ELECTRONIC ENGINEERING

VOL. 31 No. 381

NOVEMBER 1959

Commentary

THE proposal included in the manifesto of the two major political parties to appoint a Minister for Science was one of the least controversial issues in the recent General Election in this country, but the fact that both parties had put it forward is perhaps an acknowledgment that such a measure was long overdue.

The new Government has obviously felt some sense of urgency about this matter for one of its first acts has been to issue an Order in Council creating Lord Hailsham as the first holder of this very important office and transferring to him the functions previously exercised by the Prime Minister under the Atomic Energy Acts. The same Order substitutes the Minister for Science, who, in addition will retain the office of the Lord Privy Seal, "for the Lord President of the Council as member and chairman of the Privy Council committees for the five research councils, and transfers to him the functions of the Lord President in relation to those councils and certain other organizations".

The terms of reference are in themselves explicit enough but in amplification of them it has been stated that Lord Hailsham, in combining the duties of Lord Privy Seal and Minister for Science, "will be responsible to the Government for the Council for Scientific and Industrial Research, the Medical Research Council, the Agricultural Research Council, the Overseas Research Council and the Nature Conservancy. Besides exercising the ministerial functions under the Atomic Energy Acts he will exercise general supervision over the programme of space research at present being devised under the direction of the steering group presided over by Sir Edward Bullard".

He will be responsible for the broad questions of scientific policy and will receive assistance from the Advisory Council, but he will not of course be answerable for the various Government scientific establishments, the responsibility for which will remain with the ministers concerned.

In the short time that he has been appointed Minister for Science, Lord Hailsham has been able to do no more than survey the situation which the new office presents, but his assessment given at a recent Press Conference appears encouraging.

He is, on his own admission, no scientist but this in no way detracts from his suitability. On the contrary, a non-scientist, providing he has sufficient appreciation of the broad problems and can give to each its proper weight and importance, is, in our view, an ideal candidate.

He has, in addition, cleared the air to a considerable extent in stating what he is not expected to be. He will not be a Master of the Queen's Rockets or a super Minister of Power, Transport, Agriculture or Health, neither will he be a Chancellor of University Chancellors.

This negative attitude may give the impression that the appointment is no more than a piece of political window dressing designed to fulfil a promise made in the heat of an election and that everything will go on in the same way as it did before, but this is far from being the true state of affairs.

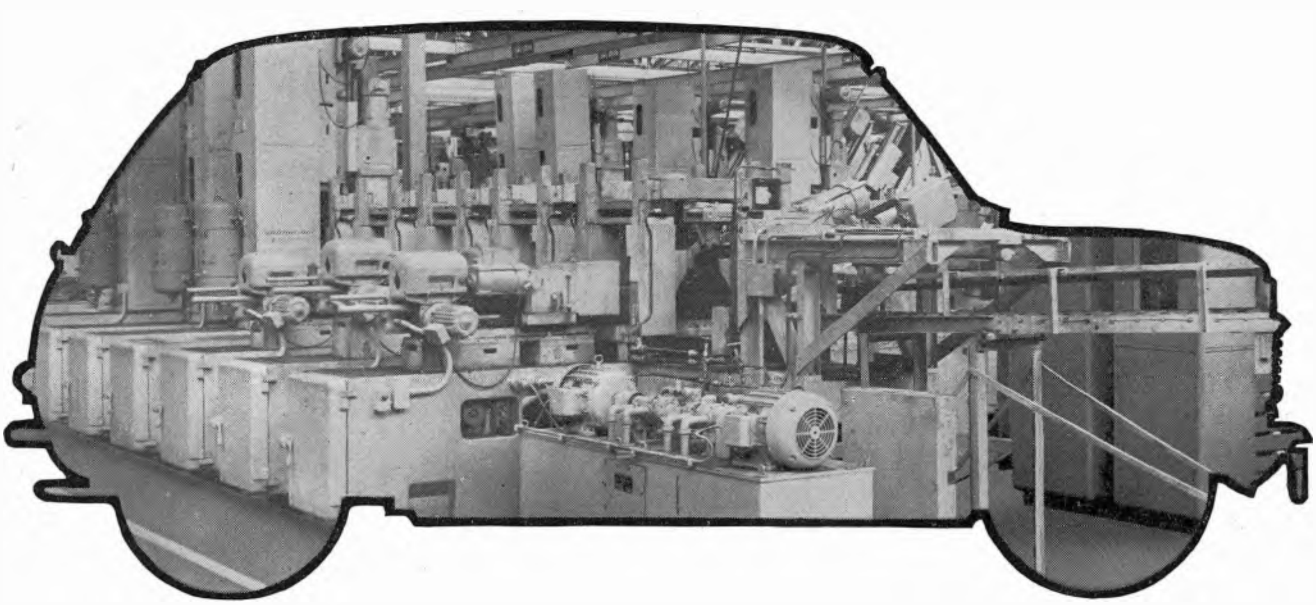
In this age when scientific developments make a greater impact on our lives than ever before, the need for this new office is real and the new Minister has made his purpose clear—"to make the voice of science coherent and articulate under Government encouragement, and in one real sense to make science self-governing under Government inspiration".

It has been made clear that while he must remain interested in the more glamorous and newsworthy fields such as space research, he has every intention of striking a proper balance—a most difficult task—between research which is carried out simply and solely for national prestige, and pure fundamental research of the kind in which this country has always excelled.

It is tempting to be the first country to put a man into outer space, but to enter this race would be financially disastrous to us. The mere projecting a man into outer space, keeping him alive there and bringing him safely back to earth would require a fantastic amount of effort and it is doubtful whether any worthwhile scientific results would be achieved. Moreover, it is acknowledged that most, if not all, the data required in outer space can be obtained by the use of instruments alone and it appears that this is the direction in which the British effort will be concentrated.

All must wish the new Minister well in his determination to make British science in all its fields an instrument of peaceful progress, for the rewards are enormous. If he succeeds he will have contributed in no small measure to the peace and prosperity of the world, and raised the esteem of this country to the greatest of heights.

ACCESS *(Austin Cold-Cathode Electronic Switching System)*



A Static Switching System using Cold-Cathode Tubes

By R. W. Brierley*

The great strides taken recently in automating production processes has resulted in very complex control systems. The majority of these are of the ON-OFF *type using relays as the logical elements. Above a certain level of complexity, it becomes difficult to maintain a high standard of operating efficiency with such equipment. A search has begun for more reliable alternatives to the electromechanical relay and systems using static devices have started to emerge. One such system, using cold-cathode tubes, is described.*

(Voir page 706 pour le résumé en français: Zusammenfassung in deutscher Sprache Seite 711)

The rate of development of transfer machines, conveyor systems and other forms of so-called automation has been so rapid in the past few years that their control systems have now become exceedingly complex. This rate of development shows no sign of abating and, while the existing, well-known control methods employing relay circuits are capable of almost unlimited extension, there comes a stage, nevertheless, when it is extremely difficult in practice to maintain the required standard of reliability. Relays, even in the best of enclosures, all too frequently fall prone to the hazards of industrial usage and, if the future control systems are to have increased reliability, then it would seem that some substitute, either partial or complete, should be found for the conventional relay.

Awareness of this fact has resulted in the development of static switching systems in which the contacts and moving parts of the electromechanical relay have been replaced by the purely electrical switching action of such devices as magnetic amplifiers and transistors. Surprisingly, however, there is little indication that any real attempt has been made to exploit cold-cathode tubes for this purpose except in automatic telephony and, lately, in digital computers. Due almost entirely to pioneering development in the former field, a wide range of such tubes is available with inherent characteristics that are undoubtedly suited to the application. These tubes are compact, self-indicating and extremely reliable and, while it cannot be said that their life is infinite, experience has shown that it is compatible with that of production machinery. Moreover, being purely two-state devices they are, in a logical sense, ideal switching elements for systems which derive their inputs from binary devices such as limit switches and push-buttons.

Since the great majority of machine tools and material handling systems in existence today do, in fact, function with this type of input, a static switching system based upon cold-cathode tubes should be entirely practicable. Indeed, such a system is already in use controlling nine transfer machines, and in this article it is proposed to give some idea of the simple yet versatile circuits that have been used in place of electromechanical relays.

In spite of the merits of static switching, however, it is highly improbable that the electromechanical relay will ever be entirely displaced. The present cost of a static system for a large and complex installation is less than for one employing relays but the reverse would be true for simple schemes. This fact, coupled with many others, suggests that the two systems will tend to be complementary, each having its own sphere of application yet, in many instances, being used together to obtain the best balance of first cost, reliability and, most important of all, speed of maintenance.

The above photograph shows a typical transfer machine within the outline of an 'Austin Seven'

* *The Austin Motor Co. Ltd.*

Such is the hybrid system to be described, where it will be seen that relays and contactors have been used wherever it is necessary to switch large powers for solenoids and motors. These operations could well have been accomplished with magnetic amplifiers but both cost and bulk would have been a considerable embarassment. Similarly, push-buttons and limit switches have been retained though, in the case of the latter, it is chiefly the cost of the static equivalent that hinders its replacement. Simple inductive proximity switches have been used on transfer machines in place of some limit switches for over five years now and their initial cost, which is approximately four times that of a sealed limit switch, has been well repaid by what is virtually 100 per cent reliability.

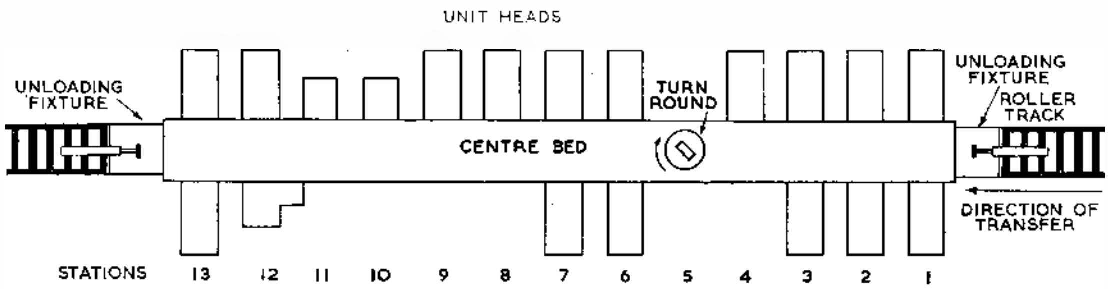


Fig. 1. Plan layout of transfer machine

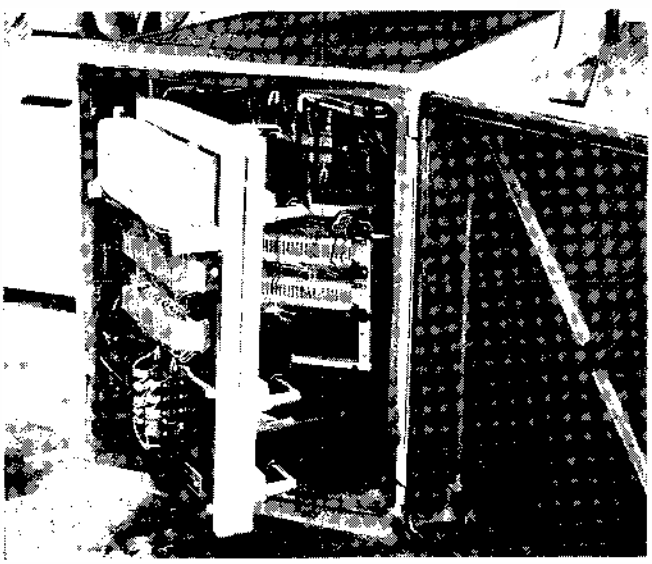


Fig. 2. Unit head control panel

Description of Transfer Machine

The type of transfer machine to which the ACCESS system has been applied is illustrated at the top of page 646. This machine forms part of a continuous machining line and is unattended in the sense that it automatically loads and unloads itself and controls the flow of components to the succeeding machine. It is some 30 yards long and has 20 machining stations in which cylinder blocks are drilled, faced, milled and bored. Certain stations, both within and at the end of the machine, re-orientate the block in 'turn-over' or 'turn-round' fixtures so that the appropriate surfaces are presented to the machining heads at the succeeding stations.

As shown in Fig. 1, a typical machine consists basically of a centre bed along which components are indexed or 'transferred', and a number of side beds on to which are fixed the unit head machine tools. By means of a standardized system of construction, these unit heads can be mounted horizontally, vertically, or at any intermediate angle. Every unit head has its own standard control panel, housed usually in a cubicle in the side bed, Fig. 2, and the circuits of these panels are so arranged that, once started by a signal from the Group Control, the unit heads automatically carry out their appropriate machining cycles with the various movements pre-set by adjustable limit switches. The cycle naturally varies according to the type of machining required; a drilling cycle, for example, consisting of the sequence:

(a) Fast approach of drills to component
(b) Slow drilling feed to correct depth
(c) Fast return to starting position

Similarly a tapping cycle would be:

(a) Fast approach of taps to component
(b) Slow tapping feed to correct depth
(c) Slow reversal of taps till clear of component
(d) Fast return to starting position

Presuming that a component is already waiting in the LOAD station, the complete cycle of operation of one of these machines is as follows:

(1)	TRANSFER	—	all components indexed forward one station.
(2)	CLAMP	—	all components at machining stations clamped rigidly in accurately-located positions.
(3a)	START HEADS	—	all unit heads signalled to start machining.
(b)	LOAD and UNLOAD	—	a new component fed into the LOAD station and a machined component ejected from the UNLOAD station.
(4)	UNCLAMP	—	all components at machining stations unclamped in readiness for the next transfer.

The power for all the various movements just outlined is either electric, hydraulic or pneumatic. The majority of unit heads, together with a swarf-collecting conveyor and input and output roller tracks fall into the first category, the transfer, clamping, loading and unloading mechanisms are hydraulic and, at those stations which precede a tapping operation, there is usually a pneumatic unit head that feeds a hollow probe into the already-drilled hole. This is not only to check that it has been fully drilled but also to blow out swarf.

