resistance value. This signal is then fed through an analogue-to-digital convertor. A 10 bit binary code is used where binary 'one' is recorded as a 10msec pulse, and a binary 'zero' as a 30msec pulse.

A translation unit is available to convert the magnetic tape output to punched tape, suitable for computer in-feed. This is also available as a postal facility from the Plessey Marine Systems Division at Ilford, Essex.

The output from the meter can also be taken directly by cable to a deck print-out unit, which shows the selected parameters as binary numbers, easily correlated with the appropriate measurement by referring to a calibration chart.

A novel feature of the Plessey recording current meter is the acoustic transducer which transmits the data being recorded on the tape allowing it to be monitored without any connecting wires up to a range of 800 metres. It also allows the correct functioning of the meter to be monitored after it has been moored, and provides an easy method of location in the event of the meter becoming detached from its moorings.

The underwater current meter



A Series Current Regulator

By T. M. Palmer*

This article describes a simple alternative to a complete current-stabilized power supply which takes the form of a self-contained series element that passes a constant current independent of applied voltage.

(Voir page 489 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 496)

A SIMPLE alternative to a complete current-stabilized power supply[1-5] is a self-contained series element that passes a constant current independent of applied voltage. With such an element, either an existing constant-voltage or an unregulated power supply can deliver a constant current to an electromagnet or other load, the resistance of which may vary providing the voltage drop across the series element remains within certain limits. The correct operating range is indicated on a voltmeter and the unit is also protected against excessive voltage.

Design

Use is made of the inherently high collector impedance of a transistor and of an amplified feedback system in which the potential across a series resistor R_5' shown in Fig. 1 is compared with the constant potential across a Zener diode *MRZ*.

The difference signal between these potentials is amplified by VT_2 and used to control a current-amplifier VT_1' which represents VT_3, VT_4 and VT_5 of Fig. 2; the voltage across VT_1' is indicated by a meter V and this voltage must lie within a prescribed range. Refinements to the simple circuit have been incorporated and are shown in Fig. 2.

First, to select any current within the design range, the effective value of R_5' (Fig. 1) must be continuously variable. Due to the low value and high dissipation (~40W) of R_5', it is more convenient to use the circuit in Fig. 2 with RV_2, R_9 and R_{10} in the base circuit of VT_2 for fine control, and to switch R_8 instead of varying it continuously. Secondly, the gain of VT_2 is increased by increasing the collector load impedance which would be R_c (Fig. 1) if the collector of VT_1 was at constant potential. However, this potential is out-of-phase and amplified G_o times relative to the base of VT_1', so that the effective collector load impedance of VT_2 is in fact R_o/G_o. G_o must be high, so R_o must be made correspondingly higher by using, in place of a resistor, the collector of VT_1 with associated circuit as shown in Fig. 2.

The effect of a change in V is compensated by introducing through R_4, RV_1 to the emitter of VT_1, an adjustable current which increases with V, thus reducing proportionately the base current of VT_3 and the collector current of VT_4, VT_5.

The increased gain of the circuit of Fig. 2 over that of Fig. 1, due to a large effective value for R_o and high current gain of the emitter-follower VT_3 driving VT_4 VT_5, makes a 'step-function' network such as R_5C_3 essential for an adequate stability margin in accordance with the Nyquist criterion.

Rapid protection against over-voltage across VT_1, VT_3, VT_4 and VT_5 is accomplished by conduction of the Zener diode MRZ_1, so maintaining the conductance of these transistors and limiting the voltage-drop across them. Sustained over-voltage causes the relay to operate and hold-on, thus protecting VT_2 and MRZ_2, and releasing only when the current is considerably reduced.

Overall Impedance

The efficacy of the unit in preventing a change in current due to a change in applied voltage or load resistance can be expressed as an overall impedance which is the ratio of the voltage change across the unit to a small current change, and has been determined experimentally by measuring the ratio of a small series-injected a.c. voltage to the resultant a.c. current, superimposed on the main d.c. current.

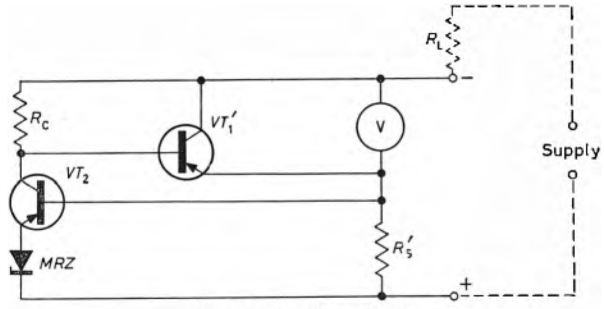


Fig. 1. Simplified circuit

Calculated Performance

The equivalent circuit used for a transistor is the hybrid π as described by Cherry[6,7] and as shown in Fig. 3. In this article, single letters are used for the parameters to avoid double subscripts, as number subscripts denote the transistor being considered. With these symbols, there is the additional condition that $GW = \beta$.

For the purposes of this discussion the circuit of Fig. 2 is divided into the following sections and these are considered in turn.

(a) The base network for VT_2
(b) The voltage amplifier VT_2
(c) The collector load for VT_2 formed by VT_1
(d) The input impedance to the base of VT_3
(e) The current amplifier VT_3, VT_4, VT_5

The performance of the circuit is obtained by calculating the effect of a change ΔI of load current in R_8 to obtain finally the resultant change in V.

(a) Considering first the effect of a change ΔI in the circuit associated with RV_2 and R_8, the voltage change at the base of VT_2 is:

$$\Delta I \,.\, R_8 \left\{1 - \frac{nR_9}{R_9 + R_{10} + n(1-n)RV_2}\right\} \ldots\ldots (1)$$

where the slider of RV_2 divides it into variable parts nRV_2 and $(1-n)RV_2$. Expression (1) may be written: $\Delta I \,.\, R_8'$.

* *National Standards Laboratory, C.S.I.R.O., Australia.*

(b) The voltage gain of VT_2:

$$G_2 = \frac{K_2}{A_{c2} \cdot R_B} \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (2)$$

where the current gain $K_2 = \beta_2$ since the external collector load admittance A_{c2} is large compared with $2/F$. The total base resistance $R_B = R_N + \beta_2 R_Z + S_2 + W_2$ where R_N is the resistance of the R_9, R_{10} circuit, R_Z of the Zener diode emitter circuit and S_2, W_2 are as shown in Fig. 3 referred to VT_2. A_{c2} is the sum of the input admittance to the base of VT_3 and the collector admittance A_{c1} of VT_1.

(c) The collector impedance R_c of VT_1 is increased considerably by the emitter resistor R_3 and, referred to the negative line, (neglecting the internal emitter resistance):

$$R_c = (F_1/2) + \frac{\beta_1 F_1 (R_3 + (1/2G_1))}{2(S_1 + R_3 + (W_1/2))} \quad \ldots\ldots\ldots \quad (3)$$

with $R_3 = 47\Omega$, $S_1 = 300\Omega$, $F_1 = 25\text{k}\Omega$, $\beta_1 = 50$

$G_1 = 0{\cdot}4\text{mho}$ and $W_1 = 125\Omega$, then $R_c = 86\text{k}\Omega$.

As previously described, the effective impedance is considerably reduced by the gain, G_o, of VT_1':

$A_{c1} = G_o/R_c$, and with $G_o \simeq 200$, $A_{c1} = {\cdot}0023\text{mho}$.

(d) The input impedance to the base VT_3 depends considerably on the effective impedance in its emitter denoted by R_{E3}. Using the suffix 'o' to denote the parameters of VT_4, VT_5 in parallel, $R_{E3} = S_o + \beta_o R_{Eo}$, where:

$$R_{Eo} = R^6/2\,(= R_8/2) \quad \ldots\ldots\ldots\ldots \quad (4)$$

In the case of VT_3, it is necessary to consider the internal feedback parameters when determining the base input resistance R_{in}:

$$R_{in} = S_3 + W_3 + \frac{2\beta_3 R_{E3} F_3 - W_3 \beta_o F_o}{2(F_3 + \beta_o F_o)} \quad \ldots\ldots \quad (5)$$

with typical values $R_{E3} = 2{\cdot}7\Omega$, and $R_{in} = 400\Omega$.

Referring back to section (b), the collector load admittance for VT_2 is:

$$A_{c2} = A_{c1} + (1/R_{in}) = {\cdot}0048\text{mho} \quad \ldots\ldots\ldots \quad (6)$$

(e) To determine the output voltage change it is necessary to calculate the trans-impedance R_{To} (ratio of voltage change at the collectors of VT_4VT_5 to the current change

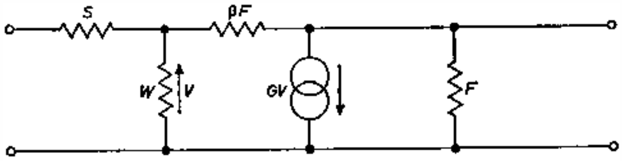


Fig. 3. Hybrid π equivalent circuit of a transistor

(corresponding to Cherry's circuit in which

$S \equiv r_{B'}$, $F \equiv r_c$, $W \equiv \frac{r_E}{1-a_N}$, $G \equiv \frac{a_N}{r_E}$ *)*

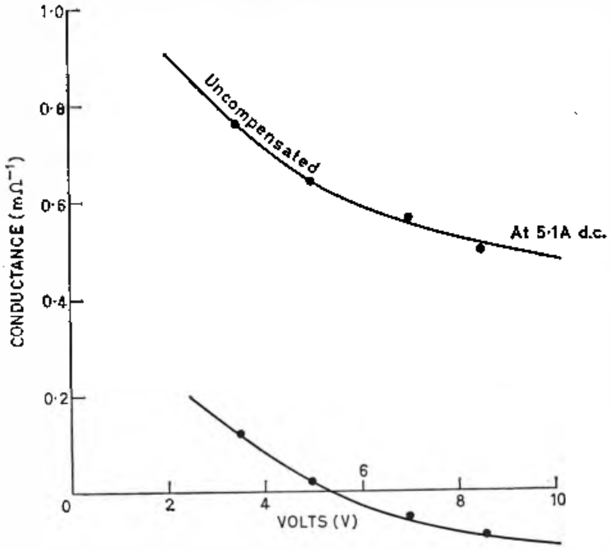


Fig. 4. Variation of conductance with V

into the base of VT_3). For an external load resistance of VR_L:

$$R_{To} = \frac{\beta_o \beta_o F_o F_3 R_L}{2R_L(\beta_o F_o + F_3) + F_o F_3} \quad \ldots\ldots\ldots\ldots \quad (7)$$

with R_L open and at an output current of 5A, $R_{To} = 80\text{k}\Omega$. The voltage gain of the final three stages $G_o = R_{To}/R_{in} = 200$.

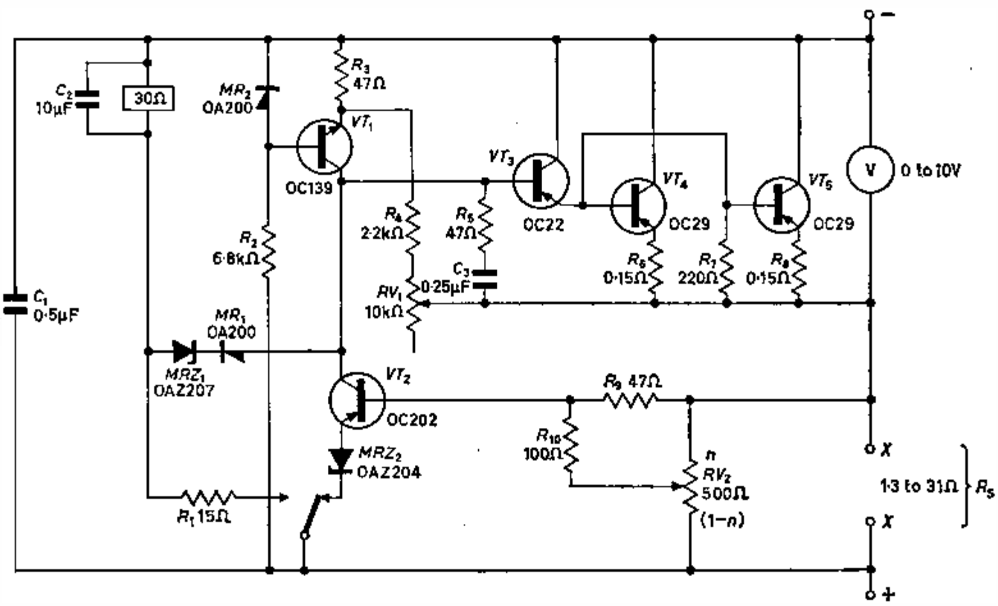
The overall impedance of the unit R_T is the product of the gain of the various parts and the trans-impedance:

$$R_T = R_s' \times G_2 \times G_o =$$

$$R_s' \times \frac{\beta_2}{A_{c2} R_{B2}} \times G_o$$

$$= 2000\, R_s' \text{ ohms}$$

Fig. 2. Full circuit



This figure is in reasonable agreement with the experimental value shown in Fig. 4 for a current of 5A, and 6V across VT_4VT_5. G_o and R_s' both increase as the reciprocal of the current so that R_T varies as the inverse square of the current. Cherry[7] shows that F varies as the square root of the collector voltage; thus G_o and R_T also vary in this way.

Adjusting RV_1 for optimum compensation raises R_T indefinitely at a particular current and voltage drop across the unit. Over a range of current and voltage an improvement of 4 to 5 times is practical and adequate.

The details of R_8 are not shown. It is a high-dissipation switched resistor suited to the particular current range required.

Temperature Compensation

The unit is mostly used in a temperature-controlled room, but the overall temperature coefficient can be reduced to about 0·02 per cent/°C by matching the positive coefficient of 3mV/°C for MRZ_2 (7V Zener diode) to that of the base-emitter potential of a silicon transistor VT_2.

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Encapsulation of Electronic Components by Transfer Moulding

The recent availability of epoxy resins that can easily be transfer moulded has created a fresh field for specialized types of compression moulding presses. These new epoxy resin materials have improved electrical properties. They have also been formulated to give extremely good metal bonding characteristics. A further feature of such materials is that they can be transfer moulded at very low pressures thereby allowing delicate components to be encapsulated. To take full advantage of these materials and their advanced moulding techniques, two new types of Daniels (of Stroud, Gloucestershire) transfer presses—called "Microjet" Presses—have been specially designed and built. One type is a 25 ton machine with a 5 ton transfer (Type 25/5E2) while the other is a 12 ton press with a 3 ton transfer (Type 12/3E2). Both of these presses are marketed in the U.K. by Alfred Herbert Ltd, Coventry.

The method of direct encapsulation by transfer moulding is relatively simple. A mould is manufactured to produce the required item and has a number of mould cavities of the correct size and shape. The electronic item is positioned in the lower half of the mould and located by its leads in slots machined to close tolerances to prevent material leakage. The mould is closed and clamped by the press at a high pressure with the electronic items correctly positioned in the mould.

The thermoset moulding material is placed in the transfer cylinder and is forced by means of the transfer unit at a controlled speed and pressure through a runner system in the heated mould into the actual mould cavities. After a predetermined period the material cures and the completed mouldings are extracted.

Moulding powders used for encapsulation work—especially the epoxy grades of material—have a very low viscosity and a good metal to metal seal between the mould halves is required to minimize flash. The high clamp force assists in obtaining this metal seal.

The hydraulic system to the press clamp gives a fast press approach speed and, at a predetermined stroke position—which is adjustable anywhere within the press daylight—the closing of the press is gradually retarded to a speed which is variable down to nearly zero. The two mould halves then gently close together and the hydraulic clamp force of 12 tons is applied.

A mould protection device is fitted to the press to protect the mould in the event of a component being incorrectly positioned in the mould as the press closes. This can be positioned anywhere within the daylight of the press and with its micro-adjustment is extremely sensitive. Should a component be displaced and cause interference whereby the mould halves do not close metal to metal, then the device is not operated, the full clamp pressure cannot be exerted and the press pump motor will stop, thereby protecting the mould. When the press pump motor is restarted the press clamp opens automatically.

There is speed control which comes into operation when the hydraulic clamp section opens on completion of a press cycle. Initial slow opening is obtained allowing the two mould halves to part slowly and—if top ejectors are used—the ejectors gently ease the mouldings out of the mould top force. The final 2in of clamp opening movement are speed controlled for bottom ejection to take place, thereby avoiding damage to mouldings during ejection.

Two electric resistance type platens (12in square) are fitted to the press and the temperature is independently controlled by means of thermostats mounted in the control panel. The top platen is bored in the centre for locating the transfer sleeve; this sleeve is normally part of the mould.

Bottom ejectors are fitted with a pneumatic reset mechanism. This system of automatically resetting the ejectors eliminates the lengthy procedure sometimes encountered of closing the guard—inching the press up until the ejector pins are clear of the mould cavity and then again opening the guard to reload the mould with inserts.

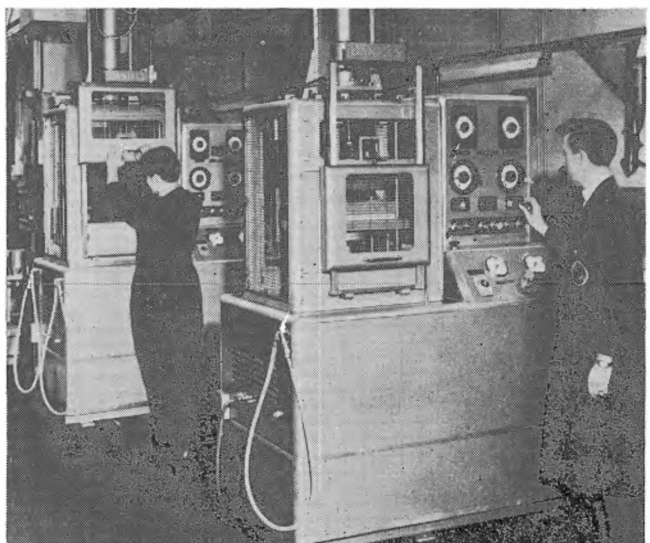
The transfer system is probably the most important feature of the press. Hydraulic pressure in the transfer circuit can be varied to give a transfer force from 165 lb. up to 3 tons. With this adjustment, various transfer pressures can be obtained using a variety of plunger tip sizes. Using a tip size of 1¾in diameter the material transfer pressure can be varied from approximately 50 lb/in² up to 2 800 lb/in². The transfer unit is arranged to give fast approach and, prior to the transfer plunger contacting the material, the speed is reduced to give a transfer speed variable from approximately 1·3in/sec down to zero speed.

After moulding is complete a small amount of cured material ('biscuit') remains in the transfer sleeve. The machine is arranged so that when the press opens the transfer ram advances a pre-set distance, ensuring that the 'biscuit' of material remains in the bottom half of the mould.

A timer is fitted which is initiated when the transfer ram advances. When the pre-determined cure period has elapsed, as set on this timer, the press opens automatically and the transfer retracts.

A preheater timer is fitted to the machine. The purpose of this is to programme the preheating of the material in a 1½kW or 2½kW 80Mc/s high frequency preheater, which can be linked to the machine, so that on completion of the press cycle, the preheated pellet is ready for loading into the press. One of the features of epoxy moulding powders is its low viscosity and it is, therefore, important that the material is offered to the machine in a consistent condition.

The 'Microjet' 12 ton press, with the 25 ton machine in the background.



LETTERS TO THE EDITOR

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

The Modified Z-Transform

DEAR SIR,—I was very interested in the article by Mr. Y. Azar[1] (February 1966). It seems to me that the conclusions of Mr. Azar concerning the modified Z-transform are too severe and should be corrected a little.

The modified Z-transform (or modified discrete Laplace transform, which is practically the same) has been introduced — apparently independently — by Cypkin[2], Barker[3] and Ragazzini and Zadeh[4], in three different forms.

Starting from formula (12) of Mr. Azar and using the letter ϵ instead of m. Suppose firstly $f(t)$ is continuous for all t. Then

$$F^*(s,\epsilon) = \sum_{n=0}^{\infty} f[(n-1)T+\epsilon T] \, . \, e^{-sTn} =$$

$$= e^{-sT} . \sum_{n-1=-1}^{\infty} f[(n-1)T+\epsilon T] \, . \, e^{-sT(n-1)} =$$

$$= e^{-sT} \sum_{k=0}^{\infty} f(kT+\epsilon T) e^{-sTk} = e^{-sT} . \phi^*(s,\epsilon) \quad \text{......... (1)}$$

where $k = n-1$, and the term of the series for $k = -1$ is zero, since $0 \leqslant \epsilon \leqslant 1$, and $f(t) = 0$ for $t \leqslant 0$ by the definition of the Z-transform. If $f(t)$ has jumps for integer t, this must be respected in the definition of the Z-transform[6]. Now, introducing in accordance with formula (5) of Azar the function

$$F^*(s,\epsilon) = \sum_{n=0}^{\infty} f(nT+\epsilon T) \, . \, e^{-sT(n+\epsilon)} \quad \text{...... (2)}$$

From equations (1) and (2), it is clear that

$$F^*(s,\epsilon) = z^{1-\epsilon} F^*(s,\epsilon) = z^{-\epsilon}\phi^*(s,\epsilon) \quad \text{..... (3)}$$

with $z = e^{sT}$ (see also Prouza[5], formulae (2, 7) and (2, 10)). $\phi^*(s,\epsilon)$ was essentially introduced by Cypkin[2], $F^*(s,\epsilon)$ by Barker[3], and $F^*(s,\epsilon)$ by Ragazzini and Zadeh[4].

It is clear from equation (1) that $F^*(s,\epsilon)$ is delayed by one sampling period with respect to $\phi^*(s,\epsilon)$ and this must be respected in the inversion formulae (see Prouza[5], formulae (7) and (8)). Erroneous results can be caused by misinterpretation, not by the transforms themselves.

The first table of $F^*(s,\epsilon)$ was published by Barker and extended substantially by Jury and his colleagues. Today, $F^*(s,\epsilon)$ is in general use in the Anglosaxon literature as $\phi^*(s,\epsilon)$ in the Soviet literature.

In my opinion there is little hope that the function $F^*(s,\epsilon)$ and the tables thereof will be introduced for wide use, as Mr. Azar suggests, although it has some advantages, as one sees from Mr. Azar's article and from Ragazzini and Zadeh[4].

Mathematically, $F^*(s,\epsilon)$ is the only correct extension of the ordinary Laplace transform (in the Stieltjes sense) for arbitrarily delayed sequences of impulses.

Yours faithfully,

L. PROUZA,
Pardubice, CSSR.

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The author replies:

DEAR SIR,—I would like to thank Mr. Prouza for his interest in my article and for throwing more light on the historical background of the various forms of the modified z-transform.

I am very delighted that he finds my proposed transform advantageous as:

(a) It avoids any misinterpretation.

(b) It avoids the tedious transformation from the z to the t plane, when one wants to determine the $Z^{-1}[F(z,m)]$ for a particular value of t_1 (equation (20) in my article).

Yours faithfully,

Y. AZAR,
The Solartron Electronic Group.

A Direct Method for Determining the Symmetry Factor of a Junction Transistor

DEAR SIR,—The degree of symmetry of a symmetrical, and in some special cases[1] of a usually unsymmetrical transistor, is its symmetry factor $k = \beta_N/\beta_I$ as the ratio of the amplification factor β_N in the normal common-emitter connexion and the amplification factor β_I in the inverse common-collector connexion, given usually at the same base current. For determining k according to definition, it is necessary to know both the amplification factors, which in themselves suppose measurements at least of two currents (if one is satisfied with a smaller accuracy and neglects the saturation currents of both the emitter and collector junctions), since $\beta_N \simeq I_C/I_B$, $\beta_I \simeq I_E/I_B$, and hence

$$k \simeq [I_C/I_E] \quad I_B = \text{const} \; .$$

This is obviously inconvenient in the measurement and checking of transistors on an industrial scale because of its slowness.

In a more suitable method, whose principle is evident from the arrangement shown in Fig. 1, it is possible to determine the symmetry factor directly in a single measuring operation. In accordance with Fig. 1, the transistor VT to be tested is connected in a bridge circuit whose two branches are formed by a fixed resistor R_C and a variable resistor R_E, whereas the residual 'branches' are represented by the collector and emitter junctions of the transistor tested. The bridge is energized by a source of direct

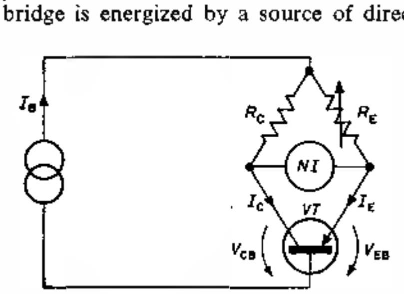

Fig. 1. The bridge measuring arrangement

current I_B (also the magnitude of the base current, at which the symmetry is determined), and is balanced out by adjusting the resistor R_E. As will be shown, at the balance (indicated by the null-indicator NI connected to the collector and emitter leads), the following equation is satisfied:

$$R_E/R_C = k \, . \quad \text{........... (1)}$$

With R_C fixed, it is then possible to read the symmetry factor directly on the scale of the calibrated resistor R_E.

To prove equation (1), the condition for the balance of the bridge will be determined. At balance, the collector and emitter are at equal potential and the current I_B will divide in both the resistors by the ratios

$$I_C = I_B \, \frac{R_E}{R_C + R_E} \text{ and}$$

$$I_E = I_B \, \frac{R_C}{R_C + R_E} \quad \text{...... (2)}$$

In accordance with the Ebers-Moll[2] model of the junction transistor, the voltages at the collector and emitter junctions are:

$$V_{CB} = V_T \ln \left(1 + \frac{I_C + \alpha_N I_B}{I_{CO}} \right) \text{ and}$$

$$V_{EB} = V_T \ln \left(1 + \frac{I_E + \alpha_I I_C}{I_{EO}} \right),$$

where V_T is the thermal voltage, α_N and α_I are the normal and inverse amplification factors in the common base connexion, and I_{CO} and I_{EO} are the reverse saturated currents of the collector and emitter junctions, respectively. The two voltages are equal, when

$$\frac{I_C + \alpha_N I_E}{I_{CO}} = \frac{I_E + \alpha_I I_C}{I_{EO}}$$

Using equation (2) and with respect to the reciprocity relation $\alpha_N I_{EO} = \alpha_I I_{CO}$, it follows from the last equation

$$\frac{R_E}{R_C} = \frac{\alpha_N (1 - \alpha_I)}{\alpha_I (1 - \alpha_N)}$$

which is in agreement with the basic equation (1). Following from this equation, note that the balancing of the bridge is independent of the magnitude of the base current I_B. Nevertheless, this is valid only with an ideal transistor with constant amplification factors. In the real case, it is necessary to respect their current and thermal dependence and to show the symmetry factor at some conventional base current and some ambient temperature.

With regard to the accuracy of the method suggested, if the balancing is indicated with an error of dV_{CE}, then the corresponding relative error in determining the symmetry factor is

$$\frac{dk}{k} = \frac{k+1}{k} \cdot \beta_N \frac{dV_{CE}}{V_T} \quad \ldots\ldots (3)$$

Thus for $\beta_N = 100$, $V_T = 26$mV at 25°C and $|dV_{CE}$K1μV, it follows from equation (3), $|dk/k| < 1$ per cent for symmetrical transistors ($k = 1$) and $|dk/k| < 0{\cdot}5$ per cent for typical unsymmetrical transistors with $k = 10$. The accuracy is limited by thermoelectric voltages in the null-indicator circuit, the main source of which being the collector and emitter joints.

Yours faithfully,

J. DOSTÁL.

Res. Inst. for Math. Mach.,
Prague.

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Some Operational Aspects of Diode Phase Sensitive Detectors

DEAR SIR,—A diode phase-sensitive detector was discussed by Leela & Sankar[1]. In this the authors discussed the percentage non-linearities of the detector. Table 1, of the text, reproduced here, illustrates different values of the percentage non-linearities for various values of n_1, and n_3, where n_1 denotes the source impedance factor (inversely proportional to the source impedance), and n_2 is the load factor (inversely proportional to the output impedance).

On critical examination of the table we find some new correlations which are reported here. Referring to Table 1, the circled values indicate the trend in the percentage non-linearities which we find are minimum as indicated by the direction of the arrow. From this it is observed that:

(1) The percentage non-linearities at different values of x reduces at

(a) lower values of n_1

(b) lower values of n_2

For a given value of the source impedance factor n_1, the percentage non-linearity increases with the load factor. This would mean that a push-pull type of phase sensitive detector will have smaller non-linearities at the open-circuit output terminals (n_2 minimum) and larger non-linearities at the short-circuit output terminals (n_2 maximum).

(2) There occurs a maximum of percentage non-linearity for a particular value of n_1 keeping n_2 constant (see Fig. 1). If the maximum non-linearity is determined for various values of n_1 and n_2, then we find that these lie on a straight line; as indicated by the value within the squares in Table 1.

From this one concludes that from an operational point of view the values of n_1 and n_2, corresponding to the straight line plot, should be avoided.

(3) For continuous operation lower values of source impedance factor may be chosen. This is necessary because the spread in the percentage non-linearities, for a given range of n_2, are smaller at these values than at higher values of source impedance factor. This is clearly brought out in Fig. 2.

TABLE 1*

The percentage non-linearity for the values of x equal to 0·25 x_c; 0·5 x_c; 0·75 x_c; 0·9 x_c and x_c for various values of n_1 and n_2

n_1 \ n_2	100	30	10	3	1	0·3	0·1	0·03	0·01
1	0·1 0·5 1·4 2·8 5·2	0·1 0·5 1·4 2·7 5·1	0·1 0·4 1·3 2·6 4·9	0·1 0·3 1·0 2·1 4·3	— 0·2 0·6 1·4 3·2	— 0·1 0·2 0·5 1·7	— — — 0·1 0·7	— — — — 0·2	— — — — 0·1
3	0·3 1·2 3·1 5·5 8·6	0·3 1·1 3·0 5·4 8·5	0·2 1·0 2·8 5·0 8·0	0·2 0·8 2·2 4·0 6·8	0·1 0·4 1·3 2·5 4·7	— 0·1 0·4 0·9 2·3	— — 0·1 0·2 0·9	— — — — 0·3	— — — — 0·1
10	0·4 1·9 4·8 8·0 11·8	0·4 1·8 4·6 7·7 11·4	0·4 1·6 4·0 6·9 10·4	0·2 1·1 2·9 5·0 8·0	0·1 0·5 1·5 2·7 4·9	— 0·1 0·4 0·9 2·1	— — 0·1 0·2 0·8	— — — — 0·3	— — — — 0·1
30	0·5 2·3 5·7 9·3 13·5	0·5 2·1 5·2 8·6 12·6	0·4 1·6 4·2 7·1 10·8	0·2 0·9 2·5 4·4 7·2	0·1 0·4 1·1 2·0 3·8	— 0·1 0·3 0·6 1·5	— — 0·1 0·1 0·5	— — — — 0·2	— — — — 0·1
100	0·6 2·4 6·0 9·7 14·1	0·5 2·0 5·0 8·3 12·4	0·3 1·3 3·5 5·9 9·3	0·1 0·6 1·6 2·9 5·0	— 0·2 0·6 1·2 2·3	— — 0·2 0·3 0·8	— — — 0·1 0·3	— — — — 0·1	— — — — —
300	0·5 2·3 5·7 9·4 13·7	0·4 1·6 4·2 7·1 10·9	0·2 0·9 2·4 4·1 6·9	0·1 0·3 0·9 1·7 3·0	— 0·1 0·3 0·6 1·2	— — 0·1 0·2 0·4	— — — — 0·1	— — — — —	— — — — —
1000	0·4 1·9 4·9 8·2 12·3	0·3 1·1 2·9 5·0 8·1	0·1 0·5 1·4 2·4 4·2	— 0·2 0·5 0·8 1·6	— 0·1 0·2 0·3 0·6	— — — 0·1 0·2	— — — — 0·1	— — — — —	— — — — —

* *From Leela and Sankar*[1]

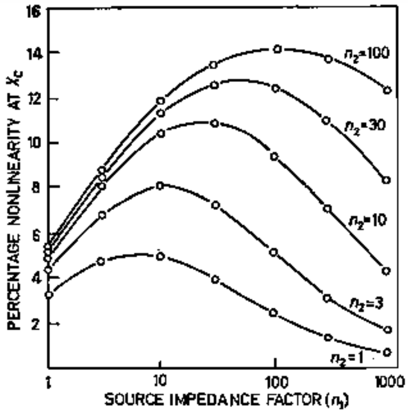

Fig. 1. Percentage non-linearity variation with n_1

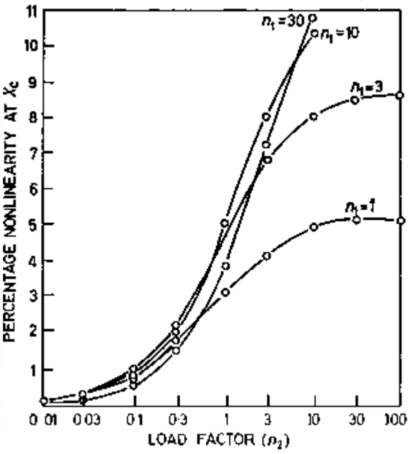

Fig. 2. Percentage non-linearity variation with n_2

From the above discussion it is obvious that a high input impedance device is indicated at the output terminals.

It is interesting to compare these results with those obtained for single ended-phase sensitive detector, by Chidambaram & Krishnan[2]. Here we observe that the percentage non-linearity increases for lower values of n_2 rather than decreasing. Also there is, an occurrence of maximum in the percentage non-linearity, which is independent of the value of n_2, but is mainly influenced by the particular value of n_1 as is evinced from the results reported by Chidambaram & Krishnan[2].

Yours faithfully,

P. K. PATWARDHAN,

V. S. INDURKAR,

Electronics Division,
Atomic Energy Establishment Trombay,
Bombay, India.

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Information and Communication Theory

By A. M. Rosie. 175 pp. Demy 8vo. Blackie & Sons Ltd. 1965. Price 32s. 6d.

THE theory described in this book is based mainly on the line taken by Shannon in his famous papers of 1948 and 1949. Since the appearance of these papers interest in the theory has been very great and development has been fairly rapid, and the subject is now being taught at undergraduate level in several universities in the United Kingdom.

The book under review is suitable for final honours year students as well as for postgraduate students; in fact, since the mathematical requirements are not very great some universities could use the book in, say, the second year of the finals course.

There are eight chapters, the first two being an introduction to the subject and elementary probability theory in which the various probability distribution are discussed. Then comes a chapter on the properties of signals where, among other things, the relation between power spectrum and autocorrelation function is dealt with, and the sampling theorem is stated and proved. The fourth is an interesting chapter on the properties of sources and channels in which the concept of entropy is introduced (Shannon's definition). After a chapter on ideal communication systems comes one on the use of binary coding to combat error introduced by noise followed by one on communication systems in the light of information theory. The final chapter discusses the application of information theory in other fields, namely psychology, biology and neurophysiology.

The overall impression made by the book on the reviewer is a good one although the earlier chapters seemed less readable. In particular the chapter on probability theory was not particularly persuasive and was hampered slightly by careless punctuation. Again although pedantic rigour is out of place in a book of this kind the common practice of introducing the Fourier integral as a limiting case of a Fourier series on the grounds that "it gives a good insight into the physical meaning of the relationship" is very questionable, especially since a proof readily acceptable to undergraduates of the Fourier integral theorem can be given (via Dirichlet's integral) without mentioning periodicity and, in particular, without introducing the rather loose idea of identifying a function of indefinite periodicity with an aperiodic function.

Apart from these rather minor criticisms the reviewer found the book interesting, well produced with clear diagrams and with many useful references.

R. H. C. NEWTON

Principles and Applications of Boolean Algebra for Electronic Engineers

By S. A. Adelfio and C. E. Nolan. 326 pp. Demy 8vo. Iliffe Books Ltd. 1965. Price 45s.

THIS book has two major parts, devoted respectively to 'Boolean Algebra' (Chapters 1-6) and to 'Logic Circuits and Electronic Counters' (Chapters 9-11). Chapters 7 and 8 were written to provide a link and will be considered first. Chapter 7 is entitled 'Basic Concepts of D.C. Principles' and deals with elementary passive circuit principles, embracing Thevenin's, Norton's and Millman's Theorems. It is unfortunate that the convention is chosen that the positive direction of current is chosen as flowing away from the negative source terminal: this can only cause confusion. In the introduction to Chapter 8, which is entitled 'Basic Concepts of Diodes and Transistors' it is stated that 'If the reader is familiar with diodes and transistors he can pass over this chapter and proceed to chapter 9.' In the reviewer's opinion this advice should also be taken by readers who are not familiar with diodes and transistors, since the physics is weak and the treatment is all too brief.

Despite the above criticisms there is no doubt that the book as a whole will be useful to electronic engineers. To the reader who is not already familiar with Boolean algebra the first chapters provide a very readable introduction. Chapter 6 is devoted to 'Problem Solving' and makes use of the background provided in the earlier chapters. The latter part of the book shows the implementation of the logic principles in practical circuits incorporating semiconductor diodes and bipolar transistors.

The presentation is in general very good and the number of typographical errors small. At 45/- the book is quite good value.

F. J. HYDE

Advanced Electrical Engineering

By A. H. Morton. 514 pp. Demy 8vo. Sir Isaac Pitman & Sons Ltd. 1966. Price 65s.

NORMALLY one would be very sceptical of a text book on advanced electrical enginering in which almost as much space is devoted to the list of contents as to the opening section on matrices. However, although the range of subject matter covered in Mr. A. H. Morton's new book is extensive the reviewer has few criticisms, mainly because of the excellent treatment each section is afforded.

The first part of the book can be loosely classified as circuit theory. The section on matrices is a little superficial and the omission of partitioned matrices is surprising. The chapters on two port networks and complexor locus diagrams are very well presented and there is a reasonable treatment of electrical transients in networks. There is a short but useful look into feedback systems and a fairly lengthy attack on transmission lines. The basic concepts of communication theory and symmetrical components are given in the last two chapters, but it is obvious that twenty pages cannot do absolute justice to a discipline as wide as communications.

The second part of the book covers electric and magnetic fields, time varying magnetic fields, and electron ballistics. The content is standard undergraduate material which is well covered in several well known texts. However, the author deals with the fields aspects in a clear and concise way and is well justified in including it.

The third section is devoted to measurements and there are chapters on limits and standards, single phase measuring circuits and instrument theory. This section is extremely brief and no doubt included solely to cover the third part I.E.E. Advanced Electrical Engineering syllabus.

Unlike many recent text books this one is written with the student in mind, and in particular it is to be recommended to students preparing for the third part I.E.E. examination. It is felt that this book, together with the author's previous text*, may well cover the common core material of most undergraduate courses. Each chapter is well supplemented by questions with answers, and a brief bibliography.

R. W. WHITEHEAD

* MORTON, A. H. SHEPHERD, J., SPENCE, L. F. Higher Electrical Engineering (Pitman).

Transistoren Physikalische Grundlagen und Eigenschaften (Transistors. Physical Bases and Properties)

(In German). By R. Paul. 508 pp. Royal 8vo. VEB Verlag Technik, Berlin. 1964.

NOTWITHSTANDING its sub-title, "Physical Bases and Properties", this book is aimed primarily at electronic engineers, particularly those interested in the application of transistors to electronic circuits, though engineers interested in the development and design of transistors themselves will also find much pertinent information. The book starts with a discussion of the conduction mechanism in semiconductors (Chapter 1) and of disturbed thermodynamic equilibrium (Chapter 2). Chapter 3 gives a particularly clear and full treatment of pn-junctions. Further chapters in the first part treat with diffusion- and drift-transistors. A rather brief chapter, 24 pages out of the total of 500, deals with the techniques of crystal-growing and junction-preparation, and with the various types of transistors in practical use today, including epitaxial and planar types.

The main part of the book discusses the technical properties and the various electrical characteristics of transistors as well as their different equivalent circuits. The last few chapters deal with pulse operation, switching characteristics, noise, heat dissipation and thermal stability. These subjects are treated in very great detail and the book forms a veritable compendium. To compress this large amount of information into 500 pages, much of it had to be set in small print. Even so, unipolar, thin-film and other types of majority-carrier transistors are omitted as being still in the stages of rapid development.

The comprehensive bibliography of nearly 2 000 references is up to date to the end of 1963. Their arrangement is somewhat puzzling, being neither by date, by author or by sequence of their appearance in the book. There is no author index and only a very brief subject index covering two pages. The book will probably serve its purpose well and will become a standard text in Eastern Germany. It is less likely to be of similar benefit to readers here, who have ready access to the literature on the subject in the English language.

E. BILLIG

Linear Systems

By R. J. Schwartz and B. Friedland. 521 pp. Royal 8vo. McGraw-Hill Book Co. 1965. Price £5

THE scope of this book is wide. The linear systems covered include: lumped-parameter continuous dynamical systems (time-invariant and time varying systems, represented by ordinary differential equations with constant and time-dependent coefficients respectively); analogous discrete systems (sampled, pulse-modulated and digitally controlled systems, etc., represented by difference equations); the treatment of random signals and noise (continuous and discrete); with a final chapter on distributed parameter systems. This field is treated in a total of thirteen chapters.

The concepts of the different varieties of dynamical systems are introduced by way of flow-diagrams and equivalent analogue computer set-ups. The mathematical treatment of linear systems proceeds by way of a reduction of higher order equations for the system to a normal system of first order equations in the 'state variables'; the differential equations of continuous systems and the analogous difference equations of discrete systems are treated in parallel—indeed the general development covers time-varying (coefficients a function of time) as well as constant coefficient equations, though there is little consideration of the mathematical complexities and the applications of the former. After introducing pulse responses (step functions, delta functions, etc) and the superposition of solutions of linear systems, the general formal treatment in the time-domain proceeds by way of matrix treatment of the first order system of normal equations in the state variables, based on the fundamental matrix of the homogeneous equations. The four following chapters (Chapters 5 to 8) study the practical solution of differential equations by means of Fourier and Laplace transform methods, and differential equations by the analogous Z transform. Two following chapters deal with the statistical basis and mathematical treatment of random signals and the problems of filters for linear systems subject to random noise. A further two chapters are devoted to stability and criteria for the determination thereof, while the final chapter selects some topics (mainly) from transmission line theory as an illustration of linear distributed parameter systems.

The foregoing review of the range of this book indicates something of its comprehensiveness. When one comes to consider its readability, and to whom it is meant to appeal it is, unfortunately, necessary to be more critical. The dust-jacket states it has been used as a course for senior undergraduates and first year graduate students at Columbia. The reviewer found the great generality in which every topic was introduced—dealing simultaneously with time-varying, constant coefficient and discrete systems—made very heavy going for the reader, not helped by such technical 'gobbledygook' as ". . . causal multiple-input, -output (possibly time-varying) system" (p. 121); he feels this regrettable in a subject which can make fascinating reading when lucidly developed. More serious, however, are the mistakes and evidence of lack of care in imprecise or inadequate presentation of new concepts (the lone student, coming upon Fourier series and transforms for the first time, would find difficulty in marrying Fig. 5.1 to the descriptive text). When efforts are made to present mathematical treatments in very general form, it behoves the writer to be most careful in defining the restrictions to his generalization—the domain in which it holds; this the authors have not always done. The statement in the Preface: "The presentation is intended to be intuitively appealing . . . ." [*rather than rigorous*?] does not licence the authors to equate a function of two variables to a function of one variable (top of p. 95) without qualification!

Although well printed, and provided with numerous problems after every chapter, for reasons given in the preceding paragraph, this work cannot be recommended as an introductory text for the student. Perhaps for the teacher, conscious of its shortcomings, it might well serve as the basis of a course for the more sophisticated student.

J. G. L. MICHEL

Dictionary of Semiconductor Physics and Electronics English-German—German-English

Edited by W. Bindman. 615 pp. Demy 8vo. Pergamon Press. 1966. Price £8

Each section of this dictionary contains some 25 000 terms relating to semiconductor physics and electronics.

Symbols and abbreviations have deliberately been dispensed with and replaced by short explanations.

Electronics Through Experiments Volume 1: Components

By E. T. Glas. 263 pp. Med. 8vo. Philips Technical Library. 1966. Price 32s.

This book is intended for use in technical schools and colleges, and also for self-tuition.

Part I deals with valves, transistors, resistors and other components together with their fundamental properties and the layout of simple circuits.

Part II covers the principles and circuit techniques related to telecommunications.

Electronics for Experimenters in Chemistry, Physics and Biology

By L. F. Phillips. 266 pp. Demy 8vo. John Wiley & Sons Ltd. 1966. Price 45s.

The aim of this book is to provide design details of electronic circuits from first principles for use by the experimenter in chemistry, physics and biology.

Valve and transistor circuits are discussed showing their similarities, and a simple treatment of transistor circuits is given, avoiding the more complicated mathematics.

Correlation Equations for Statistical Computations

By A. K. Mitropol'skii. 103 pp. Demy 4to. Consultants Bureau Enterprises Inc. 1966. Price $9.50

This book presents the methods for computing correlation equations and is based on part of the author's book Tekhnika Statisticheskikh Vychislenii.

Electrical Instruments in Hazardous Locations

By E. C. Magison. 225 pp. Med. 8vo. Plenum Press. 1966. Price $22.50

The safe economical use of electric instruments in hazardous locations is discussed together with the nature of electrical hazards and the methods by which they can be reduced.

The book is based on the work done by the Instrument Society of America's Recommended Practice Committee 8D-RP12 "Instruments for Hazardous Locations".

The Silicon Gate-Controlled Switch

Edited by G. J. Pridham. 52 pp. Demy 8vo. Sir Isaac Pitman. 1966. Price 12s. 6d.

This small book covers the five papers presented at a symposium held at the Enfield College of Technology in December 1964.

Advances in Automatic Control

328 pp. Demy 4to. The Institution of Mechanical Engineers. 1965. Price £6 10s.

This book contains the papers presented at the Convention on Advances in Automatic Control held at the University of Nottingham in April 1965 and sponsored by the Institution of Mechanical Engineers with the support of the United Kingdom Automation Council.

NEW EQUIPMENT

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

(Voir page 483 pour la traduction en français; Deutsche Übersetzung Seite 490)

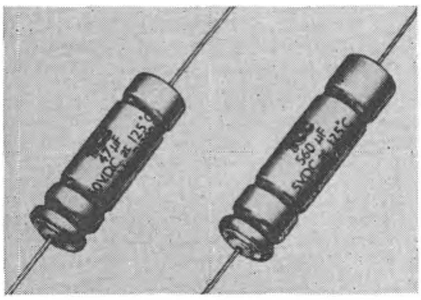
ELECTROLYTIC CAPACITORS

The Telegraph Condenser Co. Ltd, Wales Farm Road, North Acton, London, W.3

(Illustrated below)

A new range of aluminium foil electrolytics recently introduced by the Telegraph Condenser Co Ltd.—a company in the Plessey Components Group—will operate satisfactorily over the temperature range of −55°C to +125°C.

These TCC E.T.R. (Extended Temperature Range) capacitors are being manufactured in the U.K. under an exclusive Canadian Government licence, in a range of capacitances and sizes conforming to the U.S. Mil-C-39018 specification. A comparable British Defence Specification is in preparation and capacitors will be submitted for qualification approval.



E.T.R. capacitors are about one third of the weight of equivalent tantalum foil types and have a larger capacitance per unit volume. The electrical performance does in fact approach that of tantalum capacitors and they will meet a similar specification. The drop in capacitance at −55°C is small and the electrical characteristics remain very stable during life tests in excess of 2 000 hours at 125°C.

Idling characteristics are far superior, and equivalent series resistance much lower, than for ordinary aluminium electrolytic capacitors over a wide temperature range. This makes the capacitors more suitable for operation in pulse circuits where the pulse shape must be maintained.

Much of the improvement in performance is attributable to the use of a new type of electrolyte containing di-methyl formamide as the solvent, which in turn required a new type of construction and sealing.

The tubular cases are provided with extra rilling to limit the movement of the internal element in order to meet vibration requirements. Another interesting feature of the case is a weakened area designed to ensure that in the event of a high internal gas pressure being generated—such as by accidental reversed connexion of a polar unit—the case will burst non-explosively.

Life tests of E.T.R. capacitors at rated voltages and maximum temperature have shown remarkable stability of parameters; percentage changes in capacitance, power factor, and leakage current after 2 000 hours have been found considerably below Mil-C-39018 draft limits.

EE 95 751 *for further details*

ULTRA-VIOLET OSCILLOGRAPH

Southern Instruments Ltd, Camberley, Surrey

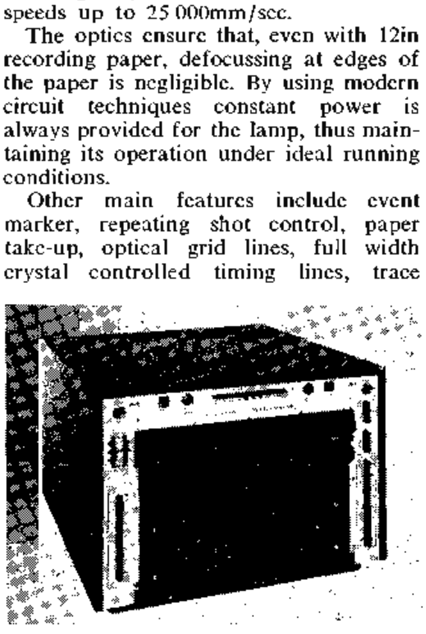
(Illustrated below)

The latest addition to the range of Southern Instruments ultra-violet oscillographs is the M1310. This instrument is of very rigid construction and provides a wide range of built-in features.

The standard instrument which is available in bench or rack mounting form, can accommodate from 1 to 50 channels and records on any paper width up to 12in. Twelve paper speeds, operated by bush-button, give a range from 4 to 2 500mm/sec. This can be extended to provide 24 speeds in the range 0·004 to 2 500mm/sec or with the ultra-high speed unit a further four speeds up to 25 000mm/sec.

The optics ensure that, even with 12in recording paper, defocussing at edges of the paper is negligible. By using modern circuit techniques constant power is always provided for the lamp, thus maintaining its operation under ideal running conditions.

Other main features include event marker, repeating shot control, paper take-up, optical grid lines, full width crystal controlled timing lines, trace identification, trace and shot numbering and remote control.