Clearly, all these various movements need to be position sensed and limit or proximity switches are provided at the

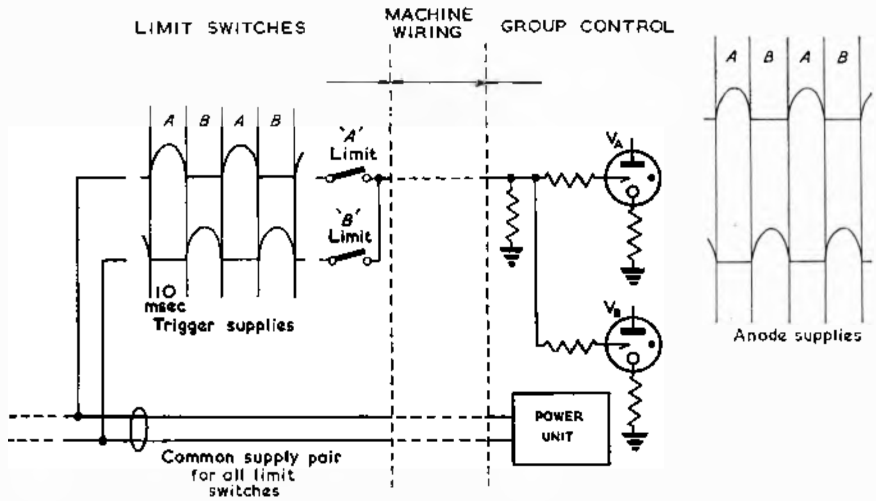
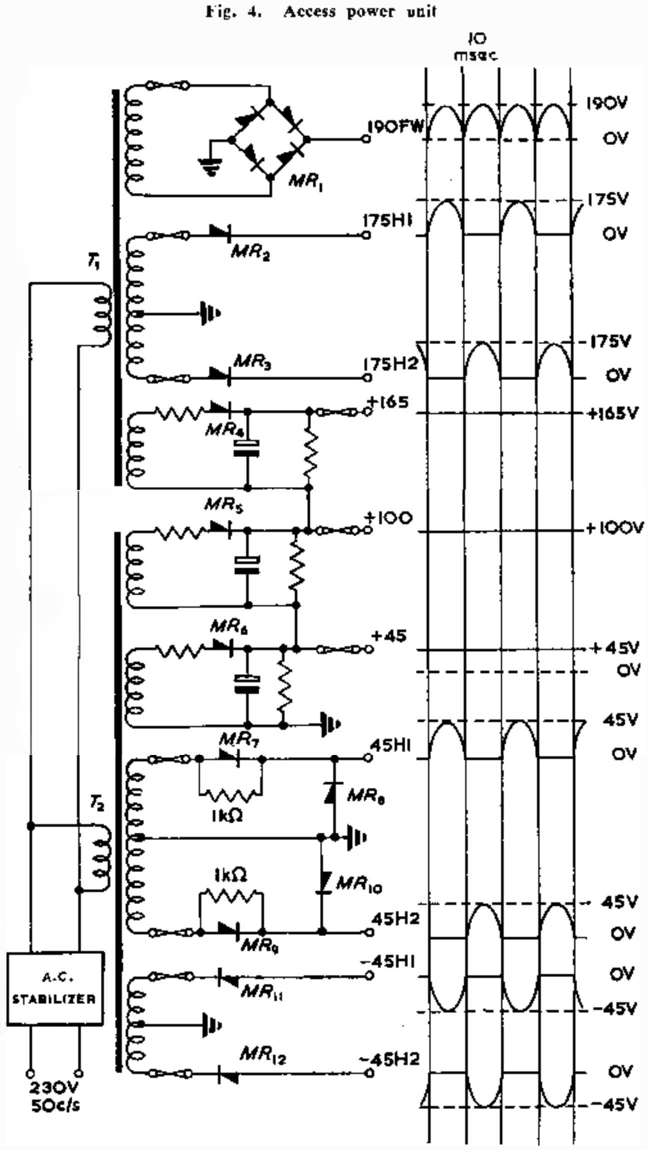


Fig. 3. Use of one conductor for two limit switches

Fig. 4. Access power unit



essential points for this purpose. Frequently, as in the case of the transfer hydraulic cylinder where it is necessary to know only if it is either fully extended or retracted, merely two limit switches are required for each particular movement. Nevertheless, the total number of such switches is considerable (over 200 on the machine illustrated) and it will be realized that, together with the many other circuits involved, both the wiring and interlocking of these limits can pose many problems.

Access—Basic Considerations

The electrical control system of a transfer machine can be divided roughly into three parts:

(a) The individual control panels for each unit head machine tool.

(b) The signal circuits receiving information from input devices such as limit switches and push-buttons.

(c) Sequence circuits for correlating this information and initiating the appropriate action.

The panels referred to at (a) are a standard type using BPO 3000 type relays and 50V d.c.-operated contactors. They have been used on many previous machines and their circuits which are quite conventional, do not warrant description. Sections (b) and (c), however, form a large part of what is known as the Group Control of the machine and it is to this unit that ACCESS has been applied.

In a conventional control system, the outputs from the various limit switches terminate in the Group Control where relays are energized in accord with the state of the switches. For a cold-cathode tube to function analogously to the relays, obviously it must conduct while a limit is closed and extinguish when a limit is opened. Further, if the switch is to control conduction via a trigger electrode, as in a triode or tetrode, then the anode must be fed from a periodically-interrupted supply. Since a high-speed of response is not usually essential in this type of machine control, a half-wave rectified anode supply at mains frequency would seem the obvious and simplest choice. Indeed, supplies of this form are used almost exclusively throughout the equipment, not only for the anodes of the tubes but also for the trigger signals from the limit switches. The reason for this is that, with a half-wave rectified anode supply, the tube is, in effect, phase-conscious and will conduct only to those trigger signals that occur in phase with the anode voltage. This leads naturally to the attractive situation represented schematically in Fig. 3 where, by using two anode and trigger supplies, each of opposite phase, the signals from a pair of limit switches can be transmitted to the Group Control over one conductor. Only the *A* limit will fire V_A, only the *B* limit V_B.

Since most of the other limit switches on the machine also work in logical pairs, e.g; TRANSFER BAR FORWARD—TRANSFER BAR BACK, TURNOVER UP—TURNOVER DOWN, they, too, can be arranged to signal over single conductors and thus halve a considerable proportion of the total wiring.

In the Group Control, many other circuit arrangements such as this occur, chiefly as a result of the use of half-wave supplies. It might not be illogical, therefore, to describe first the power unit that produces them.

Power Unit

The complete circuit shown in Fig. 4. A simple code has been adopted to designate the outputs and this, together with the relevant waveforms is shown on the circuit.

Wherever the suffix H1 occurs, it denotes a half-wave rectified supply of one phase (arbitrarily chosen) whereas the suffix H2 represents a similar waveform but of opposite phase. FW represents a full-wave rectified supply. The numerical value of any particular output is the peak voltage referred always with respect to earth while the negative sign which prefixes the two outputs -45H1 and -45H2 indicates that these two supplies are negative-going. D.C. outputs such as +165, +100 and +45 have no suffix. As will be apparent later, the 190FW, 175H1, 175H2 and +165 are anode supplies, the +100 and +45 are for biasing arrangements and the 45H1 and 45H2 are supplies for the limit switches. In conjunction with the latter, the two negative outputs −45H1 and −45H2 are used for operating diode indicator tubes.

All the rectifiers are of the silicon junction type, their excellent regulation, coupled with the stabilized a.c. input, giving constant output voltages in the simplest manner.

Possibly only one point in the circuit calls for comment and that concerns the shunt rectifiers of the 45H1 and 45H2 outputs. Without these two rectifiers MR_8 and MR_{10} and their associated 1kΩ resistors, the output impedance of the supplies is high during the half-cycles when MR_7 and MR_9 are cut-off and feedback of unwanted signals can occur. MR_8 and MR_{10}, however, conduct heavily to the negative half-cycles as passed by the 1kΩ resistors and serve effectively to clamp the output to earth while MR_7 and MR_9 are cut-off.

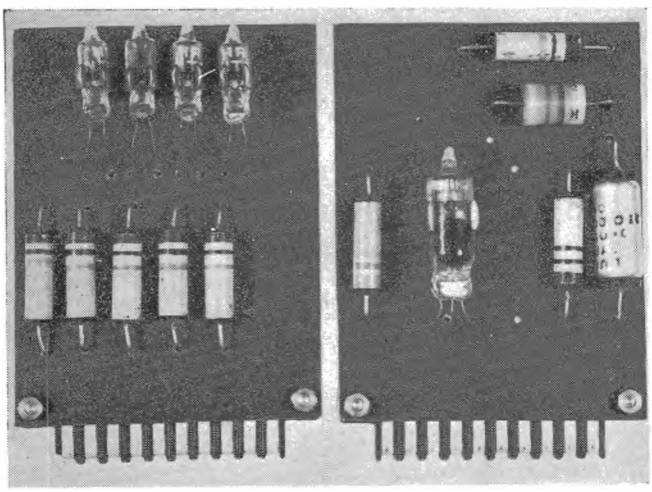


Fig. 5. Typical plug-in circuit elements

Circuit Description

All the circuits involved in this static system are built up from plug-in logical elements, two typical examples being illustrated in Fig. 5. There are in all twenty different elements and experience has shown that practically any control scheme of this type can be constructed with them. Undoubtedly, the number of different types could have been drastically reduced but the apparent simplification would have been far outweighed by the larger number of units required to achieve the same result. As it is, only ten basic printed boards are required, each board forming the basis of two or more logical elements.

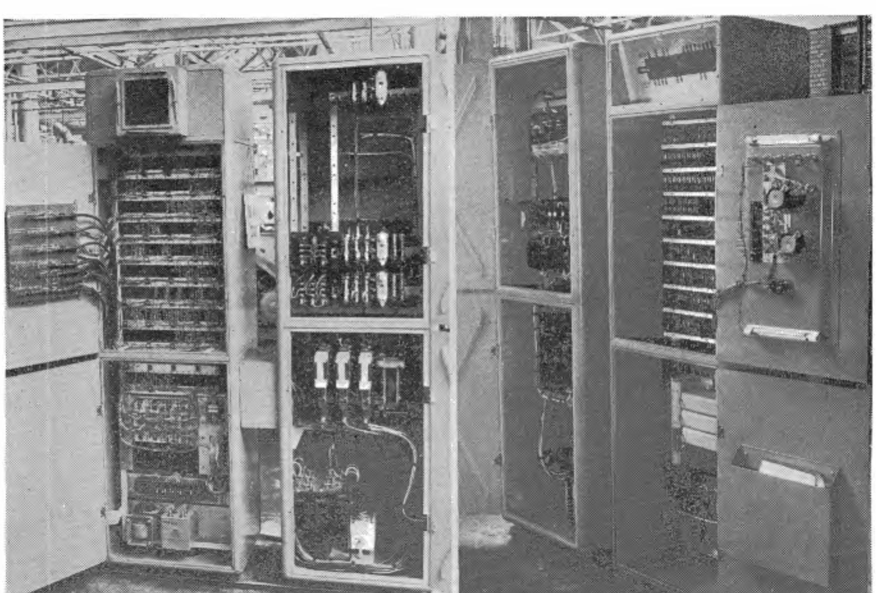
These elements will now be described and, in the diagrams that follow, their individual nature has been clearly shown since this not only aids description but also illustrates the ease with which a complete circuit is evolved. Some idea of the way in which the elements are mounted and wired can be gained from Fig. 6.

The cold-cathode tubes used in these units are Hivac types NT2, XC18 and XC23 and their main characteristics are given in Table 1.

TABLE 1
Main characteristics of cold-cathode Tubes NT2, XC18, XC23

	NT2	XC18	XC23
Minimum main gap breakdown voltage		210V	200V
Nominal main gap maintaining voltage		73V	67.5V
Nominal control gap breakdown voltage		68V	70V
Maximum cathode current		1mA	7.5mA
Nominal strike voltage	80V		
Nominal maintaining voltage	55V		
Maximum current	1mA		

Fig. 6. Internal construction of group control



ELEMENT R1

Fig. 7 is the circuit of the first standard element used in the ACCESS system.

An input signal, either 45H1 or 45H2, at terminal 1 is d.c.-restored upon the +45 normally present at terminal 4 and fires the trigger-cathode gap of the XC18. If terminal 6 be connected to 175H1, the main gap will fire only when a 45H1 is present. Similarly, if the anode is fed from 175H2, the main gap will fire only with a 45H2 signal. By removing the +45 bias from terminal 4, the tube will not fire with either signal.

When, however, the tube does fire, it develops across its cathode resistor a signal of 100V peak amplitude of whichever phase is applied to its anode. This is then fed out on terminal 9 to further elements.

It will probably be realized now that the use of half-wave rectified supplies produces a set of operating conditions that, in effect, combines the advantages of both well-known methods of firing cold-cathode tubes, namely; 'pulse-plus-bias' and 'd.c. voltage transfer'. For as long as an input signal persists, the tube fires 'continuously' to each positive half-cycle of its supply yet the facility of controlling conduction via an auxiliary input, the bias, remains.

Element RC4

The circuit, shown in Fig. 8, is used in conjunction with an R1 unit when a number of independent outputs are required from one input.

The 100V output from R1 is applied to terminal 1 and strikes the NT2 diodes. With a maintaining voltage of 55V four entirely independent signals of 45V peak amplitude are thus available at terminals 2, 4, 6 and 8.

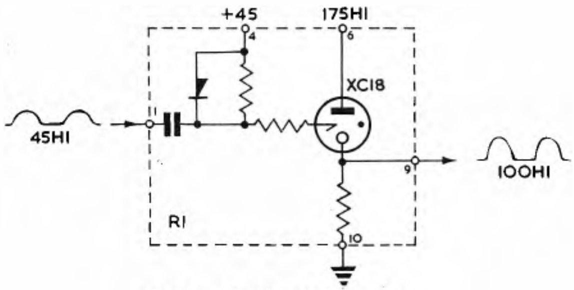


Fig. 7. Circuit of R1 element

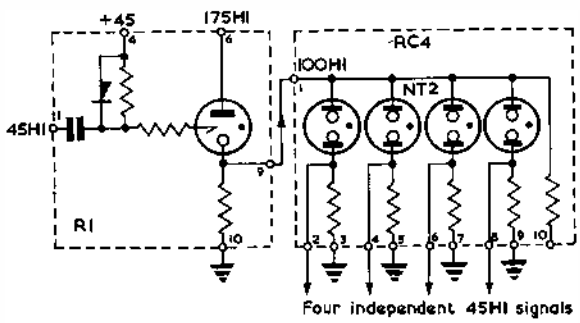


Fig. 8. Circuit of RC4 element

A similar element, RC2, is available with two diodes for when only two outputs are required.

It was mentioned earlier in connexion with the R1 element that the circuit operates in the pulse-plus-bias mode. Switching the bias gives an added control facility and this is frequently accomplished as shown in Fig. 9.

Here, in a typical transfer machine arrangement, +45 biases for a number of independent R1 circuits are provided by one RC4 element when +100 is switched to its terminal 1 by a single pole of an AUTO/MANUAL switch. In the MANUAL position the biases, and hence the firing of the R1 tubes, are brought under the control of the two MANUAL push-buttons. No matter how complicated the circuit, virtually any number of tubes can be controlled in this manner with the simplest of switching arrangements.

It is perhaps worth noting here that the outputs from both the R1 and RC4 type circuits are substantially independent of load. This follows naturally since the output voltages are derived from stabilized supplies via gas-discharge tubes which are themselves constant-voltage devices.

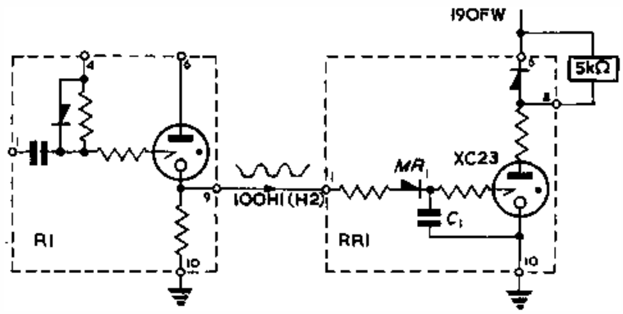


Fig. 10. Element RR1

Element RR1

This circuit, Fig. 10, can also follow an R1 unit and operates a BPO 3000 type relay for the switching of motor contactors and solenoid valves etc.

The 100V peak signal from R1 enters on terminal 1 and is converted to direct current by the rectifier and capacitor, MR_1 and C_1. The XC23 thus fires to every half-cycle of the 190V full-wave anode supply, irrespective of the phase of the input, and energizes the relay. While a half-wave anode

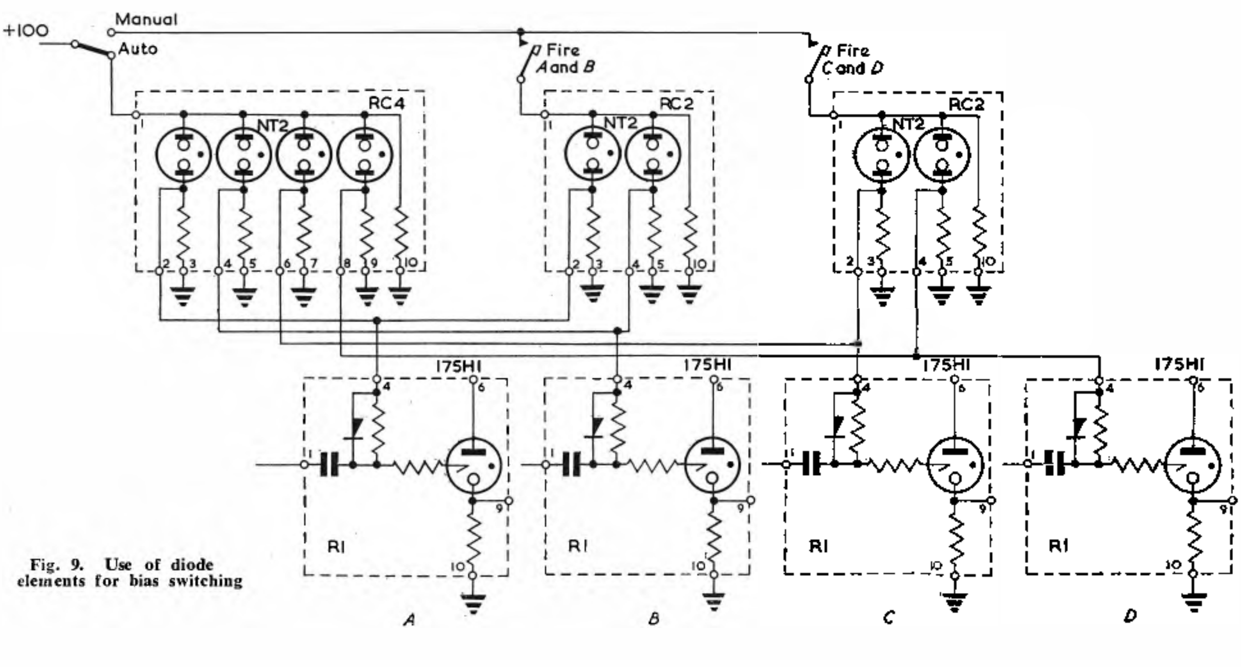


Fig. 9. Use of diode elements for bias switching

supply could have been used for this circuit, it would have led to high peak tube currents to ensure satisfactory relay operation.

Element PC1

Incoming signals from limit switches can be of two types, 45H1 or 45H2. Frequently, it is desirable to change automatically from one phase to the other and the element PC1, Fig. 11 is used for this purpose.

The limit switch first fires an R1 whose output, 100V of, say, phase H2, is then fed to terminal 1 of the PC1. The rectifier and capacitor, MR_1 and C_1, convert this signal to

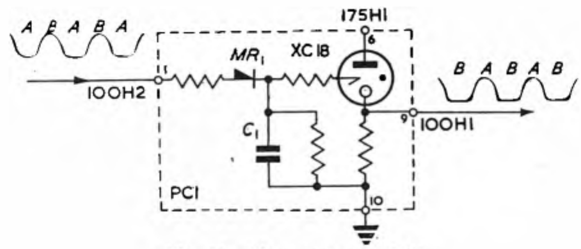


Fig. 11. Phase convertor PC1

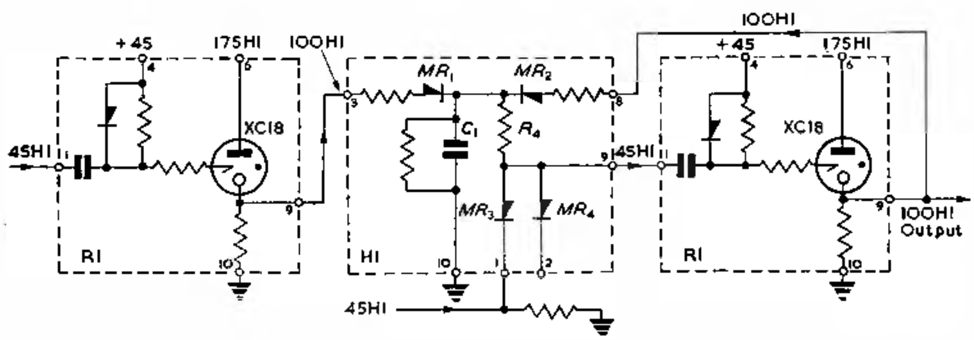
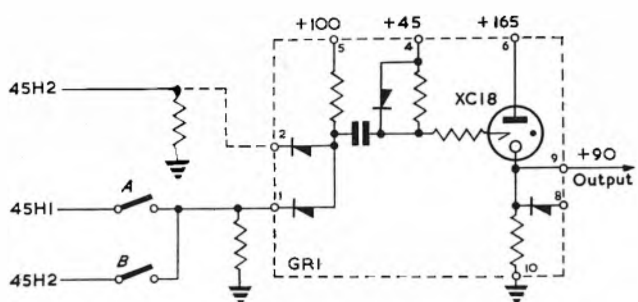


Fig. 12. Hold on element H1



Tube will fire only to limit *B*

Fig. 13. Temporary memory GR1

direct current and the XC18 conducts to each half-cycle of the 175H1 supply to which terminal 6 is connected.

The circuit thus accepts a 100H2 input and delivers a 100H1 output, or vice versa.

Element H1

Instances occur where it is required to fire a tube with one signal and hold it on with another. This mode of operation is accomplished as shown in Fig. 12.

A 100H1 or 100H2 signal from an R1 element is applied to terminal 3 where it is converted to direct current by the rectifier and capacitor MR_1 and C_1. If, now, a 45H1 (or H2) signal is present at terminals 1 or 2 of the AND matrix MR_3, MR_4 and R_4, then a 45H1 (or H2) signal is fed out on terminal 9 to fire a following R1 element. Once fired, this R1 element feeds back 100H1 (or H2) from its cathode to terminal 8 of the H1 element, where it is converted to direct current in the same manner as the original incoming 100V signal. This latter signal may now disappear but the second R1 circuit will remain on, due solely to its own feedback and the hold-on signal present at terminals 1 or 2, or both. When the hold-on signal terminates, the second R1 tube will extinguish.

Element GR1

In some cases it is necessary to operate tubes with a d.c. anode supply so that they will fire to the closing of a limit and remain fired as a temporary memory. With such a d.c. supply, the cold-cathode tube is no longer phase-conscious and it becomes necessary to introduce this feature into the trigger circuit. Such an element is the GR1, shown in Fig. 13.

Assuming that terminal 6 is connected to +165, terminal 4 to +45 and terminal 5 to +100, then with either a 45H1 or 45H2 signal at terminal 1 only, the XC18 will fire and remain fired until the anode supply is disconnected. If, however, it is required that the circuit respond only to 45H2 signals even though signals of the opposite phase may at some time or other be present, then terminal 2 of the input AND matrix may be permanently connected to 45H2. Under these circumstances, the matrix will fire the tube only when 45H2 signals appear at terminal 1.

Matrix Elements

Considerable use is made throughout the equipment of simple rectifier matrices connected so as to perform the logical operations of AND and OR. Two representative circuits are illustrated in Fig. 14.

The AND/OR element SM2, Fig. 14(a), is used to accept 45H1 (or H2) signals from limit switches and will deliver an output at terminal 9 only if similarly-phased signals are present at all of the terminals 1, 2 and 3. An output at terminal 8 will occur if a signal, of any phase, occurs at any one of the inputs.

Element CM5, Fig. 14(b), is used chiefly with the outputs of RC4 circuits. It is a simple AND configuration and will deliver an output only if similarly-phased signals are fed to all its inputs.

Timers

There is relatively little need in transfer machine control for timers but the unit T1 is available when required.

The circuit is illustrated in Fig. 15, and with no input to terminal 1, the voltage across capacitor C_1 is but a few volts above earth. When +100 is applied to terminal 1, however, the rectifier MR_1 cuts-off and C_1 commences to charge towards +165 at a rate determined by the product R_1C_1. As the voltage reaches the trigger-cathode striking potential, the XC23 conducts and the relay is energized. It will not release until the +100 input is removed.

The Complete System

Having described briefly the major circuit elements that comprise the ACCESS system, it is now possible to show how they may be combined together to effect control of a transfer machine. Only a representative selection of circuits will be described, however, since to attempt to cover the complete control would take far more space than is available here and, in any event, would be outside the strict scope of this article.

CLAMP/UNCLAMP CIRCUIT

This is a fundamental circuit in any transfer machine and serves to check that:

(a) All components are rigidly clamped in position prior to, and during machining

(b) All components are unclamped and free to move prior to transfer.

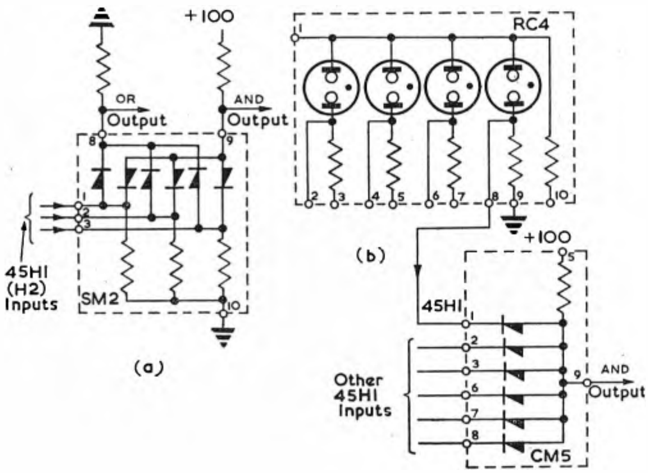


Fig. 14. (a) And/or matrix. (b) 6—input and matrix

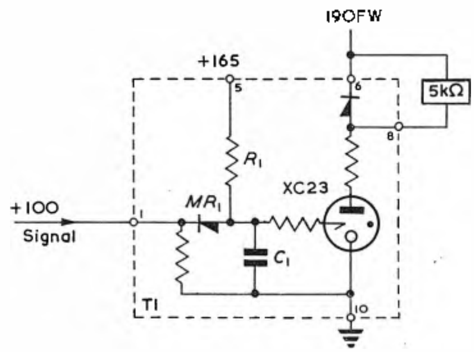


Fig. 15. Circuit of timer element

Each station of the machine is thus equipped with two limit switches. One of these is actuated when the clamping mechanism moves the component into its 'clamped' position, the other when the component is released. Signals from each pair are then fed down the machine to the Group Control where they act upon the circuit shown in Fig. 16.