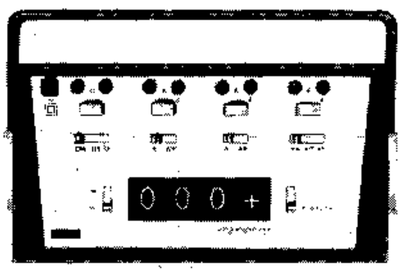
EE 95 752 *for further details*

MULTIMETER

Distributed by: Kelvin Electronics Co. (Smiths Industries Ltd), Kelvin House, Wembley Park Drive, Wembley, Middlesex

(Illustrated below)

The 'DigiKelvo' is a multimeter giving polarity and a three-figure digital display of the measured value on number tubes. The instrument has four separate input channels covering a.c. and d.c. voltage, current and resistance. Two channels are used for voltage measurements from 1mV to 1kV in six ranges. The third channel is used for measuring current from 100nA to 999mA in seven ranges and, with the use of shunt, up to 99·9A in two ranges. The fourth channel is used for measuring resistance from 1Ω to 999MΩ in seven ranges. Automatic range changing is achieved by moving a decimal point across the display. In addition to this, there is a two position course range switch, thus enabling six orders of magnitude to be covered.



The system consists of a high performance, low drift d.c. amplifier, whose output is backed off by a staircase circuit, and a counter to display the number of increments required to reach null balance. The principles used in the staircase circuit and counting circuits are conventional, but the high performance d.c. amplifier is unusual in many respects.

The d.c. amplifier has a short-term drift referred to the input of 0·5μV/°C, and an estimated long term drift of 2·5μV per year. The open-loop gain of the amplifier is 2×10^6, which, in the highest gain configuration, is reduced by feedback to 14 000. The bandwidth of the amplifier is 6000kc/s (although other features in the instrument would reduce the overall bandwidth to 100kc/s). This bandwidth gives a maximum measuring period of the instrument of 0·3sec. The linearity of the amplifier is better than 0·1 per cent, giving an accuracy of the instrument of ±1 digit in the last figure. Many of these characteristics are attributable to the use of a f.e.t. as a chopper, at the input of the amplifier.

A further feature of the 'DigiKelvo' is that internal wiring has been completely eliminated by the use of printed circuit boards and suitable connectors.

EE 95 753 *for further details*

R. F. INTERFERENCE SUPPRESSORS

Erie Resistor Ltd, South Denes, Great Yarmouth, Norfolk

Erie Resistor Ltd announce that two units out of their wide range of 1200 series R.F.I. broadband filters have gained Ministry of Aviation Approval for radio frequency interference suppression applications in military aircraft systems. The two units are designated 1200-025 and 1215-000.

The 1200 series R.F.I. broadband filters comprise a 1·5μF ceramic capacitor (the subject of a special manufacturing technique patented by Erie) and ferrite beads, to produce the series inductive effect (resistive losses at r.f.). The unit is sealed in a robust metal housing with a threaded bush section. The mechanical configuration is feedthrough, with a choice of terminations. For its performance and price the Erie R.F.I. broadband filter is claimed to be the smallest single unit commercially available in the world.

The 1200 series are 'broadband' filters, in that their operating frequency, for useful attenuation, is over the range of about 5kc/s to 1Gc/s, and even higher. Maximum attenuation is in the direction of inductive series input, since these are essentially L-section filters. Minimum attenuation, including any resonance dips that might occur, is 70dB over most of the high frequency spectrum. Continuous working voltage is 50V d.c., with approximately 50 per cent derating for 125°C operation: the operating conditions of B.S. 2G.100 are met.

EE 95 754 *for further details*

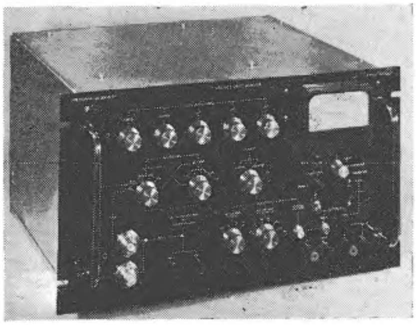
VOLTAGE DRIFT MONITOR

J. & P. Engineering (Reading) Ltd, Portman House, Cardiff Road, Reading, Berkshire

(Illustrated above right)

This drift monitor consists of a voltage reference, divider network and deviation amplifier mounted in a constant temperature enclosure. It is primarily intended for measuring the drift of power supplies in the range 0·1V to 5kV, and 100μA to 5A. For this purpose stability and sensitivity are of paramount importance.

The drift monitor contains three stages of temperature control, resulting in an overall temperature coefficient better than $0{\cdot}2 \times 10^{-6}/°C$ and a stability better



than one part per million against a 10 per cent mains change. The reference voltage is a temperature compensated Zener diode driven from a very stable current generator. The most significant divider resistors are 'loose wound' for long term stability, have a very low temperature coefficient and are mounted in a constant temperature enclosure. The air in this enclosure is circulated by a powerful fan so that the temperature rise of the measuring resistors when an input is connected is small.

The drift monitor possesses sufficient stability and sensitivity to record drifts of a few parts per million over several days, and is virtually independent of mains and temperature variations. It is entirely transistorized and is in a 19in rack mounting box with a 10½in high front panel. It was designed by J. & P. Engineering (Reading) Ltd for testing voltage and current supplies at A.E.R.E. Harwell.

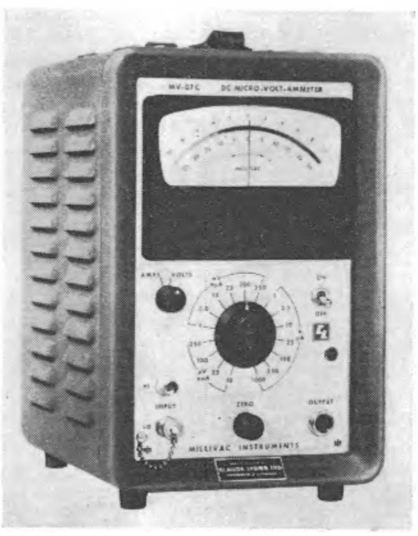
EE 95 755 *for further details*

D.C. MICRO-VOLT-AMMETER

Claude Lyons Ltd, Valley Works, Hoddesdon, Hertfordshire

(Illustrated below)

The MV-07C is a sensitive electronic voltmeter and ammeter, capable of measuring as low as 1μV and 1nA d.c. A total of 34 voltage and current ranges, in a 10-25-100 sequence, extend the measuring capability to 1kV and 1mA full scale. Accuracy is 1 per cent of full scale on most voltage ranges up to and including the 1kV range.



Features of the instrument include a low-noise chopper amplifier, careful design of the input circuit to avoid dissimilar metal connexions, individual calibration controls for all 17 voltage ranges and an individually calibrated, centre-zero 6in mirror scale galvanometer with taut band suspension.

The MV-07C employs a sensitive low-noise carrier amplifier incorporating a low-noise, low drift chopper followed by a low-noise cascode first stage. To increase signal-to-noise ratio, this amplifier has a narrow bandwidth, obtained by the incorporation of a twin-T circuit in one of the a.c. feedback loops. Heavy overall a.c. feedback stabilizes the carrier amplifier against mains voltage variations and changes in component characteristics.

A transistorized synchronous demodulator produces a d.c. output which is proportional to the input voltage or current and which is in phase with the input, enabling a centre-zero to read both positive and negative inputs without a polarity switch. The output is also taken to a connector to provide an amplified in-phase signal for external use.

Excellent gain stability is ensured by an overall adjustable d.c. feedback loop which passes through the range switch and also incorporates individual calibration controls for all 17 voltage ranges and for the six most sensitive current ranges.

This instrument is manufactured in England by Claude Lyons Ltd under an exclusive manufacturing licence from Millivac Instruments Inc., U.S.A.

EE 95 756 *for further details*

VALVE-VOLTMETER

Advance Electronics Ltd, Roebuck Road, Hainault, Ilford, Essex

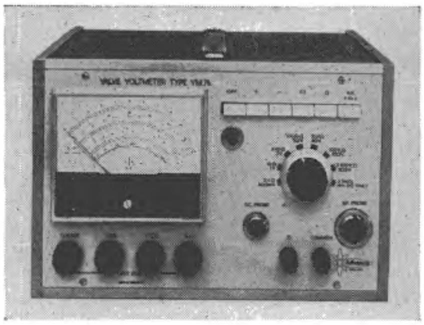
(Illustrated on page 476)

The latest in the Advance range of five voltmeters is the VM76 which extends the range to include a.c. frequencies in excess of 1000Mc/s and d.c. measurements up to 1000V with accuracies of ±2 per cent up to 100V.

Push-button selection of the various functions is a feature of this instrument. In d.c. operation this method can be used to select centre-zero readings and either positive or negative polarity. Resistance measurements between 0·02Ω and 500MΩ to an accuracy of ±5 per cent at midscale can be selected in a similar way.

Good component accessibility is provided by mounting the majority of components on a single printed circuit board. Each instrument is provided with probes and connecting leads with single plug and socket connexions. Optional accessories for all voltmeter applications are available.

The VM76 will operate in the range 0°C to +40°C and is mains operated. In line with other recently introduced Advance instruments it is built into a robust modular case which is ½ standard rack width.



The d.c. ranges cover 300mV to 1kV f.s.d. with an input impedance greater than 100MΩ. The a.c. ranges cover 30mV to 300V v.m.s. The input impedance is of the order of 7·5MΩ at 1kc/s and 150kΩ at 100 Mc/s.

EE 95 757 *for further details*

D.C. AMPLIFIERS

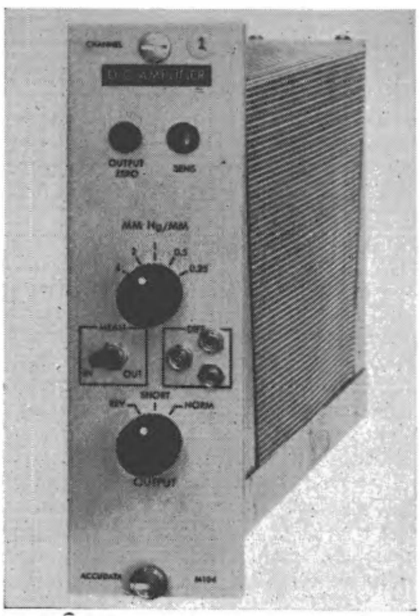
Honeywell Controls Ltd. Honeywell House, Great West Road, Brentford, Middlesex

(Illustrated below)

Honeywell is to market for the first time in Britain a series of d.c. amplifiers. Known as the Accudata 104, the range is available in a standard version, giving a gain between 10 and 250, and two other versions giving gains of 0·1 to 2·5 and 0·4 to 10.

The amplifiers have excellent noise and drift specifications and are general purpose amplifiers for d.c. excited strain gauges, thermocouples or other signal sources lacking sufficient power to drive a recorder direct. All versions are designed to work with galvanometer recorders and have sufficient output capability to drive even the highest frequency fluid damped galvanometers.

The Accudata 104 is designed to prevent transients from damaging the galvanometer when gain ranges are switched, when the input is open-circuited, or when the unit is turned on or off. A front panel control determines the polarity of connexion to the galvanometer and also enables the output to be shorted for zero setting of the galvanometer.



EE 95 758 *for further details*

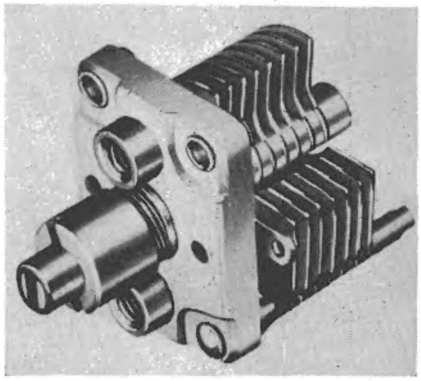
VARIABLE CAPACITORS

Wingrove & Rogers Ltd. Domville Road, Liverpool, 13

(Illustrated below)

The C.903 is an air dielectric trimmer fitted with a locking device. The ceramic base has an overall size of 0·8in × 0·6in and is fitted with two tapped 6 B.A. bushes for mounting the trimmer. This is a similar but smaller trimmer than the C.803 type.

The minimum capacitance is 3·5pF and the capacitance swing up to 34pF.



The power factor is less than ·001 at 1Mc/s and the insulation resistance is greater than 1000MΩ at 500V d.c.

EE 95 759 *for further details*

DISK STORAGE DRIVES

Distributed by: Scientific Furnishings Ltd. Poynton, Cheshire

(Illustrated above right)

The Control Data 9430 series disk storage drives combine the desirable features of magnetic tape and magnetic disk recording—fast random access and unlimited shelf storage. This is made possible by removable disk packs which may be recorded on, stored indefinitely (like magnetic tape reels) and rapidly re-inserted on-line. All disk packs are completely interchangeable among disk storage drives of the same model number; disks recorded on one unit may be read on other units. The 9430 and 9434 are also compatible with the IBM models 1311 and 2311 disk storage drives respectively.

The 9430 series all have direct seek capabilities—the last address used is retained by the disk storage drive and is available to the computer to determine the distance and direction of the next head movement. This provides considerably faster access times than the return-to-home method. Seek overlap, an additional standard feature on all models, allows simultaneous seek operations to be performed by more than one unit.

Four models are available for 6, 7 and 8 bit characters, with total capacities (6 disks per pack) between 2 000 000 and 8 192 000 characters. Maximum access time on all models is 145msec, average time being 70msec.

EE 95 760 *for further details*

WHEATSTONE COMPARATOR BRIDGE

British Physical Laboratories, Radlett, Hertfordshire

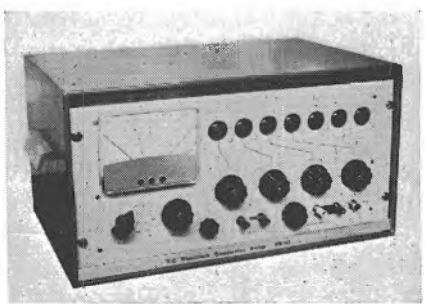
(Illustrated below)

The B.P.L. model RM 197 d.c. low-load Wheatstone comparator bridge has been developed as a result of the company's experience of low-load Wheatstone bridges, particularly the RM 196 bridge. Model RM 197 is a self-contaiained, dual purpose, instrument for measuring resistance. It consists basically of the Wheatstone bridge components comprising ratio arms and a variable arm, a bridge supply and a bridge detector.

The variable arm comprises four decades of discrete steps, its value being displayed on digital read-out indicators. The ratio arms supply the decimal point to give the magnitude of the four significant figures.

The bridge supply is constant current rather than constant voltage. This arrangement gives a sensibly constant error voltage for a certain percentage unbalance of the bridge which is desirable when used in the comparator mode.

The bridge detector is a low noise high gain chopper stabilized amplifier, the output being synchronously rectified to obtain positive and negative unbalance indications.



When in use as a Wheatstone bridge the unknown resistance is connected to the *Rx* terminals, the variable set to some arbitrary value (not 0000); the bridge is then balanced by using first the range switch, then the most significant digit. Maximum sensitivity is obtained by using the 0·5 per cent position of the f.s.d. switch. The unknown resistance value is read directly from the indicators.

When the instrument is being used in the comparator mode, the resistance value is set-up on the indicators using the range switch and decade switches. The f.s.d. switch is set to the desired full scale deflexion range, e.g. if grading 10 per cent resistors use 20 per cent f.s.d. range. The resistors are then applied to the *Rx* terminals and the meter will read directly the percentage deviation of the resistor from the set-up value.

EE 95 761 *for further details*

CHOPPER-AMPLIFIER

Kent Precision Electronics Ltd, Vale Road, Tonbridge, Kent

(Illustrated below)

The K-PAC/4 has been developed for the accurate operation of standard instrumentation with inputs in the range 0 to 100mV, from low output transducers, thermocouples etc. having inputs in the region 0 to 10mV. It consists of a low drift chopper amplifier developed by KPE and built with a self-contained power supply within the standard K-PAC construction. An approved type transistor chopper modulates the incoming millivolt signal which is then fed into a stable-gain a.c. amplifier, the output of which is de-modulated to provide a d.c. output. Feedback is then included from output to input to stabilize the gain at the value required to convert the input signal to a standard 0 to 100mV d.c. output for operation of other instruments. The gain is set within a maximum of 20 and the unit has a drift performance of less than 3μV/°C. For special instruments, the K-PAC/4A using a field-effect transistor chopper, is available for use with inputs in the region 0 to 1mV with a drift performance of within 0·5μV/°C.

The instrument is operated from an auxiliary mains a.c. supply and has an internal transformer so that the circuit is, in all cases, isolated from the mains. The unit may be supplied to have maximum or minimum voltage at open-circuit of the input terminals, so that the associated instrumentation may 'fall-to-safety'.

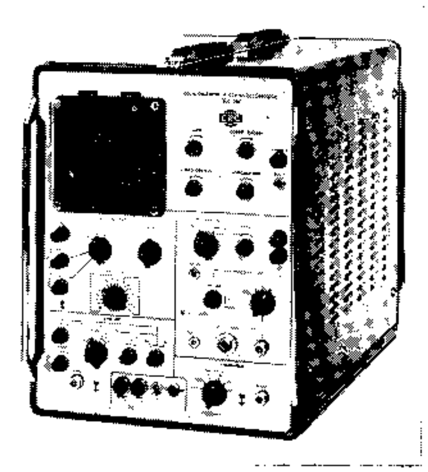
EE 95 762 *for further details*

SAMPLING OSCILLOSCOPE

Distributed by: Claude Lyons Ltd, Valley Works, Hoddesdon, Hertfordshire

(Illustrated below)

Now available in the U.K. through the Instruments Division of Claude Lyons Ltd, the C.R.C. type OCE 580 sampling oscilloscope is a single-channel instrument with 0·35nsec rise-time and



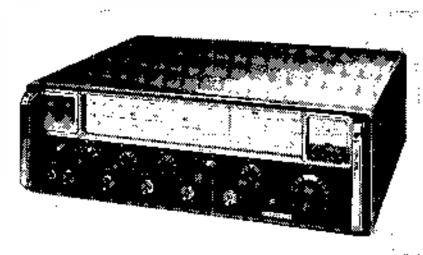
maximum sensitivity of 10mV/cm. A valuable feature is the very wide range of time-base speeds, obtained without sweep expansion. The eighteen calibrated sweep positions cover 0·2nsec/cm to 100μsec/cm. The availability of sweeps as fast as 0·2nsec/cm, without expansion enables trace brilliance to be maintained, and avoids the loss of definition found in other oscilloscopes at high trace speeds due to the need to reduce the number of samples per centimetre to obtain adequate brilliance.

EE 95 763 *for further details*

V.H.F.-U.H.F. RECEIVER

Eddystone Radio Ltd, Eddystone Works, Alvechurch Road, Birmingham, 31

(Illustrated above right)



The Eddystone model "990S" is a fully transistorized single conversion receiver for reception of a.m. and f.m. signals over the range 230Mc/s to 870 Mc/s, thus taking in the whole of Bands IV and V allocated to television transmissions. Two separate r.f. heads, incorporating trough line circuits are used, the range switch simply energizing one or other unit according to the range required.

The use of a high intermediate frequency (36·5Mc/s) ensures good image protection and the alternative bandwidths provided permit reception of signals carrying amplitude, video or frequency modulation. A low impedance output at the intermediate frequency is available for driving ancillary equipment.

The audio frequency response is excellent, as is desirable for monitoring high quality television sound channels. For convenience, a built-in monitor speaker is fitted and outputs are provided for external speaker, telephone headset, and remote lines. In the latter case, the output is restricted to allow direct connexion to lines and has a separate level control.

Basically, the "990S" receiver operates from a 12V d.c. supply and incorporated is an a.c. mains power unit furnishing this supply. The receiver can be driven direct from a suitable d.c. source (dry cells or accumulator), which is of advantage when mobile or field applications are envisaged.

Other noteworthy features include the standard Eddystone precision gear-driven slow motion drive; clear tuning scales; crystal calibrator; panel tuning meter; independent r.f. and i.f. gain controls calibrated directly in decibels; compact size and light weight.

As supplied, the receiver is in table-mounting style, but is easily converted to standard rack mounting.

EE 95 764 *for further details*

ULTRASONIC LEAK DETECTOR

Dawe Instruments Ltd, Western Avenue, Acton, London, W.3

(Illustrated on page 478)

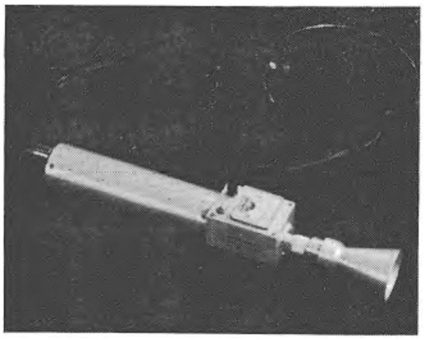
Ultrasonic energy is generated by gas or steam as it leaks from any pressurized container. The amount generated depends on the velocity of the escaping gas, which in turn is dependent on the pressure and dimensions of the orifice. In all cases, however, there is a broad frequency spectrum in the 40kc/s region, and this may be utilized to locate a leak, even in the presence of considerable audible noise from other sources.

This principle is utilized in the new Dawe ultrasonic leak detector (type 1873), a hand-held device no bigger than the average torch. The detector consists of a highly directional ceramic microphone-type transducer coupled to a tuned (40kc/s ±4kc/s) transistor amplifier. Output of the amplifier is indicated visually on a moving-coil meter, or audibly by means of detachable headphones. Apart from the headphones the whole unit is contained in a robust aluminium casing measuring 2in diameter × 13½in long. Weight is 2 lb complete with battery. One battery is sufficient for at least 150h of continuous use, and a battery check is provided.

In operation the detector is simply orientated manually to obtain a maximum meter reading or audible signal. The leak is then in the direction indicated, and may be accurately located by following through.

Sensitivity may be adjusted by a volume control which raises or lowers the output level, thus making it possible to search in close proximity to a high-velocity leak or as much as 45ft from an orifice of 0·01in diameter at a pressure of less than 10 lb/in^2. Where there is sufficient pressure the orifice may be as small as 0·002in diameter.

While the unit is principally designed for leak detection, it is also capable of locating any source of 40kc/s such as a worn bearing, turbulence (hydraulic or pneumatic), friction, corona discharge, etc. Special probes may be required in these cases.



EE 95 765 *for further details*

MICROMINIATURE BATTERY

Mallory Batteries Ltd, Gatwick Road, Crawley, Sussex

A new mercury-oxide battery which is only 0·220in diameter, 0·130in high and weighs 0·3g is now available from Mallory Batteries Ltd.

This battery, type RM-212, is rated at 1·35V and possesses the long life, voltage stability and steady discharge characteristics for which this type of battery is noted.

This cell was developed specifically for a new type of hearing aid which fits completely into the ear duct but it is equally suitable for other uses such as in 'radio pills' and other diagnostic probes.

EE 95 766 *for further details*

BISTABLE RELAY

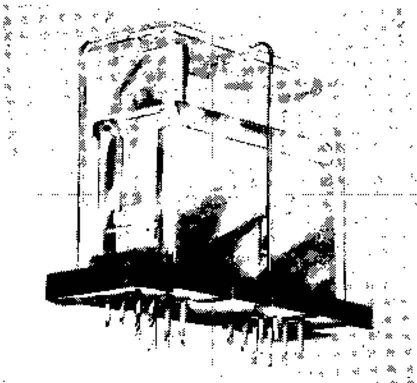
P.A.R. Ltd, Talbot Works, Talbot Street, Nottingham

(Illustrated above right)

This new bistable relay has four changeover contacts, which are mechanically locked in position by means of two dynamically balanced armatures. The latch is interlocked and reset electrically.

A robust design enables the relay to meet severe conditions of shock in any plane without causing its latch to either lock or unlock.

This feature is of particular value to



manufacturers of coin operated machines, burglar alarms, signal time recorders, counting devices and of moving or vibrating machinery.

Other typical uses may be found in industrial controls and memory devices and requirements where a momentary signal must provide continuous indication with the facility for being reset.

The relay is compact, being only 1in × $2\frac{1}{4}$in × 2in high including the transparent cover which is fitted as standard. Contact and coil tags are 0·1in grid for printed circuit applications.

EE 95 767 *for further details*

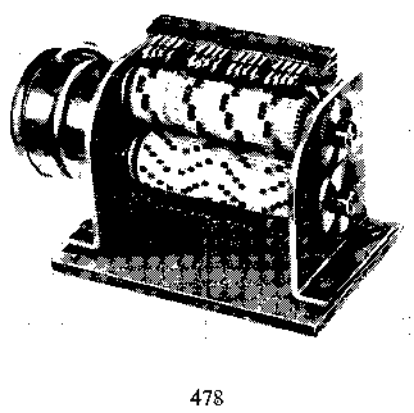
PROGRAMMING SWITCH

Sealectro Ltd, Walton Road, Farlington, Portsmouth, Hampshire

(Illustrated below)

A new dual drum Actan programming switch, providing an extremely long term programming capability, is now available from Sealectro. This new switch features a drive drum with rotational speeds from 5 rev/min to one revolution in 28 days. The secondary drum moves one step with each 360° rotation of the drive drum providing the extremely long term control function. The unit illustrated has 36 programme positions and 16 independent contact stations per drum. Other units are available with up to 120 programme positions and 57 independent contact stations per drum. This new programming switch may also be driven with a remotely controlled solenoid actuator to provide stepping and counting functions. Sliding contact activators provide simple and fast field adjustment of programmes without the need for special tools.