Here it will be seen that the clamp and unclamp limits are fed from the two common supply lines 45H1 and 45H2, each pair signalling the Group Control via a single conductor. Within the control, these signals are applied to three-input AND matrices, type SM1 and also to pairs of neon indicators whose other electrodes are returned to −45H1 and −45H2. These neons form part of a plan-position display board and if it is assumed that, say, all the CLAMP limits are closed, then it will be realized from the waveforms involved that only those neons that are returned to −45H2 will be fired. Similarly, when all the UNCLAMP limits are closed and 45H1 signals are being transmitted, then only across the UNCLAMP indicators is there developed 90V peak-to-peak, which is sufficient to strike them. During this time, the CLAMP indicators have never more than 45V across them at any instant in the cycle and, consequently, remain extinguished.

Returning to the AND matrices referred to earlier, it will be seen that their terminal 9 outputs are commoned together and applied to two R1 elements, one in the UNCLAMP circuit, the other in the CLAMP circuit. By virtue of its 175H2 anode supply, the R1 in the CLAMP circuit can fire only to a 45H2 input and, hence, will do so when all CLAMP limits are closed and a 45H2 output is produced by the SM1 elements. The 100H2 output of the R1 element now passes to an RR1 element, where as already explained, it causes relay *CL* to be energized. This relay serves to apply power to the machine tool control panels only while the components are clamped and so avoids any possible accidental machining in the unclamped condition.

Three independent signals indicating that the machine is clamped are also required for use in the sequence circuits and as will be apparent later, these must be of phase H1. Therefore, in addition to feeding the RR1 element the 100H2 output of the R1 unit is applied to a PC1 circuit

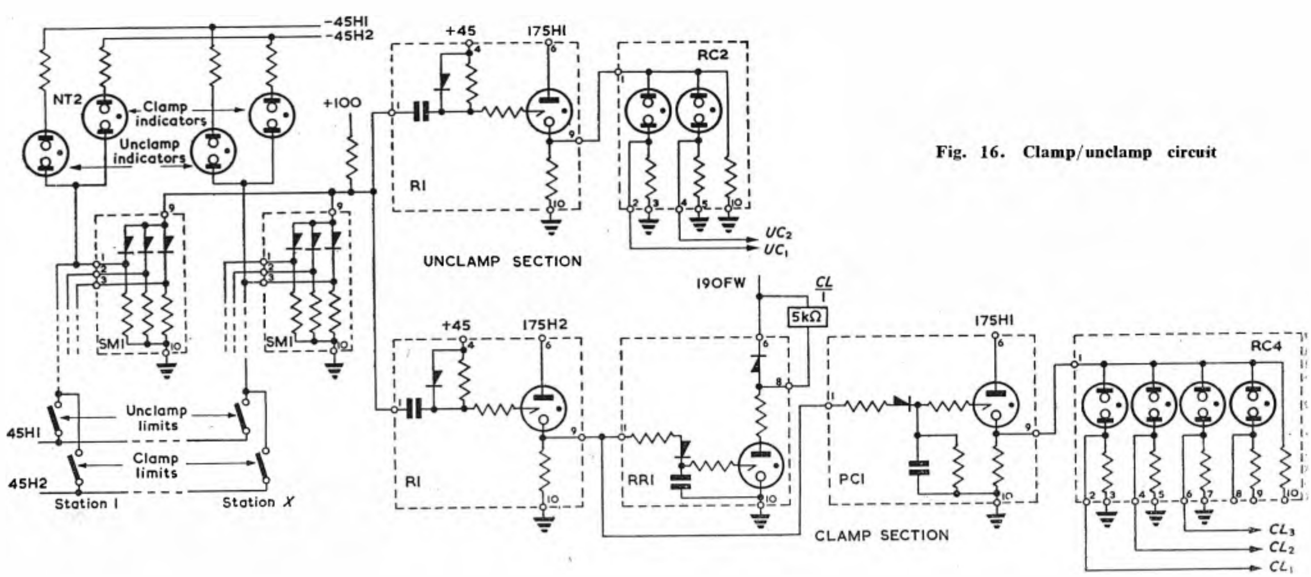


Fig. 16. Clamp/unclamp circuit

where it is converted to 100H1. This is then fed to the RC4 element and, as shown, the three 45H1 signals, CL_1, CL_2 and CL_3, are produced at terminals 2, 4 and 6.

Turning now to the UNCLAMP section of this circuit, it will be obvious that with a 175H1 anode supply, the tube of the R1 element can fire only to a 45H1 input. This will occur when all inputs of the SM1 matrices are receiving 45H1 signals and, hence, is indicative of the fact that all UNCLAMP limits are closed.

Only two signals signifying the unclamped condition are required later in the sequence circuits and these are derived from an RC2 element as illustrated.

It may be argued that to check that all the stations are clamped or unclamped, the simplest method is merely to connect all the appropriate limits in series. Experience has shown, however, that even when suitable indication is provided, the location of faulty limits in such a system can be very time-consuming and often reminiscent of locating burnt-out lamps in Christmas tree lights.

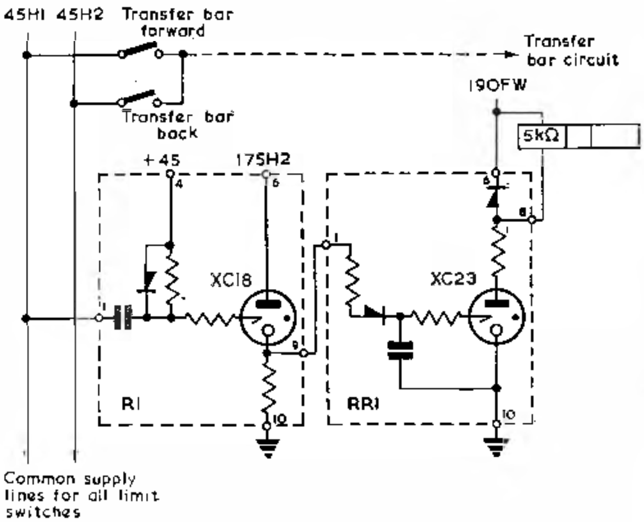


Fig. 17. Limit fault circuit

Limit Fault Circuit

One of the several ways in which this static control has improved upon previous systems is that, by a very simple circuit, it is able to detect the failure of practically every limit switch and stop the machine before any possible damage can occur.

In Fig. 17, two limit switches are shown transmitting either a 45H1 or 45H2 signal to the Group Control when the transfer bar of the machine is either fully forward or fully back respectively. The 45H1 and 45H2 supply lines are common to all limits on the machine.

It can probably be well imagined what would happen to a machine if, when operating automatically, the TRANSFER BAR FORWARD limit stuck closed and indicated to the sequence control that the bar was fully forward when, in fact, it was not. This condition, however, can be detected before damage occurs, because, if it is assumed that the bar goes forward normally, operates the FORWARD switch and returns leaving the switch stuck closed then, at the instant the bar re-operates the BACK limit switch, the 45H1 and 45H2 supplies will be connected together and become a full-wave supply, 45FW.

In the LIMIT FAULT circuit illustrated, the XC18 tube of the R1 element does not normally conduct since its anode is fed from 175H2 while its input is permanently connected to the 45H1 supply line. When, however, the 45H1 line becomes 45FW, as just outlined, then this tube does fire and its output, via the RR1 circuit, energizes one side of a mechanically-held two-position relay.

Operation of this relay trips out the main breaker and thus the damage that would occur if the bar went forward again is avoided.

Sequence Circuit

This, in effect, is the main circuit of the Group Control. It initiates all the various machine movements in the correct order, checking prior to each step that all the conditions for safety are satisfied. On the machine in question, there are five major steps in a cycle of operation and, in Table 2, these are given together with the conditions involved.

From Fig. 18, it can be seen that the circuit begins with five AND matrices having an appropriate number of inputs. Each matrix feeds an R1 element which, in turn, fires a common RR2 element so as to drive a uniselector one step forward. At each step, bank 2 of the uniselector energizes appropriate relays which then switch either pneumatic or hydraulic solenoid-operated valves.

The input signals to each of the five AND matrices are marked with the abbreviated circuit notation set out in Table 2. These signals are all 45H1 and emanate from diode elements such as the RC2 and RC4 shown in the CLAMP/UNCLAMP circuit, Fig. 16. Thus, for example, the input UC_1 on the TRANSFER matrix represents either a 45H1 signal when the machine is unclamped or, when the machine is not unclamped, a resistance-to-earth condition.

Assuming, then, that the machine is just unclamping and that all the other conditions necessary for TRANSFER are satisfied, 45H1 signals will be present at all the used inputs of the TRANSFER matrix with the exception of UC_1. At the

TABLE 2

Sequence of operations of transfer machine

SEQUENCE STEP	OPERATION	CONDITIONS TO BE SATISFIED	SIGNAL NOTATION
1.	TRANSFER	All stations unclamped	UC_1
		All unit heads back, clear of components	TL_2
		Unload station vacant	USV_1
		Unload roller track vacant	UTV_1
		Component available for loading	CA_1
2.	CLAMP ALL COMPONENTS	Transfer bar forward	TBF_1
		All unit heads back, clear of components	TL_3
		Load and unload bars back	$LUBB_1$
3.	LOAD AND UNLOAD BARS FORWARD	All components clamped in position	CL_1
		Load station vacant	LSV_1
	AND START UNIT HEADS	All heads back, clear of components	TL_4
		All components clamped	CL_2
4.	LOAD, UNLOAD AND TRANSFER BARS BACK	Load and unload bars forward	$LUBF_1$
		All components clamped	CL_3
5.	UNCLAMP ALL COMPONENTS	Transfer bar back	TBB_1
		Load and unload bars back	$LUBB_2$
		Load station full	LSF_1
		All heads back, clear of components	TL_1
		All heads have machined to depth	BL_1

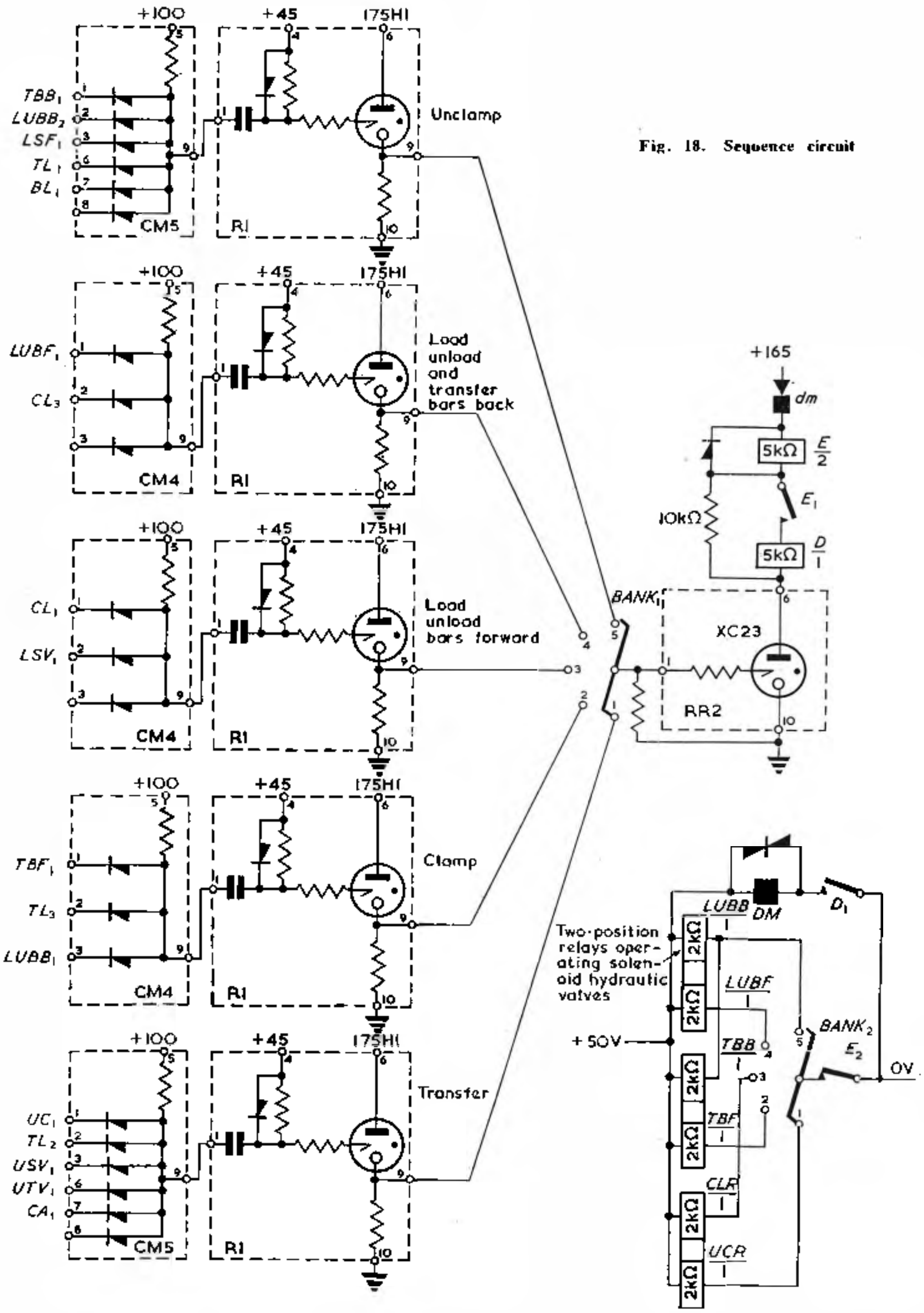


Fig. 18. Sequence circuit

instant the machine completes unclamping, however, 45H1 will appear at UC_1, the R1 element will fire and a 100H1 signal will be fed through bank 1 of the uniselector to fire the RR2 element. This XC23 tube now conducts to +165 via the uniselector interrupter contacts *dm*, the relay *E* and the 10kΩ resistor in shunt with contact E_1 and relay *D*. Relay *E* thus energizes and at contact E_2 interrupts the 50V supply from bank 2 to the sequence relays prior to stepping. Simultaneously, contact E_1 switches relay *D* into the XC23 anode circuit and contact D_1 subsequently energizes the drive magnet *DM* of the uniselector, which is of the forward-acting type. This now steps forward and as the wiper leaves contact 1 of bank 1, the 100H1 signal is disconnected from terminal 1 of the RR2 unit. The tube continues to conduct, however, until the uniselector completes its step, when the interrupter contacts *dm* open and break the d.c. supply. Both relays now release, *E* being slow due to the shunt rectifier, and at contact E_2, the 50V supply is restored to bank 2 where, via contact 2, it energizes the TRANSFER BAR FORWARD relay, *TBF*.

At the completion of the transfer movement, the TBF_1 signal will appear at the CLAMP matrix. If both the TL_3 and $LUBB_1$ signals are also present. the associated R1 tube will fire and the uniselector will make yet another step forward. This time, bank 2 will energize relay *CLR* and the machine will clamp up.

Thus it will be apparent that, provided all the 45H1 interlock signals are present at the relevant times, the sequence circuit will continue stepping, energizing the appropriate solenoids to give an automatic cycle. Manual facilities whereby each step can be controlled individually and, in some cases, be performed out of sequence, are also included but their circuit arrangements are omitted for clarity.

Conclusion

Although it was not practicable in an article such as this, to describe the entire system, it is hoped, nevertheless, that the circuits selected have presented a reasonable picture of the way in which ACCESS is applied.

Of the nine transfer machines so far equipped, the first was completed in December 1957 and has now been in continuous use, day and night, for some eighteen months. During this period the ACCESS Group Control has needed only 8½ hours maintenance and the replacement of two triodes and four diode indicators that failed very early in the life of the equipment.

It may be of interest in conclusion to explain why cold-cathode tubes were chosen as the basis of the system. Unlike both magnetic amplifiers and transistors, this type of tube possesses the enviable property of self-indication. To the maintenance man who has to keep an expensive machine running, this is of inestimable value and enables him to trace faults quickly and with the very minimum of test gear.

Acknowledgment

The author would like to thank the Directors of The Austin Motor Co. Ltd for permission to publish this article and to acknowledge the encouragement and advice given by Mr. F. W. Highfield who has since retired from the position of Superintendent of the Electronics Department.

A Vectorscope Unit

By K. G. Freeman*, B.Sc.

A vectorscope unit is described which provides a vectorial display of NTSC-type chrominance signals. Of high sensitivity, it incorporates a number of features, including means for continuously checking gain and quadrature adjustments. The instrument permits rapid assessment of chrominance signal phase and amplitude characteristics—which are of particular importance in direct-decoding operation of single gun colour television tubes.

(Voir page 706 pour le résumé en français: Zusammenfassung in deutscher Sprache Seite 711)

IN the NTSC colour television system the chrominance information is carried by a sub-carrier which lies in the high-frequency region of the video spectrum and is modulated in quadrature by the so-called I and Q colour difference signals. In addition a reference phase is provided by a burst of unmodulated sub-carrier at a fixed phase with respect to the quadrature modulation axes. Thus, any chrominance signal of given hue and saturation is represented by a phasor of appropriate phase and amplitude with respect to the reference phasor. Fig. 1(a) shows a single line of a typical NTSC colour bar composite video signal and Fig. 1(b) the corresponding chrominance signal phasor diagram.

It will be evident that if such a signal is synchronously detected along two axes in quadrature and the outputs from the detectors applied to the X and Y plates of an oscilloscope tube the display on the tube will be similar to the phasor diagram of Fig. 1(b). This is the basic principle of the vectorscope in which the arrangement of synchronous detectors is virtually identical with that in the low-level decoder used in early three-gun tube receivers. For a colour bar signal the vectorscope display will consist of a number of bright spots whose angular position and distance from the centre correspond to the phase and amplitude of the sub-carrier signal forming the bar pattern, joined by faint lines corresponding to the transitions between successive colours. For a normal colour picture, of course, the display will consist of a rather complicated 'fuzz', so that it is usual to employ the colour-bar signal to assess the chrominance signal characteristics.

The vectorscope display of the observed colour-bar chrominance signal may differ from the NTSC standard for several reasons, such as incorrect encoding, and phase distortion in the transmitter and the receiver i.f. amplifier etc. However, this form of display is also of considerable value in the study of direct-decoding circuits for single-gun colour tubes.

In order to improve the quality of reproduction of colour pictures with these tubes various forms of signal processing may be employed, which have the effect of modifying the phases and amplitudes of the chrominance phasors. In such circumstances, a vectorscope provides the easiest means of examining the form of the modified chrominance signal.

Although various types of vectorscopes have been described in the literature[1-6], the design of the one described in this article represents a somewhat different approach, and complete circuit details are given.

Basic Design Considerations

The vectorscope was intended primarily as a laboratory tool, but as it seemed possible that there might ultimately be a more general demand for such an instrument, for example, in the servicing field, it was decided that it should be as simple as possible consistent with the availability of adequate controls. The vectorscope unit was therefore designed to directly drive the X and Y deflexion plates of the Mullard oscilloscope tube type DG 13-2, as used in the Mullard type L101 dual trace oscilloscope. The requisite

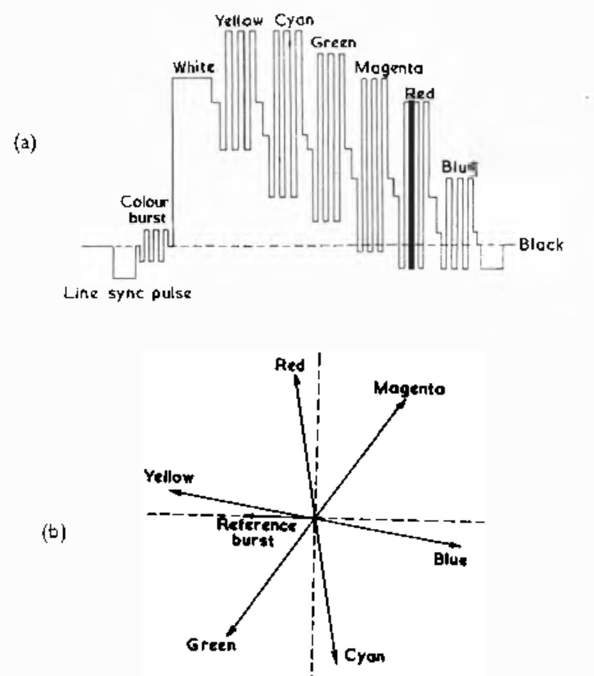


Fig. 1. (a) Single line of a typical NTSC colour bar composite video signal (b) Phasor diagram of chrominance signal of Fig. 1(a) colour bar pattern

gain and shift controls are incorporated in the driving amplifiers following the synchronous detectors.

A further simplification is possible, at the expense of some slight sacrifice of performance, by making use of the X and Y amplifiers of the L101 oscilloscope. In this case the long-tailed pair output stages can be dispensed with—although some means of gain control will still be required.

For operation of the synchronous detectors a source of reference frequency (sub-carrier) is required which is phase-locked to the colour 'burst'. In more elaborate vectorscopes this is derived from the signal under investigation by means of a burst gate and a crystal oscillator which is locked by an automatic phase-control (a.p.c.) loop. This was felt to be an unnecessary complication as a source of locked sub-carrier is usually available in practice in either the encoder or decoder.

For the vector display to have any meaning it is essential that the synchronous detectors operate exactly in

* *Mullard Ltd.*

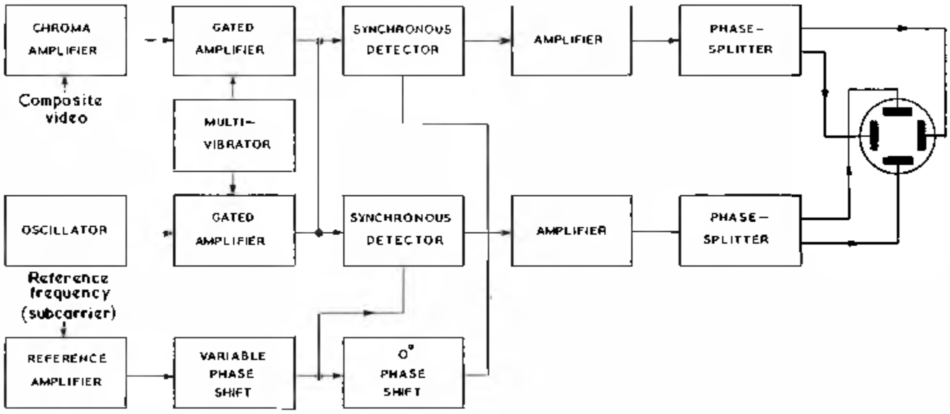
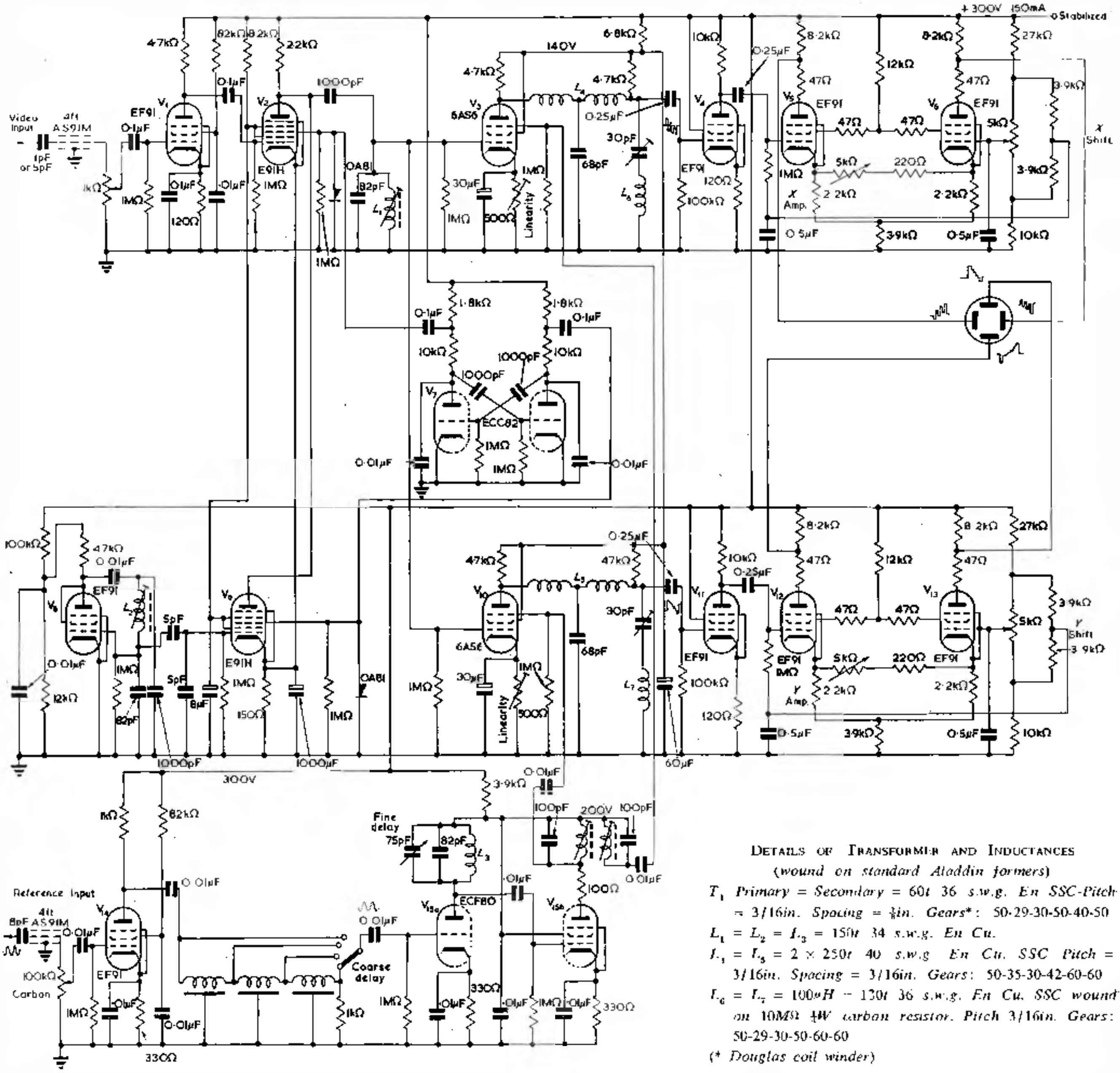


Fig. 2. Block diagram of vectorscope unit

Fig. 3. Vectorscope unit translator *(all resistor 1/4W except where otherwise stipulated)*



DETAILS OF TRANSFORMER AND INDUCTANCES
(wound on standard Aladdin formers)

T_1 *Primary = Secondary = 60t 36 s.w.g. En SSC-Pitch = 3/16in. Spacing = 1/8in. Gears*: 50-29-30-50-40-50*

$L_1 = L_2 = L_3$ *= 150t 34 s.w.g. En Cu.*

$L_4 = L_5$ *= 2 × 250t 40 s.w.g En Cu. SSC Pitch = 3/16in. Spacing = 3/16in. Gears: 50-35-30-42-60-60*

$L_6 = L_7$ *= 100µH — 130t 36 s.w.g. En Cu. SSC wound on 10MΩ 1/4W carbon resistor. Pitch 3/16in. Gears: 50-29-30-50-60-60*

(Douglas coil winder)*

quadrature and that the overall X and Y channel gains be equal. Provision for ensuring this state of affairs is therefore necessary, and in this unit it is done as follows. A stable oscillator free runs at a frequency near to sub-carrier and this is gated alternately with the chrominance signal being examined. This produces on the screen of the tube an elliptical or circular trace superimposed upon the vectorial display due to the chrominance signal. Under correct gain and phase conditions the trace is a circle and it is a simple matter to make the appropriate adjustments to obtain such a circular trace, which is always present, thus providing a continuous check of the operating conditions.