Special 10A contacts, are also avail-



able to meet special control application requirements.

In addition to this new long term programming device, Sealectro offers a complete line of programming switches with features including removable drums, stacked contacts, adjustable contacts, multiple contact positions and fixed programme units.

EE 95 768 *for further details*

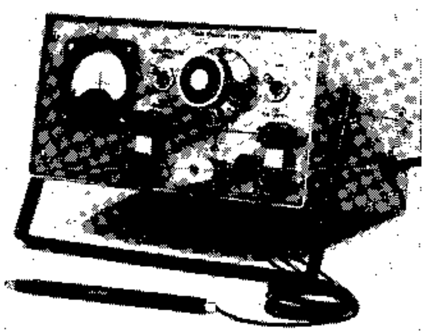
FIELD PLOTTERS

Servomex Controls Ltd, Crowborough, Sussex

(Illustrated below)

Servomex Controls Ltd has produced an improved field plotter for the Teledeltos resistance analogue method, designed for solving field problems in electrostatics, magnetism, mechanics, heat transfer, and hydraulics.

The plotter is available in two models, FP 130, which is powered by a 9V battery and FP 144 which is mains opera-



ted. Both instruments measure the potential in the field by a potentiometer method using the source voltage for the slide wire as well so that the measurement is unaffected by variations in the supply voltage.

This method uses Teledeltos paper which is relatively cheap and has a suitable resistivity and one can draw the analogue and start plotting the field in less than thirty minutes. A special silver paint of an improved type is used to paint in the equipotentials. An electric voltage derived from the apparatus is applied between the electrodes, and the potential in the field mapped by the potentiometer.

The battery operated FP 130 model is useful for students and uses a single turn potentiometer of good quality giving an accuracy of plus or minus 1 per cent but an electrical 'band spread' allows greater discrimination in the range 0 to 10 per cent. A plug and socket are provided to which can be connected any supply up to 100V d.c.

The mains operated FP 144 model has a low voltage of 5 to 10V for ordinary Laplace fields and a high voltage of 50 to 100V for feeding 'constant current' through high value resistors into points on the field for Poissonian fields so that the flow of current is virtually independent of the rest of the pattern. This model uses a good quality ten-turn potentiometer with a direct reading dial which can be read to 0·1 per cent.

The Servomex field plotters are cheap, time saving and proof against short-circuits. They are compact and easily portable, the battery operated version weighs only $1\frac{1}{2}$ lb and the mains operated model 5 lb.

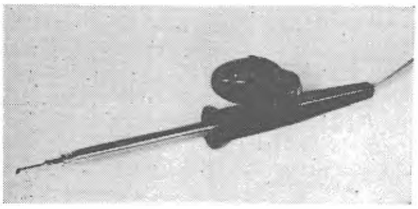
EE 95 769 *for further details*

DE-SOLDERING IRON

Distributed by: Soldering Instrument Components Ltd, Adcola House, Gauden Road, London, S.W.4

(Illustrated below)

A soldering iron designated M207 and whose primary use is the removal of solder from tags, components and printed boards, is announced by Adcola. The instrument is manufactured by



Adcola Products Pty, Ltd of Melbourne, and is available through Soldering Instrument Component Ltd, a subsidiary of Adcola Products Ltd.

The instrument is provided with a hollow stainless-steel-lined tip, which can be used for normal soldering as well as for de-soldering, or alternatively, solid bits can be inserted. The bulb protruding from the handle, which enables operation with the forefinger, draws solder through the tip and a gauze filter situated at the other end of the bit collects the solder which may then be blown out by pressing the bulb. The gauze filter may be removed from the filter chamber for cleaning.

A range of 12 tips is available for use with the M207, six hollow for de-soldering and six solid. The range includes tips with face diameters of $\frac{1}{16}$, 3/32, $\frac{1}{8}$, 5/32, 3/16 and $\frac{1}{4}$in, the hollow $\frac{1}{8}$in tip being included as a standard fitting. The tips are iron plated to prevent seizure in the barrel and to prolong the life of the face.

Heater elements, de-soldering and soldering bits, filter and filter chambers are stocked by Soldering Instrument Components Ltd, who also provide servicing facilities for the U.K. Instruments can be supplied for any voltage range between 6 and 250V.

EE 95 770 *for further details*

MICROWAVE NOISE ANALYSERS

James Scott (Electronic Engineering) Ltd, Carntyre Industrial Estate, Glasgow, E.2

James Scott (Electronic Engineering) Ltd announce a new range of improved high power noise measuring equipments type 123, capable of even more sensitive measurements than previous equipments.

Previously the sensitivity for f.m. and a.m. noise in 150kc/s from the microwave carrier signal, expressed as a signal-to-noise power ratio for total sideband power and measured in a 100kc/s bandwidth was as follows:

Modulation Frequency	1kc/s	10kc/s	100kc/s	150kc/s
A.M.	125dB	134dB	135dB	135dB

The f.m. sensitivity achieved was 0·0025c/s r.m.s. deviation.

The new equipments can achieve the following sensitivity:

Modulation Frequency	1kc/s	10kc/s	100kc/s	150kc/s
A.M.	125dB	143dB	163dB	168dB

The f.m. sensitivity achieved is 0·0008c/s r.m.s. deviation.

These figures have been achieved by use of a new form of construction which allows a higher power sample to be taken from the source under test. In addition to this, the effects of acoustic noise and vibration have been minimized. Complex feedback loops have been utilized for:

(a) Maintaining the bridge cavity on balance. A novel patented system including a thermo syphon servo element is used.

(b) Enabling noise measurements to be made on certain frequency modulated carriers by cancellation techniques.

(c) Eliminating spurious modulation effect from the power supply of the source under test.

(d) Frequency locking the local oscillator.

Improvements in the analysis equipment permits examination of the noise spectrum between 50c/s and 1kc/s from the microwave carrier in a 30c/s bandwidth.

These equipments are already available in X and J Band versions and will shortly be available for C and S Bands.

EE 95 771 *for further details*

HIGH GAIN AMPLIFIER

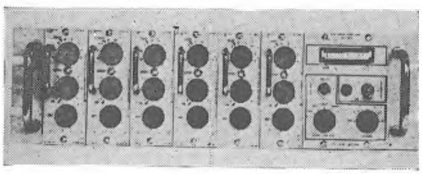
S.E. Laboratories (Engineering) Ltd, North Feltham Trading Estate, Feltham, Middlesex

(Illustrated below)

A low-noise, high-gain amplifier unit SE.436 with a flat response from 5c/s to 10kc/s, is being introduced by S.E. Laboratories (Engineering) Ltd. Its primary use is with amplitude-sensitive mechanical/electrical transducers, e.g., d.c.-energized strain gauges, self-generating velocity pick-ups, etc.

Two integrating networks of different time-constants are additionally incorporated in the variant linear/integrating amplifier SE436/I.

When switched into circuit, either one

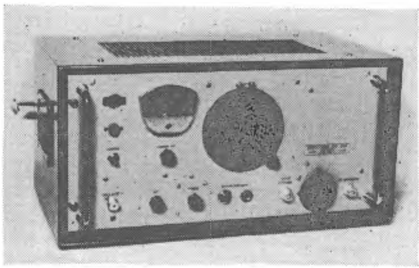


of them enables the output signal of self-generating velocity pick-ups to be integrated for subsequently indicating or recording the vibration amplitude.

These amplifiers' outputs are suitable for driving high frequency recording galvanometers, tape recorders, oscilloscopes, etc. Multiple feedback loops ensure a high degree of gain stability and in-phase rejection.

Separate voltage and current outputs are provided. Due to the voltage output's low source impedance, the amplifier is suitable for all types of voltage-sensitive recording or display equipment.

EE 95 772 *for further details*



U.H.F. WOBBULATOR

Compagnie Générale de Métrologie, B.P.30, Annecy, France

(Illustrated above)

The u.h.f. wobbulator type 241A, used in conjunction with an oscilloscope and a marking generator, constitutes a curve-tracing bench. It may be used to determine the amplitude/frequency response curve of active or passive circuits operating within the frequency range of 470 to 870Mc/s.

Its applications include adjustment of u.h.f. amplifiers, determination of the overall curve of a television receiver (channels IV and V) from aerial to display, measurement of the s.w.r. of aerials, coaxial cables and other circuits, in conjunction with a matched bridge type HCX35A and a l.f. amplifier type HCX52A.

For this purpose the wobbulator delivers, through a Lecher-line oscillator, a frequency-modulated u.h.f. voltage which is continuously recorded on a dial graduated in Mc/s.

The frequency is modulated by means of a vibrating capacitor mounted on a rectangular bronze spring type support. This arrangement has the advantage that it yields a linear displacement while at the same time ensuring an overall robust design.

In order to eliminate parasitic amplitude modulation and preserve a constant output level independent of the frequency displayed, a servo circuit adjusts the voltage by acting on the oscillator supply.

Wobbulator type 241A has an extra u.h.f. output which enables the curve displayed to be marked.

The unit can be supplied in a number of designs to suit requirements as to impedance output and general appearance.

EE 95 773 *for further details*

SHORT NEWS ITEMS

The Sixth International Conference on Microwave and Optical Generation and Amplification (formerly the Conference on Microwave Tubes), sponsored jointly by the Electronics Division of the Institution of Electrical Engineers and the Institution of Electronic and Radio Engineers, will take place at the University of Cambridge from 12 to 16 September, 1966.

The aim of the conference is to stimulate the exchange of ideas on the research that has taken place since the Fifth Conference in Paris, in September 1964, in microwave and optical generation, amplification and other active processes. Contributions on recent outstanding scientific and technical progress in these subjects will be presented and discussed.

The programme will cover microwave and optical generation by solid state, plasma and other active devices, as well as by the classical forms of microwave tubes. In addition it will include the application of various principles and phenomena relevant to the generation and amplification of coherent electromagnetic waves, without strict frequency limitations.

Accommodation will be available in colleges, if required.

Further information about the conference and registration forms will be available from the Joint Conference Secretariat, IEE, Savoy Place, London, W.C.2.

The date of the next IMEKO Symposium on Microwave Measurement at Budapest Hungary has been changed to 18 to 20 October, 1966.

This change has been made so that the Symposium will follow the General Meeting of the International Electrotechnical commission being held in Tel-Aviv, Israel.

The British Standards Institution has published B.S. 2401 entitled, 'Polyester film dielectric capacitors for direct current for use in telecommunications and allied equipment' and dealing with fixed capacitors with a rated voltage not exceeding 6300V and using as dielectric either polyethylene terephthalate or a material with similar properties. The standard is divided into three sections.

Section 1 defines terms, gives a classification into categories according to B.S. 2011 and gives standard values and tolerances for rated capacitance and voltage.

Section 2 establishes uniform requirements for judging the electrical, mechanical and climatic properties of the capacitors and gives the necessary test methods. It requires reference to be made to B.S. 2011. Appendices give examples of specified tests in this section.

Section 3 which is still under consideration will state standard dimensions for the capacitors.

The standard is identical in technical content and layout with IEC Publication 202, except where indicated to the contrary.

Copies of B.S. 2401 may be obtained from the B.S.I. Sales Branch, 2 Park Street, London, W.1. Price 15s. each (postage 1s. will be charged extra to non subscribers).

The British Radio Valve Manufacturers' Association (BVA) and the Electronic Valve and Semiconductor Manufacturers' Association (VASCA) announce that during the first quarter of 1966 exports of valves and tubes were 11 per cent up on the first quarter of 1965 and semiconductors were 16 per cent up.

The biggest increases were in microwave tubes under 1kW and television picture tubes and other electron optical devices including camera tubes.

The full details based on the customs and excise returns are as follows:

Valves and Tubes	Exports £
Thyratrons, etc.	44 497
Photo cells, etc.	129 400
Microwave tubes exc. 1kW	31 617
Microwave tubes: other ...	282 158
Television picture tubes ...	566 152
C.R.Ts.: other	83 204
Other electron optical devices	314 098
Valves over 50W	370 156
Valves: other	948 355
Total	2 769 637

Semiconductors	£
Diodes and rectifiers up to 200mA	148 849
Thyristors	31 056
Other diodes and rectifiers	70 339
Transistors: germanium ...	255 682
Transistors: other	162 290
Integrated circuit assemblies	9 745
Other crystal semiconductor devices	11 832
Total	689 793
Grand total	£3 459 430

The Office for Scientific and Technical Information has awarded a three year grant of £54 140 to the Institution of Electrical Engineers for the support of a project to develop an information service for physicists.

The continuing rapid growth of the output of information in physics, as in other branches of science and technology, makes it clear that the traditional methods used for the dissemination and retrieval of the information will become inadequate in the foreseeable future. The Institution of Electrical Engineers, who produce Physics Abstracts, devised plans for setting up a physics information service making use of computer techniques for handling large quantities of information quickly.

During the course of this project it is intended to design and test the feasibility of the computer-aided production of Physics Abstracts. The techniques and computer programmes developed during this project should be suitable for the computer production of abstract journals in other scientific disciplines. The proposed system would enable a range of additional information services to be provided without the application of further intellectual effort.

The project was devised after a series of discussions between the Institution of Electrical Engineers and the American Institute of Physics and, it is hoped, will form part of a co-operative venture between the two organizations.

The MSF standard time and frequency transmissions on 60kc/s will be radiated for 24 hours a day instead of the present daily one hour period from 1430 to 1530 GMT as from 1 June, 1966.

For an interim period the present one hour schedule of A2 modulation pulses consisting of 5 cycles of a 1000c/s tone for the seconds pulses and 100 cycles of 1000c/s tone for the minute pulses will be retained.

The new extension to the service for the remaining 23 hours of the day will operate with a modulation of interrupted carrier pulses of 100msec duration for the seconds pulses and 500msec duration at the minute, the epoch of the pulses in each case will be at the start of the break in the carrier signal.

The estimated decay time of the pulses is 0·5msec.

A shut down period for maintenance will be observed on the first Sunday of every month between the hours of 10.00 a.m. and 2.00 p.m.

G.E.C. Road Signals Ltd (a subsidiary of the General Electric Company) has been awarded a contract by the Ministry of Transport for the provision, installation and maintenance of vehicle queue detectors for the West London Area Traffic Control experiment.

The contract, worth more than £50 000, involves the installation of 173 double loop detectors with associated logic circuits throughout the six square mile area covered by the scheme. The equipment is entirely solid state and will be housed in sealed containers below the road surface at the critical approaches to, and exits from road junctions controlled by traffic signals. This will be the first large-scale installation of sub-surface detectors in the U.K.

The queue detectors will be an integral part of the computer controlled traffic control scheme. They will assess the speed of vehicles passing over them and will compare it with a predetermined speed level. If vehicles' speeds fall below this level or if vehicles become stationary, a queueing conditions will be indicated and the detector will signal the condition to the central computer. This essential information on the presence of queues will be continuously fed into the traffic control system and will be used in calculating the optimum traffic signal settings to achieve the fastest and most effective traffic flow.

A. T. & E. (Bridgnorth) Ltd—a subsidiary of the Plessey Electronics Group—has developed a sun-powered radio telephone system known as SUNCALL, for installation on the 12 miles of the new motorway between Accra and Tema in Ghana.

The system comprises four pairs of telephone boxes at intervals along the motorway, powered by solar energy convertors and connected to low power transistorized radio transmitters.

The radio equipment is hermetically sealed, weather protected and mounted on or close to an aerial pole adjacent to a standard motorway booth. A similar telephone booth on the opposite carriageway is connected in parallel with the radio controlled booth by a vermin-proof cable under the carriageway.

Call boxes are placed up to two miles apart and one section of the system can cover some 10 to 18 miles of motorway according to the geography of the countryside. Further outstations can be added at the opposite side of the base station on the same pair of frequencies. This enables an economic use to be made of frequencies over some 20 to 36 miles of motorway.

The Ministry of Aviation has awarded a contract to The Marconi Co. Ltd. to develop and supply 31 television cameras of new design to be used to transmit vital air traffic control data in a variety of air traffic control applications. New techniques are being employed, including the use of very stable thin film circuits, to ensure that the highest possible picture quality is maintained.

Three of these new cameras will be used at the new R.A.F. Air Traffic Control Centre at West Drayton, where they will provide controllers with airfield availability data and meteorological information. A very large number of radar displays and data handling equipment are already on order from The Marconi Company for this new centre.

Other cameras will be installed at London (Heathrow) Airport, at the London Air Traffic Control Centre, at the Telecommunications Engineering Establishment at Gatwick, and at the Civil Aviation Establishment at Bletchley. They will be used both operationally, and for the development of new systems and the training of operators.

Telefunken AG, Berlin/Ulm, has sold a minority holding of 23 per cent of the shares of Elektronische und Luftfahrtgerate GmbH (ELEKLUFT), Bonn, to the Hughes Aircraft Company, U.S.A. With 51 per cent of the shares Telefunken continues to hold a majority of the capital of ELEKLUFT, in which the General Electric Company, U.S.A., already has a 26 per cent holding.

This transaction, which was undertaken in agreement with General Electric, is aimed at expanding the business basis of ELEKLUFT. The company services electronic equipment and supplies the necessary spares and replacements for this purpose.

The Compagnie Française Thomson-Houston recently carried out at their research laboratories at Bagneux an experimental transmission of live television over a laser beam.

The laser was a solid state device producing an output of about one milliwatt, and the receiver consisted of a photoconducting diode placed at the centre of an optical system.

An International Conference on the Automatic Operation and Control of Broadcasting Equipment is to be held at the Institution of Electrical Engineers, London, on 9 to 11 November this year.

The conference, sponsored by the Institute of Electrical and Electronic Engineers, the Institution of Electrical Engineers, the Institution of Electronic and Radio Engineers and the Television Society will embrace all aspects of the automatic operation and control of broadcasting equipment including the following:

Programme Distribution Systems;
Picture and Sound Generation Equipment (including Automatic Lighting Control);
Switching and Distribution Equipment;
Communication Equipment;
Primary and Relay Transmitting Equipment;
Monitoring Equipment;
(Reference to Domestic Receivers will not be included).

Standard Telephones and Cables Ltd is to install a computer-based message switching system which will make the Shannon Aeradio Telecommunications Centre at Ballygirreen, six miles north of Shannon Airport, one of the most advanced in the international avation telegraph network.

The contract is for a dual 6300 ADX system which will operate in accordance with ICAO recommendations.

Shannon Aeradio is a focal point linking aircraft in flight in the North Atlantic Major World Air Route Area (MWARA) with the Aeronautical Fixed Telecommunications Network (AFTN) and thus provides a vital service to international flying on the North Atlantic routes. With the constant increase in North Atlantic air traffic, expected to reach 60 000 crossings in 1966, more sophisticated capability at Shannon Aeradio has become necessary.

In addition to dealing with the air/ground inputs, the 6300 ADX will also transmit messages over the Scot-Ice cable (a simplex circuit linking Shannon, Gander and Reykjavik) and over separate party lines to Cork and Dublin Airports, apart from international circuits to London and Prestwick.

Associated with these party lines will be three Multi Output Selection Units (MOSU's), two situated at Dublin Airport and one at Cork Airport. These MOSU's are switches, remotely operated by the message from Shannon. Each unit will automatically connect up to eight teleprinter terminals in the airport areas to the line from the ADX at Shannon in accordance with a code in the heading of the message itself.

The Television and Radio Show, sponsored by the Radio and Television Retailers' Association (RTRA) is to be held at Earls Court, London, on 22 to 26 August this year.

This will be a trade exhibition only and will for the first time, be international. Included in the exhibition will be audio and video equipment, and colour television receivers.

It is being organized by Industrial and Trade Fairs Ltd, Commonwealth House, 1-19 New Oxford Street, London, W.C.1.

The Elettra Sound Systems Department of the Marconi International Marine Co. Ltd, is to supply 122 units of an h.f. radio inter-train communication system and a wired public address and intercommunication installation for the entire rolling stock of London Transport's new Victoria Line.

Each train operator's cab is to be fitted with a telephone handset and selector panel, three positions on which will enable the trainman to select 'public address', 'intercommunication' or 'inter-train' communication' as required.

The radiotelephone inter-train communication system has been designed for use primarily when assistance is required from the train following. A low power v.h.f. transmitter and receiver will be housed in each cab with an aerial at the front of each motor car. Selection of 'inter-train communication' on the cab selector panel and the depression of an associated push-button will, after alerting the called train operator, enable a normal two-way radiotelephone conversation to take place.

When operating, the Victoria Line trains will normally comprise two four-car units with a cab at each end of the unit. To enable the train operator in the forward cab to communicate with other personnel in any of the other cabs in the train a facility providing for inter-cab communication will be included in the system. Selection of 'intercommunication' on the selector panel and the depression of a button will produce a calling tone in a small loudspeaker in the other cab units throughout the train. When any one of these cab handsets is lifted the tone call will immediately cease, permitting two-way telephone communication.

The third facility, which will normally be used when the train is stopped, will enable the train operator to convey information to the passengers simply by selecting 'public address' on the selector panel. His information and/or instructions will be heard over the six loudspeakers which are to be fitted in each car.

A Committee of European Associations of Manufacturers of Passive Electronic Components (CEPEC) has been set up as a result of a meeting in Paris last month.

Members of the Committee are representatives nominated by the national associations of Belgium, France, Germany, Italy, the Netherlands and the United Kingdom; these associations being—

Belgium: Fédération des Entreprises de l'Industrie des Fabrications Metalliques (FABRIMETAL).

France: Syndicat des Industries de Pièces détachés et Accessoires Radioélectriques et Électroniques (SIPARE) and Syndicat des Constructeurs Francais de Condensateurs Électriques Fixes (SCFCEF).

Germany: Zentralverband d. Elektrotechnischen Indusrie E.V.—Fachverband Schwachstromtechnische Bauelemente 23 (ZVEI).

Italy: Associazione Nazionale Industrie Elettrotechniche ed Elettroniche Gruppo 14 e. 15 (ANIE) C.

Netherlands: Groep Fabrieken voor Passieve Elektronische Bouwelementen in Nederland (FAPEL).

United Kingdom: Radio and Electronic Component Manufacturers Federation (RECMF).

The Committee was formed with the object of carefully observing any technical difficulties liable to impede the development of the interchange of products in Europe and to co-operate in overcoming such impediments.

CEPEC has dealt so far with problems of technical standards (norms and specifications). It is not intended to replace national specifications but to arrive at a harmonization through recommendations to appropriate standardization bodies such as I.E.C., N.A.T.O., etc. A technical committee and working groups have been formed and this approach has proved to be both efficient and successful.

At present, at the request of the U.K., CEPEC is considering the 'Burghard Report' (see page 425, this issue) which contains British proposals for common standards for electronic components, which could well form a basis for a European system.

The M.E.L. Equipment Co. Ltd has now developed an equipment consisting of a transmitter-receiver, modulator and aerial switches for the Phantom supersonic military aircraft currently on order by Britain from the United States.

The M.E.L. equipment is part of a 'sideways looking' radar which has been specially developed to scan the terrain on each side of the track of an aircraft travelling at subsonic and supersonic speeds. The output can be used for providing the aircrew with a continuous photographic map and for providing a video tape record. The output is also suitable for transmission so that the radar picture in the aircraft can be viewed simultaneously at a ground station.

M.E.L. is collaborating closely with EMI Electronics Ltd which is embodying the equipment in a special reconnaissance pack.

The Science Research Council has made a grant of £50 000 over one year to Professor G. Porter, F.R.S., Firth Professor of Chemistry at Sheffield University to provide lasers and part of other equipment needed to enable laser technology to be applied to the flash photolysis technique for the study of chemical states which have a lifetime of less than one microsecond.

The range of the flash photolysis technique is limited by the ability to produce a short flash of adequate energy. To obtain the shorter, high energy flash required, Professor Porter will use a Q-switched ruby laser which gives pulses of several joles energy of 20nsec duration; frequency doubling will be employed to give sufficient output at a wavelength of 3472Å. For monitoring purposes d.c. continuous lasers will be used.

Professor Porter's research group will first study the triplet states of carbonyl compounds and of chlorophyll. They will also investigate the possibility of shortening the laser pulse still further by developing improved chemical Q-switches.

TRW Inc of California and the Mitsubishi Electric Co. of Japan have formed an ocean-spanning team for the design, construction, installation, maintenance and operation of communication satellite ground stations. This international team, joined by Hawker-Siddeley Dynamics, Australia, a division of Hawker de Havilland of Australia, is now conducting design studies preliminary to submitting a proposal to the Australian Government for a ground station to be installed near Sydney.

The Australian ground station will be acquired and owned by the Oversea Telecommunications Commission of the General Post Office. It will consist of an 85ft or larger parabolic antenna; electronic equipment for receiving, transmitting and landline distribution of television and telephone communications; and maintenance facilities. Completion of the station is estimated to occur about the end of 1967, in time for operations of the Apollo communications satellite network now being developed by the Communications Satellite Corporation (COMSAT) for the U.S. National Aeronautics and Space Administration. COMSAT's international commercial communications system is expected to become operational by mid-1968. The spacecraft for this system are to be designed and built by TRW for the Comsat Corporation.

An International Medical Engineering and Automation (MEDEA) Exhibition is to be held at Earls Court, London, 13 to 17 March 1967.