Finally, since the phase of the chrominance signal and its reference burst varies through the circuit under examination the trace on the screen may have any angular orientation. As it is desirable to be able to have a fixed orientation of trace (e.g. with the colour burst phasor in the $-x$ direction) for comparison against an engraved scale, it is therefore necessary to have provision for changing the phase of synchronous detection. This is most easily achieved by having a variable delay in the reference frequency amplifier with a range of 0° to 360° of one cycle of sub-carrier. Again this is a feature included for convenience in the laboratory and, in general, could be omitted if a rotating graticule is used.

Fig. 2 shows the block diagram of the vectorscope unit embodying these principles.

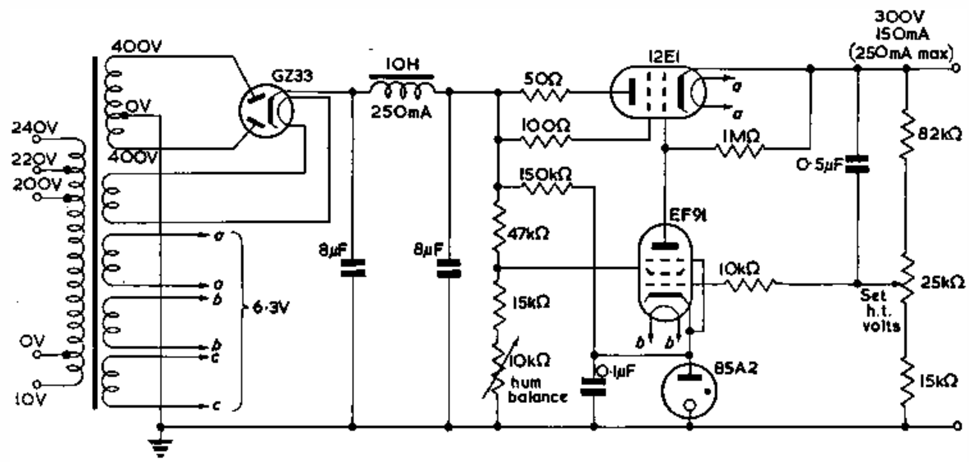


Fig. 4. Vectorscope unit power supply

Detailed Description of Vectorscope Unit Translator

Fig. 3 shows the complete circuit diagram of the vectorscope unit (less its power supply) which will now be described in more detail. The composite video under investigation is applied via a 1pF probe to the amplifier V_1. Its low capacitance ensures that the signal is not significantly modified by the probe, and it also has the additional advantage of removing most of the unwanted lower-frequency luminance component. The output from V_1 is then applied to the first grid of V_2 (E91H) which is a dual-control valve. On the first grid of a second E91H, V_9, is applied the output from the oscillator V_8.

This oscillator is adjusted to free-run at a frequency within some 10 to 20kc/s of sub-carrier, and is operated from an h.t. line of only about 30V to maintain stability and keep radiation to a minimum. Even so, the unit is so sensitive that additional attenuation is necessary to give the desired low-level of local oscillator voltage on the grid of V_9.

The two valves, V_2 and V_9 are operated alternately for equal periods of time by driving the second control grids with the output from a symmetrical multivibrator V_7 which has a frequency of about 300c/s. The multivibrator operation is conventional but 0·01μF capacitors are connected from the triode anodes to earth to slow the multivibrator waveform edges, and give a gated output from V_2 and V_9 devoid of overshoots. D.C. restoration of the multivibrator waveform enables the use of a minimum amplitude and better operating conditions for the E91H's.

The gated output appearing at the commoned anodes of V_2 and V_9 is limited to a bandwidth of about 1Mc/s centred on the sub-carrier frequency by the damped tuned anode circuit, which ensures that only chrominance video information is passed to the synchronous detectors. Since the vectorscope is used only on colour-bar test signals and the edges corresponding to the fast chrominance transitions are not of prime importance, the damped single-tuned circuit is adequate. For convenience 6AS6 short-suppressor-grid-base pentodes are used as the synchronous detectors and with some adjustment of operating conditions give a sufficiently linear performance at the low signal levels involved. E91H dual-control valves were tried but found unsatisfactory. Reference frequency (sub-carrier) for the operation of the synchronous detectors is obtained through a low-capacitance probe, amplified by V_{14} and applied to a terminated length of 1kΩ delay cable (delay ~70mμsec/ft) which is tapped so that delays of 0°, 90°, 180° and 270° of one sub-carrier cycle are obtainable. Fine delay is obtained by detuning of the anode circuit of V_{15} triode. The output is fed to the pentode of V_{15} which drives a double-tuned transformer T_1. By suitably tuning the primary and secondary circuits, sub-carrier outputs with the desired quadrature relationship for operation of the synchronous detectors are readily obtained. The overall sensitivity of the reference amplifier is such that correct synchronous detection—independent of reference amplitude—is obtained for a reference voltage input to the 8pF probe of more than approximately 0·5V r.m.s. As in receiver-decoder circuits the synchronous detectors are followed by low-pass filters, but in this case they are identical, with a cut-off frequency of about ½Mc/s. To further attenuate the reference frequency component in the output, series-tuned circuits resonant at sub-carrier are connected between earth and the low-pass filter outputs.

The outputs from the filters, after amplification by V_4 and V_{11}, are applied to the long-tailed pair output stages V_5V_6, $V_{12}V_{13}$. These are of convential design, with gain and shift controls and provide sufficient output to drive directly the X and Y plates of Mullard L101 oscilloscope tube type DG13-2 via a yard or so of four-core unscreened lead.

The overall sensitivity using the 1pF probe is such that a full-size trace on this oscilloscope tube is obtained for a chrominance signal amplitude of about 0·3V r.m.s. For rather smaller chrominance signals a 5pF probe should be adequate.

The Vectorscope Unit Power Supply and Layout Considerations

It is clear that an instrument of this sensitivity needs a well-stabilized and hum-free power supply. Fig. 4 shows a suitable power supply for the unit described, which has a hum level of only about 2mV and gives a sufficiently hum-free trace.

The greatest difficulty encountered in the design and construction of this vectorscope unit is that radiation between various stages of the 'translator' leads to distortion of the trace due to the presence of spurious sub-carrier at the input to the synchronous detectors. The greatest care in the design of the chassis layout is therefore recommended, particular attention being paid to ensuring that the reference and chrominance amplifiers and the free-running oscillator are all adequately screened from one another.

Conclusions

A vectorscope unit for laboratory work which is likely to be of general use has been described. Of adequate stability and high sensitivity it enables rapid checks to be made on NTSC type encoders, decoders and signal translators.

Acknowledgments

The author wishes to thank the Director of Mullard Research Laboratories and the Directors of Mullard Limited for permission to publish this article.

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A Ship Hydrodynamics Laboratory

A new Ship Hydrodynamics Laboratory of the Department of Scientific and Industrial Research was opened officially by H.R.H. the Duke of Edinburgh at Feltham, Middlesex, on 19 October.

The new Laboratory, which will be part of the Ship Division of the National Physical Laboratory, has three main facilities:

(1) A large towing tank 1 300ft long, 48ft wide and 25ft deep with an associated wavemaker.
(2) A large water tunnel with a measuring section of 44in diameter.
(3) A sea-keeping and manœuvring basin 100ft square with a water depth of 8ft.

In addition, there are workshops for the making of model hulls and propellers, a vibration laboratory and instrument laboratories for development, testing and calibration purposes.

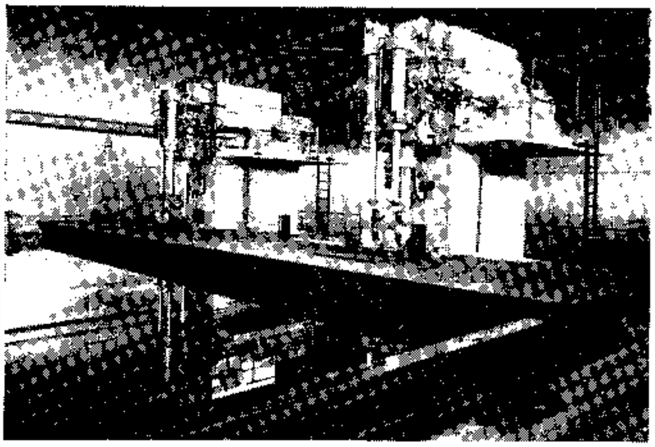
These new and up-to-date facilities for hydrodynamic research which are claimed to be the most advanced in Europe are designed to reproduce in the Laboratory more nearly realistic sea conditions than has been possible in the past. The main aim of the research programme will be to build up basic knowledge for the design of ships that can maintain high speeds in rough seas without danger and with a minimum of discomfort.

The large towing tank will enable much bigger models to be used than at Teddington and it will also allow studies to be made on submerged models of submarine cargo ships and tankers when these become necessary.

The wavemaker in the towing tank consists of a large wedge-shaped plunger weighing 20 tons, installed at the remote end of the tank.

The plunger is set in motion by hydraulic cylinders and is controlled by an electronic-hydraulic servo mechanism which continuously monitors the position of the plunger relative to a correct position determined by a control setting and applies a correcting action if any error exists.

The wavemaker installation at the remote end of the large towing tank



The electronic equipment for the system was supplied by Associated Electrical Industries Ltd, Electronic Apparatus Division, Leicester.

The circuits can be divided into two groups:

(a) Circuits required to produce signals which direct the movement of the wedge in accordance with specified requirements.
(b) Circuits which constrain the wedge to follow the director signals to the required degree of accuracy.

The wedge produces waves having amplitudes and frequencies which are infinitely variable between set limits, the amplitude limits being 0 to 1·5ft, the frequency being 1 cycle of 2·8sec to 1 cycle of 0·95sec. The frequency is achieved by driving a Selsyn through a precision gearbox from a motor controlled by a high-accuracy speed regulator. The output from the Selsyn excites an inductive potentiometer which provides a control of the amplitude. The setting of the potentiometer is remotely controlled from a desk by means of an accurate position-control sub-servo.

The controls operate through sub-servos for the purposes of convenience of control and prevention of excessive acceleration and velocity on the plunger. An auxiliary function of the sub-servos is to establish a sequence of operations when starting and finishing tests.

Controlling the movement of the wedge is the most difficult problem associated with the wavemaker. Because of the great length and mass of the wedge, it was considered impracticable to drive it from a single actuator, and in consequence, the ends of the wedges are driven by ram systems which are completely separate from one another. Thus, the servos associated with each end of the wedge must be carefully matched. This is achieved by a comprehensive scheme of local feedback loops, so designed that the hydraulic elements within the loops have their individual characteristics largely overridden.

The feedback loops are excited from displacement pick-off devices measuring the movement of hydraulic valves, and from tacho-generators and Selsyns measuring the position and speed of the main pump's wash and the main ram.

Tests on hydrofoil sections under controlled pressures up to six atmospheres can now be made in the new 44in diameter water tunnel. The tunnel will also permit tests on propeller models up to 24in diameter.

The sea-keeping tank is designed so that experiments can be carried out on a model in waves at varying angles to its course and in irregular and confused seas. Models will be free running and remotely controlled and considerable thought has been given to the design of the control and recording equipment which has to be kept as light and compact as possible. Instrument data will be either recorded in the model or telemetered ashore.

Fitting up and trimming of models which may be up to 10ft in length will be carried out in the manœuvring tank.

Compound Semiconductors

By D. A. Wright*, D.Sc., M.Sc., F.Inst.P.

This article first discusses the more important parameters governing the behaviour of semiconductors. These considerations are then applied to the various types of semiconducting compound which have so far been studied, and values of the mean parameters are quoted. The usefulness of the materials for different applications is discussed in the light of these figures. In each type of compound, definite trends are observed in the relationship between the semiconducting parameters and other physical properties. These trends are described and their interpretation is indicated. The parameters which are important for thermoelectric applications are mentioned briefly, together with the bearing of the earlier discussion on the choice of materials for these applications.

(Voir page 707 pour le résumé en français: Zusammenfassung in deutscher Sprache Seite 711)

RESEARCH and development on semiconductors have been concentrated during the last ten years primarily on germanium and silicon, because of their application to rectifiers and transistors. However, at the same time interest in semiconducting compounds has been maintained and has increased recently. Several semiconducting oxides and sulphides have been known for many years, e.g., zinc oxide, copper oxide and lead sulphide. Since 1950 numerous other semiconducting compounds have been studied, in particular binary compounds between a metal and either selenium, tellurium, phosphorus, arsenic or antimony. The objectives in this work have been twofold; firstly to try to find materials which might be superior to germanium and silicon for making pn junction diodes and triodes; secondly, to find new effects which might ultimately have commercial applications or new materials in which known effects might have much greater magnitude, and therefore usefulness, than in materials known at present.

Probably the most thorough investigation has been made of the compounds between Group III_B elements, e.g., aluminium, gallium or indium with Group V_B elements, e.g. phosphorus, arsenic or antimony; these compounds have similar crystal structure to that of germanium, and are not too difficult to prepare with a high degree of purity and in stoichiometric proportions.

Another class fairly extensively investigated is that formed by compounds between Group VI_B elements, e.g. sulphur, selenium or tellurium with Group II_B metals, e.g. zinc, cadmium or mercury, or with lead, which is related to these metals. These are more difficult to make stoichiometric, while some of them can exist either in a hexagonal crystal structure or in cubic form related to that of germanium. The transition between these forms as a crystal grows often leads to 'stacking disorder' in regions where locally the atoms form patterns intermediate between those characteristic of the two stable structures. These imperfect regions may have a considerable effect on the properties of the crystals, as discussed below.

The elements germanium, silicon and diamond have a crystal structure in which each atom has four nearest neighbours and is bonded to each of these with two electrons symmetrically placed between the atoms. This is a 'covalent' bond. In many of the compounds referred to above the structure is similar, but the electrons in the bond are not situated symmetrically so that there is a small effective charge on each of the atoms. Compared with purely covalent bonds, these partial ionic charges increase the extent to which electrons are scattered when moving in the crystal. The consequences of this increased scattering are also discussed below. In the more extreme cases of compounds between Group I and Group VII elements, e.g., sodium chloride or silver bromide, the asymmetry is much greater. The former compound consists of positively charged sodium ions and negatively charged chlorine ions, the bond being ionic and not covalent. These compounds are insulators at ordinary temperatures, and on heating develop ionic rather than electronic conductivity.

There are numerous other types of binary compound with somewhat more complex structures which are now known to be electronic semiconductors. Some such compounds and their properties are indicated in Tables 5, 6 and 7. There are in addition many possible types of ternary compound semiconductor, some of which are being investigated at the present time. Many of these are related structurally with one or another of the series of binary compounds already described. Because of this some rough prediction of their properties can be made, and it is possible to select out of a very large number of possibilities those compounds which are most likely to be worth studying. Those investigated up to the present are difficult to prepare as single crystals, and to obtain in stoichiometric proportions.

Most of the compounds mentioned so far fit broadly into the theoretical approach to semiconductor physics built up in recent years, based on the 'band theory' of solids. When this is so, the compounds may be described as 'normal' semiconductors. However, there are other compounds which have semiconducting properties, but to which the band theory does not apply, and some of these may become of increasing technical importance. The series In_2Te_3, Ga_2Se_3, etc., fall into this class, as do the refractory oxides of nickel, iron and other transition metals containing appropriate impurities. These have not been very fully studied so far; the main differences between them and normal semiconductors are discussed below.

Electrical Conductivity

The flow of electric current in a semiconductor depends on n, the density of free carriers of electric charge, and on μ, the 'mobility' of these charge carriers. The mobility is the drift velocity acquired by the carriers in the crystal in unit field, so that the conductivity σ has the value $ne\mu$, where e is the charge on the electron. The carriers can be either electrons or 'holes', according to the type of semiconductor, or both may be present at the same time, in which case the density and mobility of both must be taken into account.

In a pure crystal, having stoichiometric proportions of its constituents if it is a compound, the density of electrons, n_e, is equal to that of holes, n_h. The density depends on the

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'effective masses' of the electrons and holes, and varies exponentially with temperature and with the forbidden energy gap of the crystal. This is the 'intrinsic' case. The variation with temperature is illustrated in Fig. 1. When certain types of impurity or of lattice defect are present, at a density n_0, they may provide localized donor or acceptor levels within the forbidden gap. When these are shallow, at a small energy separation from the nearer band edge, the levels are all ionized at quite low temperatures, and the density of free carriers is then equal to n_0. The carriers concerned are electrons if the levels are 'donor' levels near the edge of the conduction band, holes if they are 'acceptor' levels near the edge of the valence band. As shown in Fig. 1, the carrier density is constant at the value n_0 over a considerable temperature range, until ultimately as temperature rises the intrinsic carrier density becomes greater than n_0. Over the range where it is less, the semiconductor is 'extrinsic'. Semiconductors are normally used in the temperature range where the extrinsic

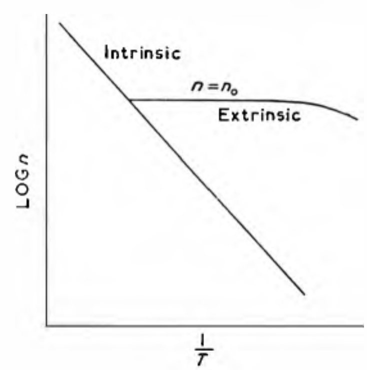

Fig. 1. Variation of carrier density n with absolute temperature t

carrier density has the constant value n_0, either electrons (n-type) or holes (p-type). The values of n_0 required lie typically between 10^{14} and $10^{18}/\text{cm}^3$, according to the application. Whatever value is decided upon to meet the requirements of the device must be obtained quite accurately and reproducibly, so that the technology for making and controlling the semiconductor is correspondingly difficult.

If the energy gap for a semiconductor is QeV, the intrinsic carrier density is proportional to $\exp(-Q/2kT)$, where T is the absolute temperature and k is Boltzmann's constant. The temperature at which a semiconductor becomes intrinsic, i.e., at which the intrinsic carrier density becomes greater than n_0, depends on the value of n_0, but is clearly larger, the larger the value of Q. For values of n_0 near 10^{16}, which are typical for many applications, InSb with a gap of 0·17eV becomes intrinsic at 20°C. Ge and Si respectively have gaps of 0·65 and 1·1eV and become intrinsic at 150°C and 300°C. For most applications therefore these materials can only be used below these temperatures.

Mobility

For most applications μ is required to be as large as possible, so that it is necessary to consider factors controlling its magnitude. It should be noted that in general it is not required to have large values of $\sigma = ne\mu$. The operation of semiconductor devices frequently depends on changes in the value of σ due to changes in n, and such changes are of greater significance if n and therefore σ are not initially too large. However, in transistors one needs a high μ in order to obtain a high diffusion constant $D = (kT/e)\mu$ and a high diffusion length $L = \sqrt{(D\tau)}$, where τ is the 'lifetime' discussed below. A detailed consideration of transistor action shows that the mobility of both types of carrier is important, and that there is a figure of merit for high frequency operation given by $\mu_e\mu_h/\epsilon^4$, where ϵ is the dielectric constant. μ_e is the mobility of the electron and μ_h that of the hole. Again, in applications of the Hall effect, a high open-circuit voltage is obtained provided n is small, but to obtain a useful power output μ must be large, in fact the power obtainable is proportional to μ^2. Photoconductivity depends in magnitude directly on μ as well as on the lifetime τ, while as will be discussed further below, thermoelectric refrigeration or generation requires, among other things, a high value of μ.

The mobility depends on the 'effective mass' of the charge carriers and on the details of the processes by which they are scattered when moving through the crystal. The concept of effective mass arises because the electron moving in the crystal, even in the absence of any collision processes, is in a region of electrostatic potential different from that in vacuum, so that its laws of motion are quite different. When its energy alters, part of the change in momentum is shared with the crystal lattice, so that the effective mass differs from that in vacuum, and is moreover a function of energy and of direction of motion in the crystal. Different methods of averaging are used to define effective mass parameters applicable to different types of process.

Solid-state theory is not yet developed sufficiently to make it possible to calculate and predict from the crystal structure what the details of the effective mass will be in any semiconductor. There is a certain amount of experimental information, but it is available in detail only for germanium, silicon and indium antimonide.

As regards the scattering processes, the electrons may be scattered by vibrations of the crystal lattice, by charged impurity atoms, or by uncharged impurities or defects in the crystal. The scattering by the lattice, leading to a mobility μ_L, may involve the 'acoustic mode', in which neighbouring atoms are displaced in the same direction, or 'polar modes', in which they are displaced in opposite directions. With covalent bonding the former dominate the scattering process, and lead to a lattice mobility which falls as temperature rises, varying as $T^{-3/2}$, as shown in Fig. 2. In crystals where there is some ionicity, the opposite charge on neighbouring ions results in considerable scattering by vibrations in the polar mode. This scattering is likely therefore to be greater than that produced by the acoustic mode. It leads to a lattice mobility which again falls as temperature rises, though the relation may be either more or less steep than $T^{-3/2}$. As the ionicity increases this scattering effect becomes larger, until with the alkali halides, e.g., sodium chloride referred to above, polar mode scattering causes very low mobility. When the ionicity is intermediate, the scattering is larger for a given effective mass than it would be for a covalent crystal, especially for low effective mass, thus even the very large mobilities for indium antimonide and indium arsenide (Table 2) are lower than they would be in a covalent crystal with electrons of similar effective mass. In the Group III-V compounds the atoms have an effective charge between 0·3 and 0·5 of the electronic charge due to the asymmetry of charge distribution between the atoms referred to earlier.

When impurities are present in the semiconductor, and the temperature is high enough to ionize them, there is an additional scattering process leading to a mobility μ_I which is proportional to $T^{3/2}$, and inversely proportional to the density of ionized impurity centres. Fig. 2 shows log μ plotted against T, and indicates that at low temperature the net value, given approximately by $1/\mu = 1/\mu_L + 1/\mu_I$, is determined by impurities, whereas at high temperature it is determined by lattice scattering. Curve A refers to a low impurity density, B to a higher one, both for acoustic mode lattice scattering. Scattering by uncharged impurities, stacking disorder and dislocations has an effect which is itself independent of temperature and which is in most cases only significant at low temperatures. In the absence of ionized impurities, neutral impurities or dislocations lead to the behaviour shown dotted at low temperatures instead of the increase characteristic of μ_I. It is clear that crystal imperfections and impurities must be avoided in order to obtain the mobility μ_L at the lower temperatures, and that with high defect densities the depression of μ below μ_L will be appreciable at room temperature. It is clear also that if in order to attain a required conductivity a

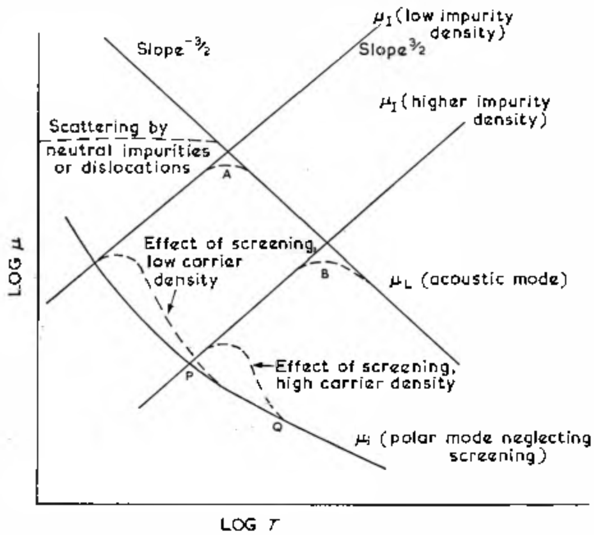

Fig. 2. Variation of mobility with temperature

relatively high value of n_0 is required, this is likely to lower μ below μ_L. Ultimately σ may not continue to rise as n_0 rises, because of the decrease in μ.

Exceptions to the general rule that mobility will fall as n_0 rises may occur when the lattice scattering is due to polar rather than acoustic mode vibrations. Then, as the carrier density rises, the carriers 'screen' the interaction between any one electron and the dipoles produced by the vibrations. When this screening effect is appreciable, the value of μ_L will increase as n_0 rises. This will be negligible at the higher temperatures, and will result in an increase in the net mobility for an intermediate temperature range, as shown between P and Q in Fig. 2.

With all 'normal' semiconductors, the mobility decreases as temperature rises in the high temperature range as in Fig. 2, whatever the details of the scattering mechanism. However, in the abnormal ones referred to at the end of the introduction, the electrons move by a 'hopping' process from one atom to another, so that energy is needed in order to free them. This results in an increase in mobility as temperature rises, though mobility at room temperature is usually very much less than in a normal semiconductor. It may be noted, however, that similar mechanisms may occur in normal semiconductors at low temperatures in the case of conductivity in an 'impurity band.'

Lifetime

If the electron and hole densities are perturbed from their thermal equilibrium values, the lifetime τ is the time-constant of the process of return to equilibrium after the stimulation ceases. This is important in photoconductivity and in pn junction devices where the operation of the device involves the injection across the junction of 'minority' carriers, i.e. of holes into an n-type semiconductor or vice versa. The device will not operate unless these minority 'live' long enough to reach another pn junction or a collecting electrode. The lifetime depends on the processes of recombination of electrons and holes, which are profoundly influenced by impurities, lattice defects and dislocations.

High Field Effects

It may be noted that the current does not necessarily increase linearly with voltage in a semiconductor. One departure from linearity arises if the electrons become 'hot', i.e., if they are accelerated by the field E until they are no longer in thermal equilibrium with the crystal lattice. When this occurs, above a critical field strength the mobility μ varies as $E^{-\frac{1}{2}}$, so that the current rises with field, but only as $E^{\frac{1}{2}}$. At still higher fields the current saturates. Another effect, which may be superposed on the first, is that electrons may acquire sufficient energy between collisions to free further electrons from the ions of the crystal when they collide with them. This leads to multiplication of the carrier density to a high value and a consequent large increase in current.

At low temperatures most of the electrons associated with impurities will be bound to the impurity atoms. The few free electrons can be accelerated in quite low fields until their energy is sufficient to ionize these impurities, so raising the free-carrier density near to its maximum extrinsic value n_0. Thus a considerable increase in current density can occur at quite low fields, e.g., at a few volts per centimetre in germanium at 4°K. Similar effects have been observed recently in indium antimonide and indium phosphide. New types of application are likely to result both from 'hot' electron processes and multiplication processes.

Classes of Semiconductor

The Elements

Germanium and silicon are the semiconductors most thoroughly investigated and most completely under control. Their properties are listed in Table I. Donor levels in germanium, at a depth of the order 0·01eV, are fully ionized above 20°K. In silicon they are deeper and are fully ionized above about 90°K. The mobility in Ge is high enough to achieve transistor operation at high frequencies, up to 500Mc/s or more in pn junction devices with base widths of about 2μ. Lifetimes in Ge as high as 1msec can be obtained, though these are not necessary for high frequency applications. In silicon the mobility is lower, and for a given geometry the upper frequency limit is below that of germanium. If transistor behaviour is desired above 200°C, a higher gap is needed than that of silicon; for high frequencies this should ideally be combined with a mobility at the operating temperature comparable with that of germanium at room temperatures. This would however call for a material with a very high mobility at room temperature, in view of the fall of μ_L as T rises, discussed under 'Mobility'.