This Exhibition, sponsored by the Electronic Engineering Association and the Scientific Instrument Manufacturers' Association, and organized by Industrial Exhibitions Ltd of 9 Argyll Street, London, W.1 will cover all aspects of medical electronics and instrumentation, biological engineering, medical automation, hospital physics, and electromedical and X-ray equipment.

The Second European Symposium on Medical Electronics will be held at Earls Court during the Exhibition.

Correction

In the article 'A Computer Controlled C.R.T. Display System' by W. Bial in the June 1966 issue there is an error in Fig. 8 on page 358. The feeder lines from 'Odd comb' and 'Even comb' should have series capacitors inserted into them.

NOUVELLES Réalisations

Traduction des pages 474 à 479

Une description basée sur des renseignements fournis par les fabricants de nouveaux organes, accessoires et instruments d'essai

CONDENSATEURS ELECTROLYTIQUES

The Telegraph Condenser Co. Ltd, Wales Farm Road, North Acton, London, W.3

(*Illustration à la page 474*)

La nouvelle série de condensateurs électrolytiques à feuilles d'aluminium récemment mise au point par la société The Telegraph Condenser Co. Ltd, qui fait partie du Plessey Components Group, assurera d'excellents services dans la gamme de températures de −55° C à +125° C.

Ces condensateurs électrolytiques à gamme de température étendue sont fabriqués dans le Royaume-Uni sous licence exclusive du gouvernement canadien, dans une gamme de capacités et de dimensions conforme à la spécification militaire américaine Mil-C-39018. Une spécification militaire britannique analogue est en cours de préparation et ces condensateurs seront soumis au contrôle d'homologation.

Les condensateurs à gamme de température étendue (Extended Temperature Range) pèsent environ un tiers du poids des condensateurs équivalents à feuille de tantale et ont une plus grande capacité par volume d'unité. Leur performance électrique s'approche en effet de celle des condensateurs au tantale et ils répondent à une spécification semblable. La chute de capacité à −55° C est faible et les caractéristiques électriques demeurent très stables durant les essais de vie dépassant 2000 heures à 125° C.

Les caractéristiques de repos sont nettement supérieures et la résistance en série équivalente est bien inférieure à celle des condensateurs électrolytiques ordinaires à feuille d'aluminium dans une gamme de température étendue. Ces avantages en font des condensateurs beaucoup plus appropriés pour le fonctionnement dans des circuits à impulsion où la forme des impulsions doit être maintenue.

Une bonne partie de l'amélioration de la performance peut être attribuée à l'emploi d'un nouveau type d'électrolyte contenant du formamide di-méthyle comme solvant, ce dernier nécessitant à son tour un nouveau type de construction et de scellement.

Les boîtiers tubulaires sont conçus de manière à limiter le mouvement de l'élément interne afin de répondre aux conditions de vibration. Le boîtier se distingue en outre par une "zone affaiblie" prévue pour que le boîtier puisse éclater sans explosion au cas où se produirait une pression de gaz intérieure élevée, telle que celle qui pourrait être causée par l'inversion accidentelle d'un élément polaire.

Les essais de durée de vie des compensateurs à gamme de température élevée aux tensions nominales et à des températures maxima ont montré une stabilité remarquable des paramètres; les changements de pourcentage de capacité, du facteur de puissance et du courant de fuite après une durée de 2000 heures ont été considérablement inférieurs aux limites prescrites dans la spécification Mil-C-39018.

EE 95 751 *pour plus amples renseignements*

OSCILLOGRAPHE À RAYONS ULTRAVIOLETS

Southern Instruments Ltd, Camberley, Surrey

(*Illustration à la page 474*)

Le dernier-né de la gamme des oscillographes à rayons ultraviolets produits par la société Southern Instruments Ltd. est le modèle M1310. Cet instrument est de construction très robuste et comporte un grand nombre de dispositifs incorporés.

L'instrument standard fourni pour montage sur bâti ou sur bande d'essai, peut recevoir de 1 à 50 canaux et enregistre sur n'importe quelle largeur de papier jusqu'à 30,48 cm. Douze vitesses d'avance du papier, mises en action par bouton-poussoir, donnent une gamme de 4 à 2500 mm/sec. Cette gamme peut être étendue pour fournir 24 vitesses dans la gamme de 0,004 à 2500 mm/sec. Ou, en employant l'élément à très grande vitesse, quatre vitesses complémentaires allant jusqu'à 25000 mm/sec.

L'optique de l'appareil réduit la défocalisation sur les bords du papier à une proportion négligeable même lorsqu'on emploi le papier d'enregistrement de 30,48 cm. Grâce à l'emploi des techniques de circuit modernes, un courant constant est toujours fourni à la lampe, maintenant ainsi son fonctionnement dans des conditions optima.

D'autres particularités intéressantes de l'instrument comprennent: un marqueur d'événements, une commande de répétition, un enrouleur de papier, des lignes de réseau optique, des lignes de minutage à contrôle par cristal sur toute la largeur, l'identification de la trace, le numérotage de la trace et du coup et la télécommande.

EE 95 752 *pour plus amples renseignements*

MULTIMÈTRE

Distributeurs: Kelvin Electronics Co. (Smiths Industries Ltd), Kelvin House, Wembley Park Drive, Wembley, Middlesex

(*Illustration à la page 474*)

Le "DigiKelvo" est un multimètre donnant la polarité et un affichage numérique à trois chiffres de la valeur mesurée sur des tubes numérotés. Cet instrument comporte quatre canaux d'entrée séparés couvrant les tensions alternatives et continues, le courant et la résistance. Deux canaux sont utilisés pour la mesure de tension de 1 mV à 1 kV en six gammes. Le troisième canal est utilisé pour mesurer le courant de 100 nA à 999 mA en sept gammes et, par l'emploi de la dérivation jusqu'à 99,9 A en deux gammes. Le quatrième canal est utilisé pour mesurer la résistance de 1 Ω à 999 MΩ en sept gammes. Le changement de gamme automatique s'effectue en déplaçant un point décimal à travers l'affichage. Il y a, en outre, un commutateur de gammes à deux directions qui permet d'obtenir six ordres de grandeur.

Le système se compose d'un amplificateur de tension continue à faible dérive et à haute performance dont la sortie est contrebalancée par un circuit en escalier, et un compteur qui affiche le nombre d'incréments requis pour atteindre l'équilibre nul. Les principes utilisés dans le circuit en escalier et les circuits de comptage sont des principes classiques mais la performance élevée de l'amplificateur à tension continue est exceptionelle à plus d'un égard.

L'amplificateur à tension continue a une dérive à court terme par rapport à une entrée de 0,5 μV/°C, et une dérive estimée à long terme de 2,5 μV par an. Le gain de circuit ouvert de l'amplificateur est de 2×10^8 qui, dans la configuration de gain la plus élevée, est réduit par réaction à 14 000. La largeur de bande de l'amplificateur est de 600 kH (bien que d'autres caractéristiques de l'instrument peuvent réduire la largeur de bande globale à 100 kHz). Cette largeur de bande assure à l'instrument une période de mesure maxima de 0,3 sec. La linéarité de l'amplificateur est supérieure à 0,1 %, donnant à l'instrument une précision de ±1 chiffre dans le dernier nombre. Bon nombre de ces caractéristiques peuvent être attribuées à l'emploi d'un transistor à effet de champ comme relais modulateur à l'entrée de l'amplificateur.

Le "DigiKelvo" se distingue également par le fait que le bobinage interne a été complètement supprimé grâce à l'emploi de plaquettes de circuit imprimé et de connecteurs appropriés.

EE 95 753 *pour plus amples renseignements*

FILTRE ANTI-PARASITES HF

Erie Resistor Ltd, South Denes, Great Yarmouth, Norfolk

La société Erie Resistor Ltd vient d'annoncer que le ministère de l'aviation a homologué deux appareils de sa gamme étendue de filtres anti-parasites HF, série 1200, pour matériels d'aviation militaire. Ces deux appareils portent les références 1200-025 et 1215-000 respectivement.

Les filtres anti-parasites à large bande de la série 1200 comprennent un condensateur en céramique de 1,5 μF (conçu suivant une méthode de fabrication brevetée par la société Erie) et des chenilles en ferrite qui produisent l'effet inductif en série (pertes résistives aux hautes fréquences). Le filtre se trouve dans un robuste logement métallique scellé comportant une douille filetée. Le montage mécanique est à transmission d'entrée avec un choix de bornes. Compte tenu de sa performance et de son prix, le filtre anti-parasites à large bande HF Erie peut êre considéré, à juste titre, comme le plus petit élément du genre offert sur les marchés mondiaux.

Les filtres de la série 1200 sont à large bande, dans ce sens que leur fréquence de fontionnement, afin de permettre une atténuation utile, est dans la gamme d'environ 5 kHz à 1 GHz et même d'avantage. L'atténuation maxima est dans la direction d'entrée en série inductive puisqu'il s'agit essentiellement de filtres à section L. L'atténuation minima, ainsi que toute inclinaison de résonance pouvant se produire, est à 70 dB au-dessus de la plus grande partie du spectre de haute fréquence.

La tension de régime continu est de 50 V c.c., avec environ 50% de diminution de taux pour le fonctionnement à 125°C: les conditions de fonctionnement de la spécification britannique 2G.100 sont donc remplies.

EE 95 754 *pour plus amples renseignements*

CONTRÔLEUR DE DÉRIVE DE TENSION

J. & P. Engineering (Reading) Ltd, Portman House, Cardiff Road, Reading, Berkshire

(*Illustration à la page 475*)

Ce contrôleur de dérive se compose d'une référence de tension, d'un réseau de division et d'un amplificateur de déviation monté dans une enceinte à température constante. Il a été principalement conçu pour mesurer la dérive d'alimentation dans la gamme de 0,1 V à 5 kV et de 100 μA à 5 A. La stabilité et la sensibilité sont d'une importance primordiale pour ces utilisations.

Le contrôleur de dérive comporte trois étages de contrôle de température donnant un coefficient général de température supérieur à $0{,}2 \times 10^{-6}$/°C et une stabilité supérieure à une partie par million pour un changement de tension secteur de 10%. La tension de référence est une diode Zener à compensation de température actionnée par un générateur de courant très stable. Les principales résistances de division sont bobinées de manière "lâche" pour leur assurer une stabilité à long terme, leur coefficient de température est très bas et elles sont montées dans une enceinte à température constante. On fait circuler l'air dans cette enceinte au moyen d'un ventilateur puissant de façon à assurer une faible montée de température des résistances de mesure lorsqu'une entrée est reliée.

Le contrôleur de dérive est d'une stabilité et d'une sensibilité suffisantes pour enregistrer des dérives de quelques parties par million pendant plusieurs jours et il est pratiquement indépendant des variations de courant secteur et de température. Il est entièrement transistorisé et monté dans un coffret de 48,2 cm ayant un panneau frontal de 26,6 cm de haut. Il a été conçu par la société J. & P. Engineering (Reading) Ltd, pour contrôleur les alimations en tension et en courant aux établissements de recherche nucléaire à Harwell.

EE 95 755 *pour plus amples renseignements*

AMPÈREMÈTRE MICROVOLTS À COURANT CONTINU

Claude Lyons Ltd, Valley Works, Hoddesdon, Hertfordshire

(*Illustration à la page 475*)

Le MV-07C est un voltmètre-ampèremètre électronique des plus sensibles pouvant mesurer des tensions continues minima de 1 μV et 1 nA. Un total de 34 gammes de tension et de courant, en une séquence de 10-25-100, portent l'étendue de mesure à 1 kV et 1 mA sur la totalité de l'échelle. La précision est de 1 % sur la plupart des gammes de tensions, jusque et y compris la gamme de 1 kV.

Cet instrument se caractérise par son amplificateur interrupteur à faible bruit, la conception judicieuse du circuit d'entrée pour éviter des connexions métalliques dissemblables, des commandes d'étalonnage individuelles pour toutes les 17 gammes de tension et son galvanomètre à échelle de 15,24 cm à zéro central et à étalonnage individuel avec suspension de bande tendue.

Le MV-07C emploie un amplificateur sensible à faible bruit comprenant un relais modulateur à faible bruit et à dérive réduite suivi d'un premier étage à cascode à faible bruit. Pour augmenter le rapport signal/bruit, cet amplificateur a une largeur de bande étroite, obtenue en ajoutant un double circuit en T dans un des cadres à réaction à courant alternatif. La forte réaction globale de courant continu stabilise l'amplificateur à courants porteurs contre les variations de tension secteur et les changements de caractéristiques de composantes.

Le démodulateur synchrone transistorisé produit une sortie en continu proportionnelle à la tension ou au courant d'entrée et en phase avec l'entrée, ce qui permet à l'instrument de mesure du zéro central de lire tant les entrées positives que les entrées négatives sans commutateur de polarité. La sortie est également transmise à un connecteur pour fournir un signal en phase amplifié pour l'usage extérieur.

Une excellente stabilité du gain est assurée par un circuit réglable de réaction en tension continue qui passe à travers le commutateur de gamme et qui comprend aussi des commandes d'étalonnage individuelles pour toutes les 17 gammes de tension et pour les 6 gammes de courant les plus sensibles.

Cet instrument est fabriqué en Angleterre par Claude Lyons Ltd, sous licence de fabrication exclusive de la société Millivac Instruments Inc., U.S.A.

EE 95 756 *pour plus amples renseignements*

VOLTMÈTRE ÉLECTRONIQUE

Advance Electronics Ltd, Roebuck Road, Hainault, Ilford, Essex

(*Illustration à la page 476*)

Le voltmètre électronique VM76 est le plus récent de la gamme Advance de cinq voltmètres et il permet d'inclure les fréquences en alternatif dépassant 1000 MHz et la mesure de tension continue allant jusqu'à 1000 V avec une précision de ±2 % jusqu'à 100 V.

L'instrument se caractérise par le fait qu'il permet de choisir par boutonpoussoir les différentes fonctions voulues. Pour le fonctionnement en continu, cette méthode peut être utilisée pour choisir des lectures à zéro central et une polarité positive ou négative. Des mesures de résistance entre 0,02 Ω et 500 MΩ d'une précision de ±5 % au milieu de l'échelle peuvent être effectuées de manière analogue.

Une bonne accessibilité aux composants est assurée par le montage de la plupart des composants sur une seule plaquette à circuit imprimé. Chaque instrument est muni de sondes et de câbles de connexion à fiche et douille simple. Des accessoires facultatifs sont prévus pour toutes les applications du voltmètre.

Le voltmètre VM76 fonctionne dans

la gamme de 0°C à +40°C et il est mû par courant secteur. De même que d'autres instruments Advance qui viennent d'être introduits, le VM76 est logé dans un robuste coffret modulaire dont la largeur est du 2/3 d'un bâti standard.

Les gammes de tension continue vont de 300 mV à 1 kV de déviation sur la totalité de l'échelle avec une impédance d'entrée supérieure à 100 MΩ. Les gammes de tension alternative vont de 30 mV à 300 V efficaces. L'impédance d'entrée est de l'ordre de 7,5 MΩ à 1 kHz et de 150 kΩ à 100 MHz.

EE 95 757 *pour plus amples renseignements*

AMPLIFICATEURS DE TENSION CONTINUE

Honeywell Controls Ltd, Honeywell House, Great West Road, Brentford, Middlesex

(Illustration à la page 476)

La société Honeywell va prochainement mettre sur le marché pour la première fois en Grande-Bretagne une série d'amplificateurs de tension continue. Cette série, qui porte le nom de Accudata 104, est livrable en modèles standard donnant un gain de 10 à 250 et en deux autres modèles donnant des gains de 0,1 à 2,5 et de 0,4 à 10.

Ces amplificateurs ont d'excellentes spécifications de bruit et de dérive et ils constituent des amplificateurs universels idéaux pour les extensomètres excités au courant continu, les thermocouples ainsi que d'autres sources de signaux manquant de puissance suffisante pour entraîner directement un enregistreur. Toutes les versions sont conçues pour l'utilisation avec des enregistreurs galvanométriques et ont une puissance de sortie suffisante pour entraîner même les galvanomètres amortis par fluide de fréquence élevée.

L'amplificateur Accudata 104 à été étudié pour empêcher les phénomènes transitoires d'endommager le galvanomètre lorsque des gammes de gain sont commutées, lorsque l'entrée est à circuit ouvert ou lorsque l'appareil est mis en marche ou arrêté. Une commande de panneau frontal détermine la polarité de connexion au galvanomètre et elle permet également de mettre la sortie en court-circuit pour le réglage à zéro du galvanomètre. Des adaptateurs arrière variables sont prévus.

EE 95 758 *pour plus amples renseignements*

CONDENSATEURS VARIABLES

Wingrove & Rogers Ltd, Domville Road, Liverpool, 13

(Illustration à la page 476)

Le condensateur d'appoint C.903 est un trimmer biélectrique muni d'un dispositif de verrouillage. La base de céramique mesure 0,8 pouce × 0,6 pouce et elle comporte deux douilles à prise pour le montage du trimmer. Ce modèle est semblable mais légèrement plus petit que le trimmer C.803.

La capacité minima est de 3,5 pF et elle peut s'élever jusqu'à 34 pF. Le facteur de puissance est inférieur à 0,001 à 1 MHz et la résistance à l'isolement est supérieure à 1000 MΩ à 500 V c.c.

EE 95 759 *pour plus amples renseignements*

COMMANDES DE MÉMOIRES À DISQUE

Distributeurs: Scientific Furnishings Ltd. Poynton, Cheshire

(Illustration à la page 476)

Les commandes de mémoires à disque Control Data série 9430 permettent l'enregistrement à la fois sur bande magnétique et sur disque magnétique et assurent un accès aléatoire rapide et un emmagasinage illimité. Ces avantages sont rendus possibles grâce aux boîtiers à disque amovibles qui peuvent être utilisés pour l'enregistrement, emmagasinés indéfiniment (comme les rouleaux de bande magnétique) et rapidement réinsérés en ligne. Tous les boîtiers à disques sont entièrement interchangeables avec des commandes de mémoires à disques du même modèle; des disques enregistrés sur un élément peuvent être lus sur d'autres appareils. Les commandes 9430 et 9434 sont également compatibles avec les commandes de mémoires à disques des modèles IBM 1311 et 2311 respectivement.

Les commandes de la série 9430 ont toutes des capacités de recherche directe: la dernière adresse utilisée est retenue par la commande de mémoires à disque et elle est fournie à la calculatrice pour déterminer la distance et la direction du mouvement de tête suivant. On obtient ainsi des temps d'accès considérablement plus rapides que par la méthode dite du retour au point de départ. Tous les modèles permettent d'effectuer des opérations de recherche simultanée par plus d'un élément.

Quatre modèles sont prévus pour les caractères à 6, 7 et 8 chiffres binaires avec des capacités totales (6 disques par boîtier) de 2 000 000 à 8 192 000 caractères. Le temps d'accès maximum sur tous les modèles est de 145 msec, le temps moyen étant de 70 msec.

EE 95 760 *pour plus amples renseignements*

PONT COMPARATEUR DE WHEATSTONE

British Physical Laboratories, Radlett, Hertfordshire

(Illustration à la page 476)

Le pont comparateur de Wheatstone à faible charge et à tension continue, modèle RM 197, mis au point par B.P.L. est le fruit de l'expérience de cette société en matière de pont de Wheatstone à faible charge, en particulier le pont RM 196. Le modèle RM 197 est un instrument autonome à double utilisation pour la mesure de résistance. Il se compose essentiellement des composants du pont de Wheatstone, comprenant des bras de rapport et un bras variable, une alimentation de pont et un détecteur de pont.

Le bras variable comprend quatre décades à plots, sa valeur étant affichée sur des indicateurs à lecture numérique. Les bras de rapport fournissent la décimale indiquant l'ordre de grandeur des quatre chiffres principaux.

L'alimentation du pont est à courant constant plutôt qu'à tension constante. Cette arrangement donne une tension à erreur sensiblement constante pour un certain pourcentage de déséquilibre du pont, souhaitable lorsque ce dernier est utilisé dans le mode comparateur.

Le détecteur de pont est un amplificateur stabilisé par relais modulateur à gain élevé et à faible bruit dont la sortie est redressée de manière synchrone pour obtenir des indications de déséquilibre négatif et positif.

Lorsque le comparateur est utilisé comme pont de Wheatstone, la résistance inconnue est reliée aux bornes Rx, et la valeur variable est réglée en fonction d'un chiffre arbitraire quelconque (pas 0000); le pont est ensuite équilibré en utilisant d'abord le commutateur de gamme, puis le chiffre le plus significatif. Une sensibilité maxima est obtenue par l'emploi de la position de 0,5% du commutateur de déviation sur la totalité de l'échelle. La valeur de résistance inconnue est lue directement, sur les indicateurs.

Lorsque l'instrument est utilisé comme comparateur, la valeur de résistance est fixée sur les indicateurs à l'aide du commutateur de gamme et des commutateurs de décades. Le commutateur de déviation sur la totalité de l'échelle est réglé à la gamme de déviation sur la totalité de l'échelle, par exemple lorsqu'on règle des résistances de 10% il faut utiliser la gamme de déviation sur la totalité de l'échelle de 20%. Les résistances sont ensuite appliquées aux bornes Rx et l'instrument de mesure indiquera directement le pourcentage de déviation de la résistance à partir de la valeur fixée.

EE 95 761 *pour plus amples renseignements*

AMPLIFICATEUR-RELAIS MODULATEUR

Kent Precision Electronics Ltd, Vale Road, Tonbridge, Kent

(Illustration à la page 477)

L'appareil K-PAC/4 a été réalisé pour l'utilisation précise d'instruments standard à entrées dans la gamme de 0 à 100 mV, à partir de transducteurs à faible sortie, de thermocouples etc. à entrées dans la région de 0 à 10 mV. Il se compose d'un amplificateur à relais modulateur à faible dérive conçu par KPE et réalisé avec un bloc d'alimentation autonome suivant la construction standard K-PAC. Un relais à transistors module le signal d'entrée en millivolts qui est ensuite injecté à un amplificateur de courant alternatif à gain stable dont la sortie est démodulée pour fournir une sortie de tension continue. La réaction est alors mise en oeuvre de la sortie à l'entrée pour stabiliser le gain à la valeur voulue pour convertir le signal d'entrée en une sortie de tension continue standard de 0 à 100 mV pour la mise en oeuvre d'autres instruments. Le gain est réglé à un maximum de 20 et l'appareil a une performance de dérive inférieure à 3 μV/° C. Pour les instruments spéciaux, le K-PAC/4A à relais modulateur à transistors à effet de champ est prévu pour l'emploi avec des entrées dans la région de 0 à 1 mV avec une performance de dérive de moins de 0,5 μV/° C.

L'instrument est actionné à partir d'une alimentation secteur auxiliaire de courant alternatif et il comporte un transformateur interne de sorte que le circuit est dans tous les cas, isolé du secteur. Il peut être prévu avec tension maxima ou minima aux circuits ouverts des bornes d'entrée de sorte que l'instrumentation connexe peut être à l'épreuve des pannes.

EE 95 762 *pour plus amples renseignements*

OSCILLOSCOPE D'ÉCHANTILLONNAGE

Distributeurs: Claude Lyons Ltd, Valley Works, Hoddesdon, Hertfordshire

(Illustration à la page 477)

L'oscilloscope cathodique d'échantillonnage type OCE 580, distribué dans le Royaume-Uni par la division d'instruments de la société Claude Lyons Ltd, est un instrument univoie à temps de montée de 0,35 nsec et d'une sensibilité maxima de 10 mV/cm. Il se caractérise en particulier par sa gamme très étendue de vitesses de base de temps, obtenues sans extension de balayage. Ses dix-huit positions de balayage étalonné vont de 0,2 nsec/cm à 100 μsec/cm. La possibilité d'obtenir des balayages d'une rapidité de 0,2 nsec/cm, sans extension, permet de maintenir la brillance de trace et obvie à la perte de définition qu'on observe dans d'autres oscilloscopes aux vitesses de trace élevées dues à la nécessité de réduire le nombre d'échantillons par centimètre pour obtenir une brillance adéquate.

EE 95 763 *pour plus amples renseignements*

RÉCEPTEUR V.H.F.-U.H.F.

Eddystone Radio Ltd, Eddystone Works, Alvechurch Road, Birmingham, 31

(Illustration à la page 477)

Le récepteur Eddystone, modèle "990S," est un appareil à simple conversion entièrement transistorisé pour la réception de signaux modulés en amplitude et en fréquence dans la gamme de 230 à 870 MHz, englobant ainsi l'ensemble des bandes IV et V attribuées aux émissions de télévision. Deux têtes HF, comportant des circuits en ligne sont utilisées, le commutateur de gamme amorçant simplement l'un ou l'autre des éléments suivant la gamme voulue.

L'emploi d'une fréquence intermédiaire élevée (36,5 MHz) assure une bonne protection de l'image et les différentes largeurs de bande permettent la réception de signaux de modulation d'amplitude, de vidéo ou de fréquence. Une faible sortie d'impédance à la fréquence intermédiare est prévue pour entraîner des appareils auxiliaires.

La réponse de fréquence acoustique est excellente, comme il se doit pour le contrôle de canaux sonores de télévision de haute qualité. Pour la commodité de l'emploi, un haut-parleur de contrôle incorporé et des sorties sont prévus pour le haut-parleur externe, le casque écouteur et les lignes à distance. Dans ce dernier cas la sortie est limitée pour permettre le connexion directe aux lignes et elle comprend une commande de niveau à part.

Fondamentalement, le récepteur 990 S fonctionne sur alimentation en tension continue de 12 V et il comprend un bloc d'alimentation en tension alternative de secteur. Le récepteur peut être entraîné directement à partir d'une source de tension continue appropriée (pile sèche ou accumulateur), ce qui constitue un avantage lorsque des applications mobiles ou de champ sont envisagées.

Parmi les autres particularités intéressantes du récepteur il faut mentionner le mécanisme d'entraînement lent par engrenage, qui est un appareil standard de précision Eddystone; les échelles d'accord; le calibreur à cristal; l'instrument d'accord de panneau; les commandes de gain moyenne fréquence et haute fréquence indépendantes étalonnées directement en décibels; les dimensions compactes et le poids léger de l'appareil.

Le récepteur est fourni pour montage sur table mais il peut être aisément converti pour le montage sur bâti standard.

EE 95 764 *pour plus amples renseignements*

DÉTECTEUR DE FUITES À ULTRASONS

Dawe Instruments Ltd, Western Avenue, Acton, London, W.3

(Illustration à la page 478)

De l'énergie à ultrasons est produite par le gaz ou la vapeur en fuite de n'importe quel conteneur préssurisé. La quantité produite dépend de la vitesse du gaz en fuite, cette dernière dépendant à son tour de la pression et des dimensions de l'orifice. Dans tous les cas, cependant, il se produit un grand spectre de fréquence dans la région de 40 kHz, qui peut être utilisé pour déceler une fuite même en cas de forts bruits audibles provenant d'autres sources.

Ce principe est utilisé dans le nouveau détecteur de fuites à ultrasons Dawe (type 1873), un dispositif manuel qui n'est pas plus grand qu'une lampe éclair ordinaire. Le détecteur se compose d'un transducteur de type microphonique en céramique, fortement directionnel et accouplé à un amplificateur à transistors accordé (40 kHz $\pm$4 kHz). La sortie de l'amplificateur est indiquée visuellement sur un instrument de mesure à cadre mobile ou de manière audible au moyen d'écouteurs amovibles. En dehors des écouteurs, l'ensemble du détecteur se trouve dans un robuste coffret en aluminium mesurant 5 cm de diamètre $\times$ 34,2 cm de long. Le poids complet, avec la batterie, est de 1,8 kg. Une seule batterie suffit pour au moins 150 heures de fonctionnement continu et un dispositif de contrôle de la batterie est prévu.

En cours de fonctionnement, le détecteur est simplement dirigé à la main pour obtenir une indication de mesure maxima ou un signal audible. La fuite est alors dans la direction indiquée et elle peut être décelée avec précision en la suivant.

La sensibilité peut être réglée par une commande de volume qui élève ou abaisse le niveau de sortie, permettant ainsi le dépistage au voisinage immédiat d'une perte à vitesse élevée ou à 13,71 m d'un orifice de 0,25 mm de diamètre à une pression inférieure à 0,680 atmosphères. Lorsque la pression est suffisante, le diamètre de l'orifice peut se réduire à 0,002 pouce.

Bien que l'appareil ait été principalement conçu pour la détection de fuites, il peut également détecter n'importe quelle source de 40 kHz telle qu'un palier usé, une turbulence (hydraulique ou pneumatique), un frottement, etc. Des sondes spéciales peuvent être nécessaires dans ces cas-là.

EE 95 765 *pour plus amples renseignements*

BATTERIE MICROMINIATURE

Mallory Batteries Ltd, Gatwick Road, Crawley, Sussex

Une nouvelle batterie à oxyde de mercure ne mesurant que 0,220 pouce de diamètre, 0,130 pouce de hauteur et ne pesant que 0,3 g est fournie à présent par la société Mallory Batteries Ltd. Cette batterie, type RM-212, a une puissance nominale de 1,35 V et possède les caractéristiques de longue durée de vie, de stabilité de tension et de décharge stable qui ont fait le renom de ce type de batterie.

Cette cellule a été conçue spécifiquement pour un nouveau type d'appareil de surdité qui s'insère entièrement dans le conduit auditif mais qui est également propre à d'autres utilisations, telles que les "pilules radio" et d'autres sondes de diagnostic.

EE 95 766 *pour plus amples renseignements*

RELAIS BISTABLE

P.A.R. Ltd, Talbot Works, Talbot Street, Nottingham

(Illustration à la page 478)

Ce nouveau relais bistable comprend quatre contacts à permutation qui sont mécaniquement verrouillés en position au moyen de deux induits dynamiquement équilibrés. Le relais est à interverrouillage et à réenclenchement électrique.

La construction robuste du relais lui permet de soutenir les chocs les plus sévères dans n'importe quel plan sans verrouiller ou déverrouiller son loquet.

Cet avantage est d'une valeur particulière pour les fabricants de machines à jetons, de dispositifs d'alarme anti-vol, d'enregistreurs de temps à signaux, de dispositifs de comptage ou de machines mobiles ou vibrantes.

D'autres utilisations typiques peuvent être trouvées dans les commandes industrielles et les dispositifs à mémoire ou un signal momentané doit pouvoir fournir une indication continue avec facilité de réenclenchement.

Le relais est compact, ne mesurant que 2,5 cm × 6,98 cm × 5,08 cm de hauteur, y compris le couvercle tranparent standard. Les cosses de contact et de bobine ont été étudiées pour les applications à circuit imprimé.

EE 95 767 *pour plus amples renseignements*

COMMUTATEUR DE PROGRAMMATION

Sealectro Ltd, Walton Road, Farlington, Portsmouth, Hampshire

(Illustration à la page 478)

Le nouveau commutateur de programmation Actan à double tambour qu'a réalisé la société Sealectro a une capacité de programmation à très long terme. Il comprend un tambour d'entraînement dont les vitesses de rotation vont de 5 tours/minute à 1 tour par 28 jours. Le tambour secondaire se déplace d'un degré à chaque rotation de 360° du tambour d'entraînement fournissant la fonction de contrôle à terme extrêmement long. L'élément que l'on voit dans notre gravure comporte 36 positions de programme et 16 positions de contacts indépendantes par tambour. Il existe d'autres éléments ayant jusqu'à 120 positions de programme et 57 positions de contact indépendantes par tambour. Le nouveau commutateur de programmation peut être également mis en action à l'aide d'un mécanisme à solénoïde commandé à distance pour fournir les fonctions de comptage et de gradation. Les mécanismes d'activation à contact glissant assurent le réglage de champ simple et rapide des programmes sans qu'il y ait lieu d'utiliser des outils spéciaux.

Des contacts spéciaux de 10 am sont également prévus pour répondre aux applications de commandes spéciales.

En plus de ce nouveau dispositif de programmation à long terme, la société Sealectro fournit une gamme complète de commutateurs de programmation comportant des tambours amovibles, des contacts entassés, des contacts réglables, des positions de contact multiples et des éléments à programme fixe.

EE 95 768 *pour plus amples renseignements*

APPAREILS DE REPORT

Servomex Controls Ltd, Crowborough, Sussex

(Illustration à la page 478)

La société Servomex Controls Ltd a réalisé un appareil de report perfectionné pour la méthode analogique de résistance Teledeltos, conçu pour résoudre les problèmes pratiques d'électrostatique, de magnétisme, de mécanique, de transfert thermique et d'hydraulique.

L'appareil de report est livrable en deux modèles, à savoir le modèle FP130, entraîné par une batterie de 9 V et le modèle FP 144 fonctionnant sur courant secteur. Les deux modèles mesurent le potentiel sur le terrain par une méthode potentiométrique qui utilise également la tension de source pour le curseur, de sorte que la mesure n'est pas affectée par les variations de tension d'alimentation.

Cette méthode utilise le papier Teledeltos qui est d'un prix relativement réduit et d'une résistivité appropriée, de sorte que l'on peut effectuer le tracé analogique et exécuter le report sur le terrain en moins de 30 minutes. Une peinture d'argent spéciale d'un type perfectionné est utilisée pour peindre les équipotentiels. Une tension électrique provenant de l'appareil est appliquée entre les électrodes et le potentiel est tracé par le potentiomètre.

Le modèle FP 130 à fonctionnement sur batterie est utile aux étudiants car il comporte un potentiomètre à un seul tour de bonne qualité donnant une précision de ±1 % mais un dispositif d'étalement de bande électrique permet une plus grande discrimination dans la gamme de 0 à 10 %. Une fiche et une douille sont fournies afin de pouvoir y relier n'importe quelle alimentation jusqu'à 100 V c.c.

Le modèle FP 144 à fonctionnement sur courant secteur a une faible tension de 5 à 10 V pour les champs de Laplace ordinaires et une tension élevée de 50 à 100 V pour injecter un courant constant à travers des résistances à valeur élevée à des points du champ pour les champs poissoniens de sorte que le flux de courant est pratiquement indépendant du reste du réseau. Ce modèle utilise un potentiomètre de bonne qualité à 10 tours avec un cadran à lecture directe dont la lecture peut aller jusqu'à 0,1 %.

Les appareils de report de champ Servomex sont peu coûteux, réduisent les pertes de temps et sont à l'épreuve des court-circuits. Ils sont compacts et facilement portatifs, la version fonctionnant sur batterie ne pesant que 0,5 kg environ tandis que le modèle secteur pèse un peu moins de 2 kg.

EE 95 769 *pour plus amples renseignements*

FER À DESSOUDER

Distributeurs: Soldering Instrument Components Ltd, Adcola House, Gauden Road, London, S.W.4

(Illustration à la page 479)

La société Adcola vient de réaliser un fer à souder, type M207, dont l'utilisation principale est la suppression des soudures de cosses, composants et plaquettes de circuit imprimé. Cet instrument est fabriqué par la Adcola Product Pty. Ltd. de Melbourne et il est vendu par la Soldering Instrument Components Ltd., une filiale de Adcola Products Ltd.

L'instrument est muni d'une pointe creuse doublée d'acier inoxydable qui peut être utilisée tant pour le soudage que pour les dessoudage. La petite protubérance sur la poignée, qui permet la mise en oeuvre avec le bout du doigt, extrait la soudure à travers la pointe et un filtre placé à l'autre extrémité de la pointe recueille la soudure qui peut ensuite être projetée au dehors en pressant la protubérance. Le filtre peut être retiré de la chambre à filtre pour être nettoyé.

Une gamme de 12 pointes peut être utilisée avec le fer M207, dont six pointes creuses pour le dessoudage et six pointes solides. La gamme comprend des pointes à diamètres de face de 0,15 cm, 0,23 cm, 0,31 cm, 0,39 cm, 0,47 cm et 0,63 cm, la pointe creuse de 0,31 cm constituant un accessoire standard. Les pointes sont à revêtement métallique pour empêcher le grippage dans le barillet et pour prolonger la durée de vie de la face.

Les éléments de chauffe, les pointes de soudage et de dessoudage, les filtres et les chambres à filtre sont en stock auprès de la société Soldering Instrument Components Ltd qui fournit également les services d'entretien pour le Royaume-Uni. Les instruments peuvent être fournis pour n'importe quelle gamme de tension entre 6 et 250 V.

EE 95 770 *pour plus amples renseignements*

ANALYSEURS DE BRUIT MICROONDES

James Scott (Electronic Engineering) Ltd, Carntyre Industrial Estate, Glasgow, E.2

La société James Scott (Electronic Engineering) Ltd vient de créer une nouvelle gamme d'instruments perfectionnés de mesure de bruit de grande puissance, type 123, pouvant effectuer des mesures encore plus sensibles que celles qu'effectuaient les appareils précédents.

Ainsi avec les appareils précédents la sensibilité pour le bruit modulé en fréquence et en amplitude dans 150 kHz à partir du signal porteur microondes, exprimée comme rapport de puissance signal/bruit pour la puissance de bande latérale totale et mesurée dans une largeur de bande de 100 Hz, était comme suit:

Fréquence de modulation	1 kHz	10 kHz	100 kHz	150 kHz
Modulation d'amplitude	125 dB	134 dB	135 dB	135 dB

La sensibilité de modulation de fréquence obtenue était de 0,0025 Hz efficace de déviation.

Les nouveaux appareils peuvent réaliser la sensibilité suivante:

Fréquence de modulation	1 kHz	10 kHz	100 kHz	150 kHz
Modulation d'amplitude	125 dB	143 dB	163 dB	168 dB

La sensibilité de modulation de fréquence réalisée est donc de 0,0008 Hz efficace de déviation.

Ces chiffres ont été obtenus par l'emploi d'une nouvelle forme de construction qui permet d'obtenir un échantillon de puissance plus élevé de la source soumise à l'essai. En plus, les effets de bruit acoustique et de vibrations ont été minimisés. Des circuits complexes de réaction ont été utilisés pour:

(a) Maintenir la cavité de pont en équilibre. Un nouveau système breveté comprenant un élément asservi à thermo-syphon est utilisé.

(b) Permettre d'effectuer des mesures de bruit sur certains porteurs modulés en fréquence par les méthodes dites d'annulation.

(c) Eliminer les faux effets de modulation de l'alimentation de la source soumise à l'essai.

(d) Verrouiller en fréquence l'oscillateur local.

Les perfectionnement qui ont donc été portés à l'appareil d'analyse permettent l'examen du spectre de bruit entre 50 Hz et 1 kHz à partir du porteur de micro-ondes dans une largeur de bande de 30 Hz.

Ces appareils sont déjà livrables dans les versions à bande J et X et ils seront prochainement livrables pour les bandes C et S.

EE 95 771 *pour plus amples renseignements*

AMPLIFICATEUR À GAIN ÉLEVÉ

S.E. Laboratories (Engineering) Ltd. North Feltham Trading Estate, Feltham, Middlesex

(Illustration à la page 479)

La société S.E. Laboratories Engineering Ltd a créé un amplificateur à gain élevé et à faible bruit, le SE. 436 à réponse linéaire de 5 Hz à 10 kHz. Son utilisation principale est en liaison avec des transducteurs électriques/mécaniques sensibles aux amplitudes, comme par exemple les extensomètres amorcés au courant continu, les capteurs de vitesse, etc.

Deux réseaux intégrateurs à constantes de temps différentes sont incorporés, en outre, dans l'amplificateur intégrateur/linéaire fourni en variante, le SE.436/I.

Lorsqu'ils sont mis en circuit, l'un ou l'autre de ces amplificateurs permet d'intégrer le signal de sortie d'un capteur de vitesse pour l'enregistrement ou l'indication subséquente de l'amplitude de vibration.

La sortie de ces amplificateurs peut entraîner des galvanomètres d'enregistrement à haute fréquence, des magnétophones, des oscilloscopes, etc. Les circuits de réaction multiple assurent un degré élevé de stabilité de gain et de rejet en phase.

Des sorties de tension et de courant séparées sont prévues. Grâce à la faible impédance de source de la sortie de tension, l'amplificateur peut être utilisé pour tous les types de matériel d'affichage ou d'enregistrement sensibles aux tensions.

EE 95 772 *pour plus amples renseignements*

WOBBULATEUR U.H.F.

Compagnie Générale de Métrologie, B.P.30, Annecy, France

(Illustration à la page 479)

Le Wobbulateur U.H.F. type 241A, utilisé en liaison avec un oscilloscope et un générateur marqueur, constitue un banc traceur de courbes. Il peut être utilisé pour déterminer la courbe de réponse de fréquence d'amplitude de circuits actifs ou passifs fonctionnant dans la gamme de fréquence de 470 à 870 MHz.

Ces applications comprennent le réglage d'amplificateurs U.H.F., la détermination de la courbe générale d'un récepteur de télévision (chaînes IV et V) de l'antenne à l'image, la mesure du rapport d'amplitude d'ondes stationnaires d'antennes, de câbles coaxiaux et d'autres circuits, en liaison avec un pont accordé type HCX35A et un amplificateur BF type HCX52A.

Pour cette utilisation, le Wobbulateur fournit, à travers un oscillateur à ligne de Lecher, une tension U.H.F. modulée en fréquence et enregistrée de manière continue sur un cadran gradué en MHz.

La fréquence est modulée au moyen d'un condensateur vibrant monté sur un support rectangulaire à ressort de bronze. Ce dispositif a l'avantage de permettre un déplacement linéaire tout en assurant une construction générale robuste.

Afin d'éliminer la modulation d'amplitude parasitique et de préserver un niveau de sortie constant et indépendant de la fréquence affichée, un circuit asservi règle la tension en agissant sur l'alimentation de l'oscillateur.

Le Wobbulateur type 241A a une sortie U.H.F. supplémentaire qui permet de marquer la courbe affichée.

Le Wobbulateur peut être fourni en différents modèles pour répondre aux besoins du client en ce qui concerne la sortie d'impédance et l'aspect général.

EE 95 773 *pour plus amples renseignements*

Résumés des Principaux Articles

Moteur élévateur sans balais à c.c. pour la conversion de données binaires en données décimales par W. Leonchick

Résumé de l'article aux pages 426 à 429

Une machine à traduire utilisant un principe vectoriel inédit pour obtenir une lecture décimale directe de données binaires est décrite dans cet article. Pour plus de clarté, le processus est indiqué en détail.

Etant donné que les moteurs élévateurs du type sans contact sont essentiellement des dispositifs numériques d'entrée/sortie, la seconde partie de l'article traite de l'utilisation de la machine comme mécanisme asservi à degrés à courant continu et à circuit ouvert. Les avantages qui en découlent sont: (a) la fiabilité, (b) le fonctionnement sans entretien et (c) moins d'espace et de poids par rapport aux circuits asservis à c.a. équivalent. Le système décrit est insensible aux variations des conditions d'environnement.

La conversion de tension numérique en tension analogique, ainsi que la nécessité d'employer un matériel de traduction de données binaires en données décimales sont éliminées grâce à la souplesse d'utilisation de la réalisation électromécanique.

Une porte pour la modulation d'amplificateurs de pré-perception par G. N. Hunter

Résumé de l'article aux pages 432 à 435

L'auteur de cet article s'attache à démontrer l'utilité de la porte M.O.S.T. pour les circuits à faible niveau. Lorsqu'on module l'entrée d'un amplificateur de perception dans une mémoire de calculatrice, le problème consiste à rejeter des impulsions de bruit élevé en très peu de tempts avant de permettre la transmission de signaux à faible niveau. L'article examine un circuit de porte pratique et montre que sa performance, avec des échantillons de courant de M.O.S.T., est compatible avec temps de cycles de mémoire jusqu'à 1µsec.

Nouveaux développements dans la réalisation des circuits de commutation à combinaisons minimales NI pour variables N par G. W. Copperwhite et D. Zissos

Résumé de l'article aux pages 436 à 442

Les auteurs décrivent une méthode pas-à-pas entièrement algébrique pour la réalisation de circuits de commutation à combinaisons minimales NI. Il s'agit essentiellement d'une méthode d'optimisation de l'expression algébrique de la fonction devant être réalisée en trois degrés. La justification théorique des degrés de réalisation est indiquée. Plusieurs exemples sont donnés pour illustrer la méthode, y compris un dispositif complet pour additions/soustractions.

Un amplificateur "en phase" et ses applications quant aux stabilisateurs de tension par G. N. Patchett

Résumé de l'article aux pages 443 à 445

Le dispositif décrit a été réalisé pour réduire l'angle de phase entre la tension d'entrée et celle de sortie d'un stabilisateur de tension alternative de précision. L'appareil mesure la phase de la tension de sortie par rapport à une référence. S'il se produit une erreur de phase, une tension de quadrature est injectée aussitôt au stabilisateur pour annuler cette composante de quadrature de la tension de sortie. Cet appareil permet de réduire l'erreur de phase à 15' dans toutes les conditions de facteur de charge et de puissance.

Rectification d'impédance dans la ligne d'alimentation verticale d'une antenne à gain élevé par A. B. Shone

Résumé de l'article aux pages 446 à 448

Il est de pratique courante au Royaume-Uni de réunir les émissions dans la bande UHF et dans de nombreux cas il est plus rentable d'utiliser l'ensemble de l'ouverture d'antenne disponible pour une antenne commune portant jusqu'à quatre canaux UHF. Cet arrangement pose le problème d'obtenir l'équilibrage d'impédance voulu dans une grande largeur de bande et également—compte tenu en particulier des émissions en couleur—d'obtenir un dessin vertical uniforme qui devrait, autant que possible, se conformer à une loi de cosécantes. Cet article suggère un moyen éventuel d'obtenir ces deux résultats. Leur application dépendrait de l'inclinaison de faisceau inhérente à ce genre de solution.

Amplificateur à résistance d'entrée infinie par J. B. Matley

Résumé de l'article aux pages 449 à 451

Cet article présente et analyse un amplificateur à deux étages à gain de tension unitaire utilisant un transistor à effet de champ dans le premier étage pouvant produire une résistance d'entrée bien au dessus de 100 mégohms. Une modification de ce circuit est étudiée qui permet de régler la résistance d'entrée à l'infini. Le circuit ainsi modifié est ensuite analysé.

Un bloc d'alimentation réversible de 10kV par M. C. Jackson

Résumé de l'article aux pages 452 à 453

Ce bloc d'alimentation stable et sûre de ±10kV se compose de circuits constitués de corps solides et il comporte un disjoncteur à maxima. Il est, en outre, conforme au modèle Harwell de la série 2000. Il a été conçu pour fournir les potentiels d'accélération et de concentration d'une source d'ions d'un spectromètre de masse. La stabilité de la tension de sortie est supérieure à 4 parties dans 10^6 par heure.

Circuits logiques analogiques et de commutation de relais par C. D. Lewis

Résumé de l'article aux pages 454 à 457

Cet article décrit une série de circuits de calcul analogique comportant un ou deux amplificateurs opérationnels. Ces circuits simulent les fonctions binaires telles que NON, ET, OU, SI, ALORS, MAIS, etc. . .

L'électronique intégrée et son rôle dans l'enseignement par K. J. Dean

Résumé de l'article aux pages 458 à 460

L'auteur se penche dans cet article sur quelques uns des plus récents développements dans le domaine de l'électronique ainsi que sur ses incidences sur l'enseignement. Il décrit des appareils permettant d'étudier en laboratoire les applications des circuits intégrés.

Redressement à l'aide de dispositifs à caractéristiques symétriques par J. F. Young

Résumé de l'article aux pages 461 à 463

L'auteur démontre qu'un dispositif ayant une caractéristique symétrique de tension-courant peut faire fonction de redresseur d'impulsions ou de formes d'ondes à harmoniques déformées. Il décrit les caractéristiques d'un tel redresseur d'impulsions symétriques et montre que la sortie redressée dépend du rapport marque/espace plutôt que de l'amplitude de l'impulsion d'entrée. Des paires de diodes Zéner ont été utilisées comme redresseurs symétriques bien que d'autres dispositifs aient des caractéristiques appropriées. Certaines possibilités d'application du redresseur symétrique sont étudiées.

Une minuterie automatique simple pour agrandisseur photographique par J. L. Linsley Hood

Résumé de l'article aux pages 464 à 466

Le circuit décrit dans cet article utilise une cellule photoconductrice pour mesurer la brillance de l'image projetée et pour contrôler, au moyen d'un circuit intégrateur simple, le temps de pose. Des temps de pose de 1sec à 40sec sont prévus.

Un régulateur de courant série par T. M. Palmer

Résumé de l'article aux pages 467 à 469

Il s'agit dans cet article d'une variante fort simple d'une alimentation de courant stabilisé qui consiste en un élément série autonome qui fait passer un courant constant indépendant de la tension appliquée.

NEUE AUSRÜSTUNGEN

Übersetzung der Seiten 474 bis 479

Beschreibung neuer Bauelemente, Zubehörteile und Prüfgeräte auf Grund der von Herstellern gemachten Angaben.

Elektrolytkondensatoren

The Telegraph Condenser Co. Ltd, Wales Farm Road, North Acton, London, W.3

(Abbildung Seite 474)

Aluminiumfolie - Elektrolytkondensatoren in neuer Bauart wurden vor kurzem von der Telegraph Condenser Co. Ltd, einer Mitgliedsfirma der Plessey-Gruppe, angekündigt. Sie arbeiten zur vollen Zufriedenheit innerhalb eines Temperaturbereiches von $-55°C \ldots +125°C$.

Diese TCC-Kondensatoren ETR (Extended Temperature Range = Erweiterter Temperaturbereich) werden in England unter einer exklusiven Lizenz der kanadischen Regierung gefertigt, und zwar in einer Auswahl von Kapazitäten und Grössen, die den US-Vorschriften MIL-C-39018 entsprechen. Ein entsprechendes britisches Pflichtenblatt ist in Vorbereitung, und Kondensatoren werden für Typenzulassung vorgelegt werden.

ETR-Kondensatoren wiegen etwa ein Drittel der entsprechenden Tantalfolien-Kondensatoren und haben eine höhere Kapazität je Volumeneinheit. Ihre elektrische Leistung entspricht fast der von Tantalkondensatoren, und sie werden ähnlichen Pflichtenblättern genügen. Der Fall der Kapazität bei $-55°C$ ist gering, und die elektrischen Eigenschaften bleiben bei Lebensdauertests von über 2 000 Stunden bei $+125°C$ sehr konstant.

Bei Leerlauf sind die Kenndaten wesentlich überlegen, und der Ersatzserienwiderstand ist über einen breiten Temperaturbereich viel niedriger als für die herkömmlichen Elektrolytkondensatoren mit Aluminiumfolie. Dadurch sind die Kondensatoren für Betrieb in Impulsschaltungen sehr geeignet, bei dem auf Beibehaltung der Impulsform Wert gelegt wird.

Ein grosser Teil der Verbesserungen ist auf Verwendung eines neuen Elektrolyten zurückzuführen, der Dimethylformamid als Lösemittel enthält und eine neue Bauart und Abdichtung erfordert.

Die zylindrischen Gehäuse werden zur Bewegungsbegrenzung des Elementes mit Extrarillen versehen, um den Vibrationsanforderungen zu genügen. Ein anderes interessantes Konstruktionsmerkmal ist ein mechanisch geschwächter Teil des Gehäuses, der das Gehäuse ohne Explosion platzen lässt, wenn—z.B. durch versehentliches Umpolen eines polarisierten Kondensators—hoher interner Gasdruck erzeugt wird.

Lebensdauertests mit ETR-Kondensatoren bei Nennspannung und Höchsttemperatur haben die bemerkenswerte Stabilität der Parameter bewiesen; die prozentualen Änderungen von Kapazität, Verlustfaktor und Reststrom lagen nach 2000 Stunden wesentlich unter den nach dem MIL-C-39018-Entwurf zugelassenen.

EE 95 751 *für weitere Einzelheiten*

UV-Schreiber

Southern Instruments Ltd, Camberley, Surrey

(Abbildung Seite 474)

Die neuste Ergänzung des Southern Instruments-Programms für ultraviolette Schreiber ist das Modell M1310, ein sehr stabil gebautes Instrument, das viele eingebaute Einrichtungen bietet.

Das Standardgerät ist im Tischgehäuse oder für Gestelleinbau lieferbar, kann 1 ... 50 Kanäle aufnehmen und auf jeder Papierbreite bis zu 305 mm registrieren. Zwölf durch Drucktasten gewählte Papiervorschübe geben Geschwindigkeiten von 4 bis 2500 mm/s. Diese Auswahl kann auf 24 Vorschübe im Bereich 0,004 bis 2500 mm/s, oder mit dem ultraschnellen Gerät um weitere vier Vorschübe auf 25 000 mm/s erweitert werden.

Die Optik gewährleistet, dass Defokussierung am Rande des Registrierpapiers selbst bei 305 mm Breite vernachlässigbar klein ist. Durch Anwendung moderner Schaltungstechnik wird der Lampe immer konstante Energie zugeführt, so dass sie immer unter idealen Betriebsbedingungen arbeitet.

Weitere besondere Merkmale sind Ereignismarken, Einrichtungen zur Wiederholung von Vorgängen, optische Rasterlinien, kristallgesteuerte Zeitlinien über die volle Breite, Kennzeichnung der Spuren, Numerierung von Spur und Vorgang, sowie Fernsteuerung.

EE 95 752 *für weitere Einzelheiten*

Mehrfachmessgerät

Vertrieb: Kelvin Electronics Co. (Smiths Industries Ltd), Kelvin House, Wembley Park Drive, Wembley, Middlesex

(Abbildung Seite 474)

Das "Digikelvo" ist ein Mehrfachmessgerät, das auf Zählröhren eine dreistellige Digitalanzeige der gemessenen Werte und Polarität gibt. Das Instrument hat vier getrennte Eingangskanäle für Gleich- und Wechselspannung, Strom und Widerstand. Zwei Kanäle sind für Spannungsmessungen von 1 mV ... 1 kV in sechs Bereichen belegt, der dritte für Strommessungen von 100 nA ... 999 mA in sieben Bereichen und—mit Hilfe eines Nebenschlusses—bis zu 99,9 A in zwei Bereichen. Der vierte Kanal wird zum Messen von Widerständen von 1 Ω ... 999 MΩ in sieben Bereichen benutzt. Automatische Bereichumschaltung wird durch Verschieben des Kommas in der Anzeige erreicht. Ausserdem gibt es noch einen Schalter mit zwei Positionen für Grobeinstellung, so dass sechs Grössenordnungen überstrichen werden.

Das System besteht aus einem Hochleistungs-Gleichstromverstärker mit niedriger Drift, dessen Ausgang durch eine Treppenschaltung kompensiert wird, und einem Zähler, der die Anzahl der zur Nullung erforderlichen Inkremente anzeigt. Die in der Treppenschaltung und den Zählschaltungen angewendeten Prinzipien sind herkömmlich, doch ist der Hochleistungs-Gleichstromverstärker in vieler Hinsicht ungewöhnlich.

Der Gleichstromverstärker hat eine auf den Eingang bezogene kurzzeitige Drift von $0,5\ \mu V/°C$, und die Langzeitdrift wird auf $2,5\ \mu V$ pro Jahr geschätzt. Die Leerlaufverstärkung ist 2×10^6 und wird durch Gegenkopplung in der Verknüpfung für die höchste Verstärkung auf 14 000 reduziert. Die Bandbreite des Verstärkers ist 600 kHz, jedoch wird sie durch andere Eigenschaften des Instruments auf insgesamt 100 kHz reduziert. Diese Bandbreite gibt für das Gerät eine maximale Messperiode von 0,3 s. Die Linearität des Verstärkers ist besser als 0,1 Prozent und gibt dem Instrument eine Messunsicherheit von ± 1 Ziffer in der letzten Stelle. Viele dieser Eigenschaften sind auf Verwendung eines FET als Chopper im Verstärkereingang zurückzuführen.

Durch Anwendung von Druckschaltungsplatten und geeignete Steckverbindungen entfällt interne Verdrahtung im "Digikelvo" völlig.

EE 95 753 *für weitere Einzelheiten*

HF-Störsperren

Erie Resistor Ltd, South Denes, Great Yarmouth, Norfolk

Erie Resistor Ltd gibt die Zulassung von zwei Modellen der HF-Breitbandstöresperren Serie 1200 für die Entstörung von Systemen in Militärflugzeugen seitens des britischen Luftfahrtministeriums bekannt. Es handelt sich um die Modelle 1200-025 und 1215-000.

HF-Breitbandstörsperren der Serie 1200 bestehen aus einem Keramikkondensator von 1,5 μF (nach Erie patentierten Verfahren hergestellt) und Ferritperlen zur Erzeugung des induktiven Serieneffektes (ohmsche Verluste bei HF). Sie sind hermetisch dicht in ein mit Gewindebuchsenteil versehenes, robustes Metallgehäuse eingebaut. Die mechanische Anordnung entspricht einer Durchführung, für die es verschiedene, wahlweise Abschlüsse gibt. Erie-HF-Breitbandstörsperren sollen in ihrer Leistungs- und Preisklasse die kleinsten in der Welt kommerziell greifbaren Einzelbausteine sein.

Die Serie 1200 besteht insofern aus Breitbandsperren, als sie bei Betriebsfrequenzen von 5 kHz bis zu 1 GHz—und sogar höher—eine nützliche Dämpfung geben. Die Höchstdämpfung ist in Richtung des induktivitätsbehafteten Serieneingangs, da es sich hier im wesentlichen um L-Glieder handelt. Über den grössten Teil des Hochfrequenzspektrums ist die Mindestdämpfung—einschliesslich etwa vorkommender Resonanzkurvensenken—70 dB. Die Dauer-Arbeitsspannung ist 50 V—mit etwa 50% Minderung für Betrieb bei 125° C. Die Modelle genügen den Betriebsbedingungen nach der Norm BS.2G.100.

EE 95 754 *für weitere Einzelheiten*

Spannungsdriftmonitor

J. & P. Engineering (Reading) Ltd, Portman House, Cardiff Road, Reading, Berkshire

(Abbildung Seite 475)

Der Driftmonitor besteht aus Bezugsspannung, Spannungsteiler und Abweichungsverstärker in einem temperaturkonstanten Gehäuse. Er ist in erster Linie für die Bestimmung der Drift von Stromversorgungen mit Spannungsbereichen von 0,1 V...5 kV bei Strömen von 100 μA...5 A bestimmt. Konstanz und Empfindlichkeit sind für Anwendungen dieser Art von ausschlaggebender Bedeutung.

Die dreistufige Temperaturregelung des Driftmonitors gibt ihm einen Gesamtkoeffizienten von besser als $0{,}2 \times 10^{-6}$. Die Bezugsspannung wird von einer temperaturkompensierten Zenerdiode geliefert, die durch einen hochkonstanten Stromerzeuger getrieben wird. Die wichtigsten Teilerwiderstände sind zur Erreichung langfristiger Konstanz lose gewickelt, haben einen sehr niedrigen Temperaturkoeffizienten und sind in Gehäuse eingebaut, deren Temperatur konstant gehalten wird. Die Luft in diesen Gehäusen wird durch einen leistungsfähigen Ventilator umgewälzt, so dass bei Beaufschlagen mit einem Eingang der Temperaturanstieg nur gering ist.

Konstanz und Empfindlichkeit des Driftmonitors gestatten das Registrieren von Drifts von nur wenigen Millionenteilen über mehrere Tage und sind praktisch von Spannungs- und Temperaturschwankungen unabhängig. Das volltransistorierte Gerät ist in ein Gehäuse für 19″-Gestelle mit 267 mm hoher Frontplatte eingebaut. Es wurde von J. & P. Engineering (Reading) Ltd zur Prüfung von Strom- und Spannungsquellen der britischen Atombehörde in Harwell entwickelt.

EE 95 755 *für weitere Einzelheiten*

Elektronisches Gleichstrom-Messgerät

Claude Lyons Ltd, Valley Works, Hoddesdon, Hertfordshire

(Abbildung Seite 475)

Das Modell MV-07C ist ein empfindliches elektronisches Volt- und Amperemeter, das so niedrige Werte wie 1 μV— und 1 nA— messen kann. Es hat insgesamt 34 Spannungs- und Strombereiche in einer 10-25-100-Folge mit Skalenendwerten bis zu 1 kV— und 1 mA—. Für die meisten Spannungsbereiche bis zu und einschliesslich 1 kV beträgt die Messunsicherheit 1 Prozent des Bereichendwertes.

Zu den besonderen Merkmalen des Gerätes gehören u.a. der geräuscharme Chopperverstärker, sorgfältiger Entwurf der Eingangsschaltung zur Vermeidung von Verbindungen unähnlicher Metalle, unabhängige Nacheichung für alle 17 Spannungsbereiche, sowie ein getrennt geeichtes 152-mm-Spiegelgalvonameter mit Spannbandaufhängung und Nullpunkt in der Mitte.

Der empfindliche, geräuscharme Trägerverstärker enthält einen geräuscharmen Chopper mit geringer Drift, dem eine geräuscharme Kaskodenstufe nachgeschaltet ist. Dieser Verstärker hat zur Erhöhung des Signalabstands eine schmale Bandbreite, die durch Einschalten einer Doppel-T-Schaltung in eine der Wechselstrom-Rückkopplungsschleifen erreicht wird. Hohe Wechselstrom-Gesamtrückkopplung stabilisiert den Trägerverstärker gegen Netzschwankungen und Änderungen in den Kenndaten der Bauelemente.

Ein transistorierter Synchrondetektor erzeugt einen Eingangsspannung oder -strom proportionalen Gleichstromausgang, der mit dem Eingang gleichphasig ist und gestattet, auf einem Messgerät mit beiderseitigem Ausschlag positive und negative Eingänge ohne Polaritätsschalter anzuzeigen. Der Ausgang wird ausserdem einer Steckverbindung zugeführt, der man für externe Anwendung ein verstärktes phasengleiches Signal entnehmen kann.

Ausgezeichnete Verstärkungskonstanz wird durch eine einstellbare Gleichstrom-Rückkopplungsschleife erreicht, die durch den Bereichschalter geht und auch unabhängig Nacheichregler für alle 17 Spannungsbereiche und die 6 empfindlichsten Strombereiche enthält.

Claude Lyons Ltd baut dieses Gerät in England unter einer exklusiven Lizenz der Millivac Instruments Inc, U.S.A.

EE 95 756 *für weitere Einzelheiten*

Röhrenvoltmeter

Advance Electronics Ltd, Roebuck Road, Hainault, Ilford, Essex

(Abbildung Seite 476)

Modell VM76 ist das neuste der fünf Voltmeter des Advance-Programms und erweitert es bis zu Wechselstromfrequenzen über 1 GHz und Gleichstrommessungen bis 1000 V, mit ± 2 Prozent Messunsicherheit bis zu 100 V.

Das Gerät ist für Drucktastenwahl der verschiedenen Funktionen ausgelegt. Bei Gleichspannungsmessungen kann auf diese Weise beiderseitiger Anschlag der Anzeige und positive oder negative Polarität eingestellt werden. Ausserdem kann man in gleicher Weise Widerstandsmessungen von 0,02 Ω...500 MΩ mit ± 5 Prozent Messunsicherheit in Skalenmitte wählen.

Der grösste Teil der Bauelemente ist auf Druckschaltungsplatten montiert, was die Bauelemente leicht zugänglich macht. Jedes Gerät wird mit Messspitzen sowie Verbindungsschnüren mit einpoliger Steckverbindung geliefert. Wahlweise gibt es auch Zubehör für alle Voltmeteranwendungen.

Das mit Netzanschluss arbeitende VM76 hat einen Temperaturbereich von 0...+40° C. Es ist wie alle in jüngster Zeit angekündigten Advance-Instrumente in ein robustes, modulares Gehäuse von 2/3 Gestellbreite eingebaut.

Die Skalenendwerte der Gleichspannungsbereiche sind 300 mV bis zu 1 kV mit über 100 MΩ Eingangsimpedanz. Die Wechselstrombereiche überstreichen 30 mV bis zu 300 V_{eff}. Die Eingangsimpedanz ist in der Grössenordnung von 7,5 MΩ bei 1 kHz und 150 kΩ bei 100 MHz.

EE 95 757 *für weitere Einzelheiten*

Gleichstromverstärker

Honeywell Controls Ltd, Honeywell House, Great West Road, Brentford, Middlesex

(Abbildung Seite 476)

Honeywell nimmt erstmalig in Grossbritannien den Vertrieb einer Reihe von Gleichstromverstärkern auf. Die mit Accudata 104 bezeichnete Baureihe ist in Standardausführung mit einem Verstärkungsfaktor zwischen 10 und 250 sowie in zwei anderen Ausführungen mit 0,1...2,5 und 0,4...10 Verstärkung lieferbar.

Die Verstärker haben ausgezeichnete Rausch- und Driftkenndaten; sie sind ideale Mehrzweckverstärker für gleichstromerregte Dehnungsmessstreifen, Thermoelemente oder andere Signalquellen, deren Leistung nicht für das direkte Treiben eines Registriergerätes reicht. Alle Ausführungen sind für Einsatz mit Galvanometerschreibern ausgelegt und haben eine selbst für das Treiben von flüssigkeitsgedämpften Galvanometern bei höchsten Frequenzen ausreichende Ausgangsleistung.

Beschädigung des Galvanometers durch Einschwingvorgänge beim Schalten von Messstreifen, Unterbrechung des Eingangsstromkreises oder beim Ein- und Ausschalten des Gerätes wird im Accudata 104 durch Schaltungsmittel verhindert. Ein Frontplattenregler bestimmt die Polarität der Galvanometerverbindungen und gestattet Kurzschliessen des Ausgangs für Nullung des Galvanometers.

EE 95 758 *für weitere Einzelheiten*

Drehkondensatoren

Wingrove & Rogers Ltd, Domville Road, Liverpool, 13

(Abbildung Seite 476)

Typ C.903 ist ein Lufttrimmer mit Feststelleinrichtung. Die keramische Grundplatte hat Aussenabmessungen von 20,3 × 15,2 mm und ist für Montage mit zwei Gewindebuchsen für 6-BA-Schrauben versehen. Er ist dem Typ C.803 ähnlich, jedoch kleiner.

Die Mindestkapazität ist 3,5 pF und der Kapazitätshub bis zu 34 pF. Der Verlustfaktor liegt bei 1 MHz unter 0,001, der Isolationswiderstand bei 500 V— höher als 1 GΩ.

EE 95 759 *für weitere Einzelheiten*

Plattenspeicherantrieb

Vertrieb: Scientific Furnishings Ltd, Poynton, Cheshire

(Abbildung Seite 476)

In Plattenspeicherantrieben der Control Data-Baureihe 9430 sind die wünschenswertesten Eigenschaften von Magnetband- und Magnetplattenspeicherung kombiniert: schneller wahlfreier Zugriff und unbegrenzte Aufbewahrung in Regalen. Das wird durch abnehmbare Plattenpakete ermöglicht, auf denen man speichern kann und die—wie Magnetbandspulen—unbegrenzt aufbewahrt und schnell wieder eingesetzt werden können. Innerhalb der Plattenspeicherantriebe derselben Modellnummer sind alle Plattenpakete frei austauschbar; auf einem Antrieb eingeschriebene Information kann auf einem anderen gelesen werden. Die Typen 9430 und 9434 sind ausserdem mit den IBM-Modellen 1311 bzw. 2311 kompatibel.

Die Baureihe 9430 kann direkt suchen, d.h. die letzte Adresse wird im Plattenspeicherantrieb gespeichert und steht dem Computer zur Bestimmung der Entfernung und Richtung der Kopfbewegung zur Verfügung. Dadurch werden bedeutend schnellere Zugriffszeiten als bei automatischer Rückkehr zur Ausgangsstellung möglich. Ein weiteres Kennzeichen aller Modelle ist das Suchüberlappen, nach dem mehr als ein Antrieb gleichzeitig suchen kann.

Es gibt vier Modelle für Zeichen mit 6, 7 oder 8 Bit, deren Gesamtkapazität bei 6 Platten je Paket zwischen 2 000 000 und 8 192 000 Zeichen liegt. Die Höchstzugriffzeit ist für alle Modelle 145 ms, die durchschnittliche Zeit 70 ms.

EE 95 760 *für weitere Einzelheiten*

Wheatstone-Vergleichsbrücke

British Physical Laboratories, Radlett, Hertfordshire

(Abbildung Seite 476)

Als Ergebnis der Erfahrungen von B.P.L. mit Wheatstone-Kleinlastbrücken, besonders der Brücke RM196, wurde von der Firma die Wheatstone-Gleichstrom-Kleinlastvergleichsbrücke RM197 entwickelt. Es ist ein in sich geschlossenes Instrument zum Messen von Widerständen in zwei Betriebsarten und besteht grundsätzlich aus den Wheatstone-Brückenelementen, d.h. Verhältniszweigen, einem veränderlichen Zweig, einer Brückenspeisung und einem Anzeigeverstärker.

Der veränderliche Zweig besteht aus vier Dekaden in diskreten Stufen mit Digitalanzeige des eingestellten Wertes. Die Grössenordnung der vier geltenden Ziffern ergibt sich aus der durch die Verhältniszweige bestimmten Kommastellung.

Die Brückenspeisung erfolgt mit konstantem Strom statt mit Konstantspannung. Diese Anordnung ergibt eine sinnvolle konstante Fehlerspannung für einen bestimmten Prozentsatz Unsymmetrie, was bei Betrieb als Vergleicher wünschenswert ist.

Der Anzeigeverstärker ist rauscharm, chopperstabilisiert mit hoher Verstärkung, dessen synchron gleichgerichteter Ausgang eine positive oder negative Unsymmetrieanzeige ergibt.

Bei Einsatz als Wheatstone-Messbrücke wird der unbekannte Widerstand an die Klemmen R_x angeschlossen und der veränderliche Zweig auf einen willkürlichen Zweig (nicht 0000) eingestellt; die Brücke wird dann erst mittels des Bereichschalters, dann der kennzeichnendsten Ziffer abgeglichen. Höchstempfindlichkeit wird in der 0,5-Prozent-Stellung des Endwertschalters erzielt. Der unbekannte Widerstandswert wird dann direkt von der Anzeige abgelesen.

Bei Anwendung als Vergleicher wird der Widerstandswert mittels Bereich- und Dekadenschalter auf der Anzeige eingestellt. Der Endwertschalter wird auf den gewünschten Vollausschlagbereich eingestellt, z.B. für das Klassieren von Widerständen mit 10 % Toleranz auf den Bereich für 20 % Vollausschlag. Der Widerstand wird dann an die R_x-Klemmen angeschlossen und die Abweichung vom Sollwert in Prozent abgelesen.

EE 95 761 *für weitere Einzelheiten*

Chopper-Verstärker

Kent Precision Electronics Ltd, Vale Road, Tonbridge, Kent

(Abbildung Seite 477)

Der K-PAC/4 wurde für den einwandfreien Betrieb von Standardinstrumentierungen mit 0...100 mV Eingangsspannungen von Wandlern und Thermoelementen usw. mit niedrigen Ausgangsleistungen. die Eingänge im Bereich 0...10 mV haben, entwickelt. Er besteht aus einem von KPE entwickelten Chopper-Verstärker mit geringer Drift, der mit einer in sich geschlossenen Stromversorgung in ein Standard-K-PAC-Gehäuse eingebaut wird. Ein Transistorchopper zugelassener Bauart moduliert das ankommende Millivoltsignal, das in einen Wechselstromverstärker mit konstanter Verstärkung gespeist wird, dessen Ausgang nach Demodulation ein Gleichstromausgangssignal abgibt. Durch Rückkopplung des Ausgangs auf den Eingang wird die Verstärkung auf einen Wert stabilisiert, der das Eingangssignal in einen Standardgleichstromausgang zwischen 0 und 100 mV für den Betrieb anderer Instrumente umsetzt. Der zwischen einem Höchstwert von 20 und 1 einregelbare Verstärkungsfaktor hat Drifteigenschaften von weniger als 3 μV/° C. Für Spezialinstrumente ist der Chopper K-PAC/4A mit Feldeffekttransistoren für Einsatz mit Eingangspegeln zwischen 0 und 1 mV und Drifteigenschaften innerhalb 0,5 μV/° C lieferbar.

Das Gerät wird mit Netzanschluss betrieben und hat einen internen Trenntransformator, der die Schaltung unter allen Umständen vom Netz isoliert. Das Gerät ist mit Höchst- oder Mindestspannung für offenen Eingangskreis lieferbar, so dass angeschlossene Instrumente für eigensicheren Betrieb ausgelegt werden können.

EE 95 762 *für weitere Einzelheiten*

Sampling-Oszillograf

Vertrieb: Claude Lyons Ltd, Valley Works, Hoddesdon, Hertfordshire

(Abbildung Seite 477)

Von dem Geschäftsbereich Instrumente der Claude Lyons Ltd kann nunmehr der C.R.C.-Sampling-Oszillograf Typ OCE 580 bezogen werden. Es handelt sich um ein Einkanalgerät mit

0,35 ns Anstiegszeit und einem Höchstablenkfaktor von 10 mV/cm. Ein wertvolles Merkmal ist die sehr breite Auswahl von Zeitmassstäben, die ohne Dehnung zur Verfügung steht. Die achtzehn geeichten Durchlaufgeschwindigkeiten liegen zwischen 0,2 ns/cm und 100 μs/cm. Die hohe Geschwindigkeit von 0,2 ns/cm ohne Dehnung ermöglicht Aufrechterhaltung der Spurhelligkeit und vermeidet die Minderung der Zeichenschärfe, die bei anderen Oszillografen mit hoher Schriebgeschwindigkeit auftritt, da zur Aufrechterhaltung ausreichender Helligkeit die Anzahl der je cm entnommenen Proben reduziert werden muss.

EE 95 763 *für weitere Einzelheiten*

UKW-UHF-Empfänger

Eddystone Radio Ltd, Eddystone Works, Alvechurch Road, Birmingham, 31

(Abbildung Seite 477)

Der Eddystone 990S ist ein volltransistorierter Einzelumwandlungsempfänger für AM- und FM-Signale im Bereich 230 ...870 MHz, der damit alle für Fernsehübertragungen zugeteilten Bänder IV und V überstreicht. Zwei getrennte Vorverstärkerköpfe mit Durchgangsleitungsschaltungen finden Anwendung; der Bereichschalter steuert einfach je nach gewünschtem Bereich den einen oder anderen an.

Anwendung einer hohen ZF (36,5 MHz) gewährleistet gute Spiegelselektion, und die vorgesehenen alternativen Bandbreiten gestatten Empfang von Signalen mit aufgeprägter Amplituden-, Video- oder Frequenzmodulation. Ein neiderohmiger Ausgang bei Zwischenfrequenz erlaubt Treiben von Hilfseinrichtungen.

Der Hörfrequenzgang ist ausgezeichnet, was für die Überwachung von Fernsehtonkanälen hoher Qualität wünschenswert ist. Zur Erleichterung ist ein Monitorlautsprecher eingebaut; externe Lautsprecher, Kopfhörer und Fernleitungen können angeschlossen werden. Im letzteren Fall wird die Ausgangsleistung für direkten Anschluss an Leitungen mittels eines getrennten Reglers begrenzt.

Der Empfänger 990S arbeitet im Prinzip mit einer 12-V-Stromversorgung, die durch einen Wechselstromnetzanschluss gespeist wird. Den Empfänger kann man direkt aus einer geeigneten Gleichstromversorgung (Trockenbatterien oder Akku) speisen, was für etwaigen Einsatz im beweglichen oder Aussendienst vorteilhaft ist.

Weitere bemerkenswerte Merkmale sind der Eddystone-Standard-Präzisionsfeinstelltrieb mit Untersetzung, klare Abstimmskalen, Quarzeicher, Abstimmanzeiger, unabhängige, direkt in dB geeichte Verstärkungsregelung für HF und ZF, kompakte Abmessungen und niedriges Gewicht.

Das Gerät wird in Tischausführung geliefert, lässt sich aber leicht für Gestelleinbau umwandeln.

EE 95 764 *für weitere Einzelheiten*

Überschall-Lecksucher

Dawe Instruments Ltd, Western Avenue, Acton, London, W.3

(Abbildung Seite 478)

Aus einem beliebigen Druckbehälter leckende Gase oder Flüssigkeiten erzeugen Überschallenergie. Die erzeugte Menge hängt von der Geschwindigkeit des entweichenden Gases ab, die ihrerseits durch den Druck und die Abmessungen der Öffnung bestimmt wird. In allen Fällen ergibt sich ein breites Frequenzspektrum im 40-kHz-Gebiet, was selbst in Gegenwart wesentlichen hörbaren Rauschens von anderen Quellen zum Aufspüren eines Lecks ausgenutzt werden kann.

Das Prinzip findet im neuen Dawe-Überschall-Lecksucher 1873 Anwendung, der nicht grösser als eine übliche Taschenlampe ist und in der Hand gehalten wird. Der Sucher besteht aus einem sehr richtungsempfindlichen mikrofonartigen Wandler, der einen auf 40 kHz ±4 kHz abgestimmten Transistorverstärker speist. Der Ausgang des Verstärkers wird visuell auf einem Drehspulinstrument oder hörbar mittels abnehmbarer Kopfhörer angezeigt. Das ganze Gerät mit Ausnahme der Kopfhörer ist in ein robustes Aluminiumgehäuse von 51 mm Durchmesser und 343 mm Länge eingebaut und wiegt komplett mit Batterien 900 g. Eine Batterie reicht für 150 h Dauerbetrieb; Batterieprüfung ist vorgesehen.

In Betrieb wird das Suchgerät einfach so ausgerichtet, dass der grösste Ausschlag oder das lautstärkste Signal erzielt wird. Das Leck ist dann in der sichtbar gemachten Richtung, die durch Nachfolgen genauer eingeengt werden kann.

Die Empfindlichkeit lässt sich mittels eines Reglers einstellen, der den Ausgangspegel hebt oder senkt. Es wird dadurch möglich, in unmittelbarer Nähe eines Hochgeschwindigkeitlecks oder in etwa 14 m Abstand von einer Öffnung mit 0,25 mm Durchmesser und bei weniger als 0,7 atü Druck zu suchen. Bei ausreichendem Druck kann die Öffnung einen Durchmesser von nur 0,05 mm haben.

Das Gerät ist in erster Linie als Lecksucher entwickelt, kann jedoch auch jede beliebige Quelle eines 40-kHz-Signals finden, z.B. ausgelaufene Lager, Wirbelströmungen (hydraulisch oder pneumatisch), Reibung, Koronaentladungen usw. In diesen Fällen könnten Spezialsonden erforderlich sein.

EE 95 765 *für weitere Einzelheiten*

Mikrominiaturbatterie

Mallory Batteries Ltd, Gatwick Road, Crawley, Sussex

Mallory Batteries Ltd kann jetzt eine neue Quecksilberbatterie liefern, die bei nur 5,6 mm Durchmesser und 3,3 mm Höhe 0,3 g wiegt.

Die Batterie, Typ RM-212, hat 1,35 V Nennspannung und alle Eigenschaften, die für die Batterien dieses Typs bekannt sind: Langlebensdauer, Spannungskonstanz und gleichbleibende Entladung.

Die Zelle wurde speziell für Hörapparate neuer Konstruktion, die völlig in den Hörkanal passen, entwickelt, kann jedoch in gleicher Weise in "Radiopillen" und anderen diagnostischen Sonden Anwendung finden.

EE 95 766 *für weitere Einzelheiten*

Bistabiles Relais

P.A.R. Ltd, Talbot Works, Talbot Street, Nottingham

(Abbildung Seite 478)

Das neue bistabile Relais hat vier Umschaltkontakte, die mechanisch mit Hilfe von zwei dynamisch abgeglichenen Ankern arretiert werden. Die Arretierung ist verriegelt und kann elektrisch rückgestellt werden.

Robuste Konstruktion gewährleistet, dass das Relais den schwersten Stossbedingungen in beliebiger Ebene ausgesetzt werden kann, ohne dass die Arretierung dadurch gesperrt oder freigegeben wird.

Diese Charakteristik ist vor allem für Hersteller von Maschinen mit Münzeinwurf, Einbrecheralarmen, Signal-Kontrolluhren, Zähleinrichtungen und sich bewegenden oder vibrierenden Maschinen wertvoll.

Andere typische Anwendungsmöglichkeiten ergeben sich in der industriellen Steuerung, Speichereinrichtungen oder Fallen, in denen Momentansignale dauernd, aber mit Rückstellmöglichkeit angezeigt werden sollen.

Das Relais ist kompakt mit Abmessungen von nur 25,4 × 70 × 51 mm hoch, einschliesslich transparenter Haube, die zur Standardausrüstung gehört. Das Grundraster der Kontakt- und Spulenfahnen ist 2,54 mm für Druckschaltungen.

EE 95 767 *für weitere Einzelheiten*

Programmschalter

Sealectro Ltd, Walton Road, Farlington, Portsmouth, Hampshire

(Abbildung Seite 478)

Sealectro kann nunmehr einen neuen Actan - Doppeltrommel - Programmschalter mit äusserst langfristiger Programmierfähigkeit liefern. Der neue Schalter hat eine Treibertrommel mit Umdrehungsgeschwindigkeiten von 5 UPM bis zu einer Umdrehung in 28 Tagen. Für jede 360°-Drehung der Treibertrommel wird die Sekundärtrommel einen Schritt vorwärts bewegt, wodurch sich die sehr langfristige Steuerfunktion ergibt. Das abgebildete Gerät hat 36 Programm- und 16 unabhängige Kontaktpositionen je Trommel. Andere lieferbare Einheiten haben bis zu 120 Programm- und 57 unabhängige Kontaktpositionen je Trommel. Der neue Programmschalter kann auch durch eine ferngesteuerte Magnetspule getrieben werden und dann Schrittschalt- und Zählfunktionen ausüben. Schiebekontaktaktivatoren gestatten einfache und schnelle Programmänderung ohne Spezialwerkzeuge an Ort und Stelle.

Für Sonderanwendungszwecke stehen 10-A-Kontakte zur Verfügung.

Ausser dieser neuen Ausrüstung für langfristiges Programmieren bietet Sealectro noch ein komplettes Sortiment von Programmschaltern an, zu deren besonderen Merkmalen ausbaubare Trommeln, gestapelte Kontakte, einstellbare Kontakte, Mehrfachkontaktpositionen und Festprogrammgeräte gehören.

EE 95 768 *für weitere Einzelheiten*

Feldauswerter

Servomex Controls Ltd, Crowborough, Sussex

(Abbildung Seite 478)

Servomex Controls Ltd kündigt einen verbesserten Feldauswerter für das Teledeltos-Widerstandsanalogverfahren zur Lösung von Feldproblemen der Elektrostatik, des Magnetismus, der Mechanik, Wärmeübertragung und Hydraulik an.

Der Auswerter kommt in zwei Modellen, dem mit einer 9-V-Batterie betriebenen FP 130 und dem FP 144 mit Netzanschluss. Beide Geräte messen das Potential des Feldes nach dem Potentiometerverfahren, wobei die Quellenspannung auch für den Schleifdraht benutzt wird, so dass die Messung von Netzspannungen unabhängig ist.

In diesem Verfahren findet Teledeltos-Papier Anwendung, das verhältnismässig billig ist, einen geeigneten spezifischen Widerstand hat und innerhalb von 30 Minuten ein Analog zeichnen und mit der Ausmessung beginnen kann. Eine verbesserte Spezial-Silberfarbe wird zum Aufmalen der Linien gleichen Potentials verwendet. Die Elektroden werden mit einer dem Gerät entnommenen Spannung beaufschlagt und das Potential im Feld mittels des Potentiometers aufgetragen.

Das batteriebetriebene Modell FP 130 ist für Studenten nützlich; es ist mit einem eingängigen Potentiometer guter Qualität ausgestattet und hat ±1 Prozent Genauigkeit; elektrisches "Bandspreizen" gestattet jedoch bessere Unterscheidung zwischen 0 und 10 Prozent. Stecker und Buchsen für Anschluss jeder beliebigen Speisespannung bis zu 100 V— sind vorhanden.

Das netzbetriebene Modell FP 144 hat eine niedrige Spannung von 5...10 V für gewöhnliche Laplacefelder und eine hohe Spannung von 50...100 V zur Speisung eines konstanten Stroms durch hochohmige Widerstände in Punkte des Feldes für Poissonsche Felder, so dass der Stromfluss vom Rest des Diagramms praktisch unabhängig ist. Das Modell ist mit einem zehngängigen Potentiometer mit direkter, auf 0,1 Prozent ablesbarer Skalenanzeige ausgestattet.

Servomex-Feldauswerter sind preiswert, zeitsparend und kurzschlussfest. Sie sind kompakt und leicht tragbar: das batteriebetriebene Modell wiegt nur 675 g, das netzbetriebene 2,25 kg.

EE 95 769 *für weitere Einzelheiten*

Entlötkolben

Vertrieb: Soldering Instrument Components Ltd, Adcola House, Gauden Road, London, S.W.4

(Abbildung Seite 479)

Adcola kündigt einen neuen Lötkolben M207 an, dessen Hauptanwendungszweck ist, Lot von Lötfahnen, Bauelementen und gedruckten Schaltungen zu entfernen. Das Instrument wird von Adcola Products Pty Ltd in Melbourne hergestellt und durch Soldering Instrument Components Ltd, eine Tochter der Adcola Products Ltd, vertrieben.

Das Gerät ist mit einer hohlen, mit rostfreiem Stahl ausgekleideten Lötspitze ausgerüstet, die für normales Löten oder Entlöten zu verwenden ist; es können aber auch Vollspitzen eingesetzt werden. Der aus dem Handgriff vorragende Ball kann mit dem Zeigefinger betätigt werden und zieht Lot durch die Spitze; ein am anderen Ende der Spitze liegendes Gazesieb sammelt das Lot, das dann durch Drücken des Balls ausgeblasen werden kann. Das Gazesieb lässt sich zum Reinigen aus der Siebkammer entfernen.

Für den M207 gibt es 12 Lötspitzen, und zwar sechs hohle für Entlöten und sechs massive. Das Sortiment umfasst Spitzen mit 1,6 mm, 2,4 mm, 3,2 mm, 4 mm, 4,8 mm und 6,35 mm Lötflächendurchmesser; die hohle 3,2-mm-Lötspitze gehört zur Standardausrüstung. Die Lötspitzen sind eisenplattiert, um Festbrennen im Kolben zu verhindern und die Lebensdauer der Lötfläche zu verlängern.

Heizelemente, Entlöt- und Lötspitzen, Siebe und Siebkammern sind bei Soldering Instrument Components Ltd auf Lager, und die Firma übernimmt auch den Service in Grossbritannien. Geräte sind für jede Spannung zwischen 6 und 250 V lieferbar.

EE 95 770 *für weitere Einzelheiten*

Mikrowellen-Rauschanalysator

James Scott (Electronic Engineering) Ltd, Carntyre Industrial Estate, Glasgow, E.2

James Scott (Electronic Engineering) Ltd kündigt eine neue Serie verbesserter Hochleistungs - Rauschmessausrüstungen Typ 123 an, die noch feinere Messungen gestatten als ihre Vorgänger.

Die Empfindlichkeit für FM- und AM-Rauschen in 150 kHz Abstand vom Mikrowellenträgersignal war—als Signal-Rauschen-Leistungsverhältnis für die gesamte Seitenbandleistung ausgedrückt und in einem 100 Hz breiten Band gemessen—wie folgt:

Modulationsfrequenz	1 kHz	10 kHz	100 kHz	150 kHz
AM	125 dB	134 dB	135 dB	135 dB

Die erzielte FM-Empfindlichkeit war 0,0025 Hz Effektivhub.

Mit der neuen Ausrüstung sind folgende Werte erreichbar:

Modulationsfrequenz	1 kHz	10 kHz	100 kHz	150 kHz
AM	125 dB	143 dB	163 dB	168 dB

Die erzielbare FM-Empfindlichkeit ist 0,0008 Hz Effektivhub.

Diese Werte wurden durch eine neue Bauart erreicht, die gestattet, bei der Abtastung der zu messenden Quelle Proben höherer Leistung zu entnehmen. Ausserdem wurde der Einfluss von akustischen Geräuschen und Vibrationen gemindert. Komplexe Rückkopplungsschleifen wurden zur Erreichung folgender Resultate vorgesehen:

(a) Aufrechterhaltung des Brückenhohlraumabgleichs. Ein neuartiges, patentiertes System mit Thermoheber-Servoelement findet Verwendung.

(b) Ermöglichung von Rauschmessungen durch Aufhebungstechnik an bestimmten frequenzmodulierten Trägern.

(c) Beseitigung des störenden, von der Stromversorgung der zu messenden Quelle herrührenden Modulationseffektes.

(d) Frequenzmitnahme des Hilfsoszillators.

Verbesserungen der Analysenausrüstung gestatten Untersuchung des Rauschspektrums zwischen 50 Hz und 1 kHz vom Mikrowellenträger in 30 Hz Bandbreite.

Die Ausrüstung ist in Ausführungen für das X- und J-Band und in Kürze auch für das C- und S-Band lieferbar.

EE 95 771 *für weitere Einzelheiten*

Hochleistungsverstärker

S.E. Laboratories (Engineering) Ltd, North Feltham Trading Estate, Feltham, Middlesex

(Abbildung Seite 479)

Ein rauscharmer Verstärker SE.436 mit hoher Verstärkung und linearem Frequenzgang von 5 Hz...10 kHz wird von S.E. Laboratories (Engineering) angekündigt. Er ist hauptsächlich für Einsatz mit amplitudenempfindlichen mechanisch-elektrischen Wandlern wie z.B. gleichstromerregten Dehnungsmessstreifen, selbsterregten Geschwindigkeitsaufnehmern usw. bestimmt.

Eine Variante, der linear-integrierende Verstärker SE.436/I, enthält zwei zusätzliche integrierende Netzwerke mit unterschiedlichen Zeitkonstanten.

Einschalten des einen oder anderen gestattet Integrieren des Ausgangssignals des selbsterregten Geschwindigkeitsaufnehmers für nachfolgende Anzeige oder Registrierung der Schwingungsamplitude.

Die Verstärkerausgänge sind für das Treiben von hochfrequenten Registriergalvanometern, Magnetbandgeräten, Oszillografen usw. geeignet. Mehrfache Rückkopplungsschleifen gewährleisten hohe Verstärkungskonstanz und Gleichtaktunterdrückung.

Getrennte Spannungs- und Stromausgänge sind vorhanden. Der Verstärker ist dank der niedrigen Quellenimpedanz des Spannungsausgangs für spannungsempfindliche Registrier- und Anzeigegeräte aller Art geeignet.

EE 95 772 *für weitere Einzelheiten*

UHF-Wobbulator

Compagnie Générale de Métrologie, B.P.30, Annecy, Frankreich

(Abbildung Seite 479)

Zusammen mit einem Oszillografen und einem Markengeber bildet der UHF-Wobbulator 241A eine Kurvenschreiberausrüstung, die für Bestimmung des Amplitudenfrequenzgangs aktiver und passiver Schaltung mit Betriebsfrequenzen zwischen 470 und 870 MHz eingesetzt werden kann.

Zu den Anwendungsgebieten gehören das Abgleichen von UHF-Empfängern, Bestimmung der Gesamtkurve von Fernsehempfängern (Bänder IV und V) von der Antenne bis zum Bildschirm, Messen des Stehwellenverhältnisses von Antennen, Koaxialkabeln und anderen Schaltungen in Verbindung mit Anpassungsbrücke HCX35A und NF-Verstärker HCX52A.

Für diesen Zweck gibt der Wobbulator mittels eines Lecherleitungsoszillators eine frequenzmodulierte UHF-Spannung ab, die laufend auf einer in MHz geeichten Skala angezeigt wird.

Die Frequenz wird mittels eines Schwingkondensators moduliert, der auf einen rechteckigen bronzefederartigen Träger montiert ist. Diese Anordnung hat den Vorteil, eine lineare Verschiebung zu ergeben und gleichzeitig eine robuste Konstruktion zu gewährleisten.

Zur Verhinderung parasitärer Amplitudenmodulation und Erhaltung eines konstanten Ausgangspegels unabhängig von der angezeigten Frequenz ist ein Regelkreis für die Oszillatorspannung vorhanden.

Der Wobbulator 241A hat einen weiteren UHF-Ausgang, der Markierung der dargestellten Kurve gestattet.

Das Gerät kann in verschiedenen Ausführungen den Anforderungen an Ausgangsimpedanz und allgemeine Formgebung entsprechend geliefert werden.

EE 95 773 *für weitere Einzelheiten*

Zusammenfassung der wichtigsten Beiträge

Bürstenloser Gleichstrom-Schrittmotor mit Binär-Dezimalumwandlung von W. Leonchick

Zusammenfassung des Beitrages auf Seite 426-429

Ein beschriebener Rechnersprachenumsetzer macht von einem einzigartigen vektoriellen Prinzip Gebrauch, um von binären Daten eine unmittelbare Dezimalanzeige zu erhalten. Die Entwicklung wird absichtlich eingehend behandelt, um volles Verständnis zu gewährleisten.

Da Schrittmotore der kontaktlosen Bauart im wesentlichen digitale Eingangs-Ausgangseinrichtungen sind, befasst sich der zweite Teil des Beitrags mit ihrer Anwendung in dem oben erwähnten Umsetzer als Gleichstrom-Servoschaltwerk in einer offenen Schleife. Vorteile sind: (a) Zuverlässigkeit, (b) wartungsfreier Betrieb und (c) Raum- und Gewichtseinsparungen im Vergleich mit Wechselstrom-Servoschaltungen. Das beschriebene System wird durch veränderliche Umgebungsbedingungen nicht beeinflusst.

Digital-Analogspannungs-Umwandlung sowie Binär-Dezimalumsetzerausrüstungen werden durch die Vielseitigkeit des elektromechanischen Bausteins überflüssig.

Eine MOST-Torschaltung für das Einblenden vor dem Leseverstärker von G. N. Hunter

Zusammenfassung des Beitrages auf Seite 432-435

Die Eignung des MOST für das Einblenden schwacher Signale wird beschrieben. Beim Einblenden des Eingangs eines Leseverstärkers in einem Computerspeicher entsteht die Aufgabe, grosse Rauschimpulse eine kurze Zeit vor Beginn der Übertragung des Kleinsignals zu unterdrücken. Eine praktische Torschaltung wird untersucht; ihre Leistung mit derzeitigen MOST-Typen zeigt, dass sie mit Speicherzykluszeiten bis zu 1 μs herunter vereinbar ist.

Weiterentwicklung des Entwurfs minimaler NOR-(NAND-) Kombinationsschaltungen für N-Variable
von G. W. Copperwhite und D. Zissos

Zusammenfassung des Beitrages auf Seite 436-442

Eine rein-algebraische Schritt-für-Schrittmethode für den Entwurf minimaler NOR- (und NAND-) Kombinationsschaltungen wird beschrieben. Im wesentlichen wird nach dieser Methode der algebraische Ausdruck der zu verwirklichenden Funktion in drei Schritten optimiert. Eine theoretische Rechtfertigung wird für die Entwurfsschritte gegeben. Zur Illustration der Methode werden verschiedene Beispiele, einschliesslich eines vollen Addier-Subtrahierwerkes, herangezogen.

Ein Gleichtaktverstärker und seine Anwendung in Spannungskonstanthaltern von G. N. Patchett

Zusammenfassung des Beitrages auf Seite 443-445

Die beschriebene Einrichtung wurde zur Verringerung des Phasenwinkels zwischen Ein- und Ausgangsspannung eines Präzisions-Wechselspannungskonstanthalters entwickelt. In der Ausrüstung wird die Phase der Ausgangsspannung mit der einer Bezugsspannung verglichen und bei Auftreten eines Phasenfehlers eine die Blindkomponente aufhebende, um 90° verschobene Spannung in den Konstanthalter gespeist. Mit dieser Einrichtung wird der Phasenwinkel unter allen Belastungs- und Leistungsfaktorbedingungen auf 15′ reduziert.

Impedanzkorrektur des Vertikalspeisesystems einer nullausgefüllten Antenne mit hohem Gewinn von A. B. Shone

Zusammenfassung des Beitrages auf Seite 446-448

Nach einem in Grossbritannien geltenden Grundsatz werden Sendungen im UHF-Band von einem gemeinsamen Aufstellungsort ausgestrahlt, und es hat sich in vielen Fällen als wirtschaftlich erwiesen, die ganze zur Verfügung stehende Antennenöffnung für eine Antenne für bis zu vier UHF-Kanäle auszunutzen. Diese Anordnung wirft die Frage auf, wie man die erforderliche Impedanzanpassung über die grosse Bandbreite erzielen kann und wie man—vor allem wenn man an Farbfernsehen denkt—ein gleichförmiges Vertikaldiagramm erreicht, das so weit wie möglich einer Näherung des Kosecans-Gesetzes folgt. Der Beitrag macht Vorschläge zur Erreichung beider Ergebnisse, die dann anwendbar sind, wenn die dieser Lösung eigene Diagrammneigung zweckdienlich ist.

Verstärker mit unendlich hohem Eingangswiderstand von J. B. Matley

Zusammenfassung des Beitrages auf Seite 449-451

Dieser Beitrag beschreibt und analysiert einen zweistufigen Verstärker mit Spannungsverstärkung 1:1. Ein Feldeffekttransistor in der ersten Stufe kann Eingangswiderstände von wesentlich mehr als 100MΩ ergeben. Eine besprochene Abwandlung dieser Schaltung gestattet Einstellen des Eingangswiderstandes auf unendlich hoch; eine Analyse der abgewandelten Schaltung wird gegeben.

Eine umpolbare 10-kV-Konstantstromversorgung von M. C. Jackson

Zusammenfassung des Beitrages auf Seite 452-453

Diese ungefährliche ±10-kV-Stromversorgung hat Festkörperschaltungen und Überlastungsschutz und entspricht konstruktionsmässig der Baureihe Harwell 2000. Ursprünglich wurde das Gerät zur Erzeugung der Beschleunigungs- und Fokussierspannungen für die Ionenquelle eines Massenspektrometers entwickelt. Die Konstanz der Ausgangsspannung ist besser als 4×10^4 pro Stunde.

Analoge Logikschaltungen, die Relais schalten von C. D. Lewis

Zusammenfassung des Beitrages auf Seite 454-457

Eine Reihe analoger Computerschaltungen mit ein oder zwei Funktionsverstärkern werden beschrieben, die die binären Funktionen NICHT, UND, ODER, AUSSCHLIESSLICHES ODER, WENN . . . DANN, ÄQUIVALENT, NOR, NAND und ABER . . . NICHT simulieren.

Integrierte Elektronik—ihr Platz in der Schaltung von K. J. Dean

Zusammenfassung des Beitrages auf Seite 458-460

In diesem Beitrag werden einige neue Entwicklungen in der Elektronik zusammen mit ihrem Einfluss auf die Schaltung besprochen. Laboratoriumseinrichtungen für das Studium der Anwendung integrierter Schaltungen werden beschrieben.

Gleichrichtung bei Verwendung von Elementen mit symmetrischen Kennlinien von J. F. Young

Zusammenfassung des Beitrages auf Seite 461-463

Es wird gezeigt, dass ein Element mit symmetrischer Spannungs-Stromkennlinie als Gleichrichter von impulsförmigen oder durch geradzahlige Harmonische verzerrten Wellenformen wirken kann. Die Eigenschaften eines symmetrischen Gleichrichters für Impulse werden abgeleitet, und es wird gezeigt, dass der gleichgerichtete Ausgang eher vom Tastverhältnis als von der Eingangsimpulsamplitude abhängt. Als symmetrische Gleichrichter wurden Zenerdiodenpaare angewendet, jedoch haben auch andere Elemente geeignete Kennlinien. Einige Anwendungsmöglichkeiten für symmetrische Gleichrichter werden besprochen.

Ein einfaches, automatisches Zeitsteuergerät für fotografische Vergrösserungen von J. L. Linsley Hood

Zusammenfassung des Beitrages auf Seite 464-466

In der beschriebenen Schaltung findet ein Photowiderstand zum Messen der Helligkeit eines projizierten Bildes und—mit Hilfe einer integrierenden Schaltung—zur Regelung der Belichtungszeit Anwendung. Belichtungszeiten können zwischen 1 und 40 Sekunden liegen.

Ein Serien-Stromregler von T. M. Palmer

Zusammenfassung des Beitrages auf Seite 467-469

Der Beitrag beschreibt eine einfache Alternative zu einem kompletten Konstantstrom-Netzteil in Form eines in sich geschlossenen Elementes, das unabhängig von der angelegten Spannung einen konstanten Strom durchlässt.