Silicon Carbide

There is a considerable electronegativity difference between silicon and carbon, so that the compound between them has a considerable degree of ionic bonding. The effective charge on the ions is about 0·9e. This leads to a mobility less than that resulting from covalent-bond lattice-

TABLE 1

	ATOMIC WEIGHT M	MELTING POINT (°C)	ENERGY GAP Q (eV)	DEBYE TEMPERATURE θ (°K)	DIELECTRIC CONSTANT ε	MOBILITY ELECTRONS μ_e ($cm^2\ V^{-1}\ sec^{-1}$)	MOBILITY HOLES μ_h ($cm^2\ V^{-1}\ sec^{-1}$)	RATIO OF EFFECTIVE MASS TO FREE ELECTRON MASS ELECTRONS m_e	RATIO OF EFFECTIVE MASS TO FREE ELECTRON MASS HOLES m_h	THERMAL CONDUCTIVITY K_0 (W/cm °C)
Diamond	12·0	3500	5·2	1950		1800	1200	—	—	1·95
Si	28·1	1420	1·1	652	11·7	1300	500	0·26	0·4	1·13
Ge	72·6	940	0·65	362	15·7	3800	1800	0·12	0·2	0·63
Sn	118·7	—	0·08	215	—	2000	1000	—	—	—
SiC	20·0	—	2·8	—	L.F. 10, H.F. 6·7	~100	~20	0·6	1·2	—

TABLE 2
III-V Compounds

	ATOMIC WEIGHT M	MELTING POINT (°C)	ENERGY GAP Q (eV)	DEBYE TEMPERATURE θ (°K)	DIELECTRIC CONSTANT ε	MOBILITY ELECTRONS μ_e ($cm^2\ V^{-1}\ sec^{-1}$)	MOBILITY HOLES μ_h ($cm^2\ V^{-1}\ sec^{-1}$)	RATIO OF EFFECTIVE MASS TO FREE ELECTRON MASS ELECTRONS m_e	RATIO OF EFFECTIVE MASS TO FREE ELECTRON MASS HOLES m_h	THERMAL CONDUCTIVITY K_0 (W/cm °C)
BN	12·5	—	~5·0	—	—	—	—	—	—	—
BP	21·0	—	—	—	—	—	—	—	—	—
GaN	42·0	—	3·3	—	—	—	—	—	—	—
AlN	20·5	—	—	—	—	—	—	—	—	—
AlP	29·0	—	3·0	—	—	—	—	—	—	—
AlAs	51·0	—	2·2	—	—	—	—	—	—	—
AlSb	74·0	1060	1·65	—	8·4	1200	150	—	0·4	—
GaP	50·5	—	2·3	—	—	—	—	—	—	—
GaAs	72·5	1240	1·3	356	—	5000	200	0·04	—	—
GaSb	96·0	700	0·8	—	14·0	4000	1000	0·05	—	0·27
InP	73·0	1070	1·2	—	—	4000	200	0·04	—	—
InAs	95·0	940	0·45	—	—	30000	200	0·03	~0·3	0·3
InSb	118·0	523	0·2	205	16·0	80000	800	0·015	0·18	0·15

TABLE 3
II-VI Compounds

	ATOMIC WEIGHT M	MELTING POINT (°C)	ENERGY GAP Q (eV)	DEBYE TEMPERATURE θ (°K)	DIELECTRIC CONSTANT ε L.F.	DIELECTRIC CONSTANT ε H.F.	MOBILITY ELECTRONS μ_e ($cm^2\ V^{-1}\ sec^{-1}$)	MOBILITY HOLES μ_h ($cm^2\ V^{-1}\ sec^{-1}$)	RATIO OF EFFECTIVE MASS TO FREE ELECTRON MASS ELECTRONS m_e	RATIO OF EFFECTIVE MASS TO FREE ELECTRON MASS HOLES m_h	THERMAL CONDUCTIVITY K_0 (W/cm°C)
ZnO	41·0	—	3·2	—	8·5	4·0	1000	—	0·07	—	—
CdO	64·0	—	2·2	415	—	5·8	100	—	—	—	—
ZnS	44·0	1850	3·6	—	—	—	—	—	—	—	—
ZnSe	72·0	1000	2·6	—	—	—	—	—	—	—	—
ZnTe	96·5	1240	2·1	—	—	—	—	—	0·2	—	—
CdS	72·0	1475	2·4	—	—	6·2	200	20	0·36	—	—
CdSe	95·2	1250	1·7	—	—	—	200	—	—	—	—
CdTe	120·5	1040	1·4	—	—	10·4	600	50	0·34	—	—
HgS	116·0	583	—	—	—	—	—	—	—	—	—
HgSe	140·0	690	~0·2	—	—	—	10000	—	—	—	—
HgTe	164·0	670	0·02	—	—	—	17000	—	—	—	0·02

TABLE 4
IV-VI Compounds

	ATOMIC WEIGHT M	MELTING POINT (°C)	ENERGY GAP Q (eV)	DEBYE TEMPERATURE θ (°K)	DIELECTRIC CONSTANT ε L.F.	DIELECTRIC CONSTANT ε H.F.	MOBILITY ELECTRONS μ_e ($cm^2\ V^{-1}\ sec^{-1}$)	MOBILITY HOLES μ_h ($cm^2\ V^{-1}\ sec^{-1}$)	RATIO OF EFFECTIVE MASS TO FREE ELECTRON MASS ELECTRONS m_e	RATIO OF EFFECTIVE MASS TO FREE ELECTRON MASS HOLES m_h	THERMAL CONDUCTIVITY K_0 (W/cm°C)
PbS	120·0	1110	0·37	194	70	17·4	600	400	0·15	—	—
PbSe	143·0	1065	0·26	—	—	20·7	900	500	—	—	—
PbTe	168·0	904	0·29	—	—	30·0	300	300	0·3	—	0·023

scattering, and much less than that in diamond or silicon, as shown in Table 1. There is in addition great difficulty in achieving crystal perfection because of the 'stacking-disorder' typical of this material. It is possible that more perfect crystals would give higher mobility than that listed, though still below that of silicon. Such low values are not very useful for transistors, but are acceptable for rectifiers; silicon carbide rectifiers can be operated above 500°C.

Group III-V Compounds

These compounds have a diamond-type crystal structure similar to that of germanium and silicon, and have partly covalent, partly ionic binding. Their electrical properties are listed in Table 2 and it will be noted that gallium arsenide and indium phosphide are suitable as regards gap-width and mobility for high-temperature high-frequency transistor operation. The life-time obtained so far is, however, inadequate for good transistor performance. High lifetime is not required for high-speed switching, or in variable reactance diodes for parametric amplifiers. For the latter their high mobility and relatively high gap-width make gallium arsenide and indium phosphide of particular interest. The mobility is high also in indium arsenide and especially in indium antimonide, but these have low gap-width. They are important for Hall effect applications and as photo-conductors sensitive in the infra-red.

In both gallium arsenide and indium phosphide, the mobilities refer to material with an impurity density not much below $10^{17}/cm^3$. This represents the highest purity obtainable so far. The mobilities as listed are quite high, but are no doubt depressed by impurity scattering. If these compounds can be further purified, electron mobilities near 10 000 are probably attainable in both. The impurities causing low lifetime must have deep seated energy levels at which recombination occurs, as distinct from shallow donor or acceptor levels. This suggests that several types of impurity remain present in the best samples of these compounds made so far.

Group II-VI Compounds

Compounds between Group II and Group VI elements have either hexagonal or cubic structure. Their properties are indicated in Table 3. They are more ionic and have larger gap-width than group III-V compounds of similar atomic weight. Zinc and cadmium sulphides are insulators in the pure state, and are of interest mainly as phosphors and photo-conductors. Cadmium sulphide with suitable impurity can, however, have a conductivity exceeding $1\Omega^{-1}cm^{-1}$. Donor levels are obtained by substituting either Group III atoms for the Group II or Group VII for the Group VI. Correspondingly acceptor levels are obtained by substituting Group I atoms for the Group II or Group V for the Group VI. The mobility is low for the compounds of low atomic weight, but large for electrons in mercury telluride and selenide, where there is less ionicity in the bond. and the effective masses are low. These, however, have low gap-widths. Cadmium sulphide and selenide are good photo-conductors, sensitive to visible light, in spite of the rather low mobility. This is because there is a large increase in electron population for a small light input, and a large lifetime for these excess electrons. This appears to arise from the details of the recombination and trapping processes characteristic of these crystals. Unfortunately the long lifetime associated with the high sensitivity results in a speed of response too slow for some applications. (~0·1sec for CdS and ~0·01sec for CdSe.) Research on these compounds is very active at the present time, and further improvements in performance can be expected.

Both CdS and CdTe can also give photovoltaic effects, with a high response to red light, so that they are potentially valuable for solar battery applications. pn junctions can now be prepared in both, but so far they rarely have the characteristics expected on the basis of energy gap and resistivity.

Group I-VII Compounds

These compounds are still more ionic than the group II-VI elements, and have high gaps and low mobilities. The alkali-halides are of no interest as semiconductors, and in fact have ionic conductivity at temperatures below those at which intrinsic electronic conductivity appears. Because of low dielectric constant, the ionization energy of impurity centres is high, so that extrinsic conductivity is very small. With the silver halides photo-conductivity plays a significant part in the photographic process, but otherwise semiconductivity is not important in them.

Group IV-VI Compounds

Lead sulphide, telluride and selenide have a rock salt structure indicating that the lead in these compounds is effectively divalent. Their properties are shown in Table 4. The interest in these until recently has mainly been in the photo-conducting properties in the infra-red, but recently lead telluride and selenide have also become important as thermoelectric materials, as discussed further below.

Other Binary Compounds

Various other families of binary compounds have been studied to a limited extent, with results as in Tables 5, 6 and 7. With all these materials, real understanding will be obtained only when single crystals of stoichiometric material have been prepared with uniform properties, when the mobility has been measured over a wide range of carrier density, and when at least some studies such as cyclotron resonance have been carried out giving information about the effective mass. This represents a considerable programme, the initial work in preparing and controlling good-quality single crystals being difficult in many cases, especially when constituents are volatile at or below the melting point. Some of these compounds are finding applications because of their thermoelectric properties, as discussed below.

Ternary Compounds

The preparation, control of purity and stoichiometry, and growing of single crystals becomes more difficult in ternary systems. Most study has been done so far on Group I-III-VI compounds (chalcopyrites) which are analogues of the compounds between Group II and Group VI elements, on Group II-IV-V compounds which are analogues of the Group III-V compounds, and on Group I-V-VI compounds. Their properties are listed in Tables 8, 9 and 10.

TABLE 5
II-IV Compounds

	ATOMIC WEIGHT M	MELTING POINT (°C)	ENERGY GAP Q (eV)	MOBILITY ELECTRONS μ_e (cm² V⁻¹ sec⁻¹)	HOLES μ_h
Mg_2Si	26	1100	0·77	400	60
Mg_2Ge	41	1150	0·74	280	110
Mg_2Sn	56	770	0·36	300	250
Mg_2Pb	85	550	? Metallic	—	—
Ca_2Si	36	920	1·9	—	—
Ca_2Sn	66	1120	0·9	—	—
Ca_2Pb	96	1110	~0·5	—	—

TABLE 6
V-VI Compounds

	ATOMIC WEIGHT M	MELTING POINT (°C)	ENERGY GAP Q (eV)	DEBYE TEMPERATURE θ (°K)	DIELECTRIC CONSTANT ε	MOBILITY ELECTRONS μ_e (cm² V⁻¹ sec⁻¹)	MOBILITY HOLES μ_h (cm² V⁻¹ sec⁻¹)	THERMAL CONDUCTIVITY K_0 (W/cm °C)
Bi_2Te_3	160	575	0·15	145	85	420	400	0·016
Bi_2Se_3	131	706	0·27	—	—	~600	—	—
Sb_2Te_3	125	620	? metallic	—	—	—	~300	0·05
Sb_2Se_3	96	617	1·2	—	—	~100	—	—
Bi_2S_3	103	—	1·3	—	—	200	—	—
Sb_2S_3	68	—	1·7	—	—	—	—	—
As_2Te_3	107	—	~1·0	—	—	170	80	0·02
As_2Se_3	77	—	~1·5	—	—	—	—	—

TABLE 7
III-VI Compounds

	MELTING POINT (°C)	ENERGY GAP Q (eV)	MOBILITY ELECTRONS μ_e (cm² V⁻¹ sec⁻¹)	MOBILITY HOLES μ_h (cm² V⁻¹ sec⁻¹)
Ga_2Te_3	793	1·3	—	—
In_2Te_3	667	1·0	—	—
GaSe	—	2·0	40	20
GaTe	—	1·7	—	—
InSe	—	1·8	—	—
TlSe	—	0·6	—	—

TABLE 8
I-III-VI Compounds

	MELTING POINT (°C)	ENERGY GAP Q (eV)	MOBILITY ELECTRONS μ_e (cm² V⁻¹ sec⁻¹)	MOBILITY HOLES μ_h (cm² V⁻¹ sec⁻¹)	THERMAL CONDUCTIVITY K_0 (W/cm°C)
$AgInS_2$	—	1·9	—	—	—
$CuInS_2$	—	1·2	—	—	—
$AgInSe_2$	773	1·2	—	—	0·03
$CuInSe_2$	990	0·9	1010	50	—
$AgInTe_2$	680	0·95	—	100	—
$CuInTe_2$	780	0·95	—	—	—
$CuFeS_2$	875	0·5	—	—	—
$CuFeSe_2$	574	0·15	—	<20	—
$CuFeTe_2$	740	0·10	—	<50	—
$CuGaSe_2$	1040	1·6	—	—	—
$CuGaTe_2$	870	1·0	—	60	—
$CuTlSe_2$	405	—	—	—	—
$CuTlTe_2$	375	—	—	—	—
$AgGaSe_2$	850	1·7	—	—	—
$AgGaTe_2$	720	1·1	—	—	—
$AgTlSe_2$	328	0·7	—	—	—
$AgTlTe_2$	290	—	—	—	—
$AgFeSe_2$	737	0·23	>250	70	0·03
$AgFeTe_2$	680	0·3	>2000	150	0·03

TABLE 9
II-IV-VI Compounds

	ENERGY GAP Q (eV)		ENERGY GAP Q (eV)
$ZnSiAs_2$	2·1	$CdSnAs_2$	—
$ZnGeP_2$	2·2	$ZnSnP_2$	2·1
$CdGeP_2$	1·8	$CdSnP_2$	1·5
$ZnGeAs_2$	~1·0	$BeSiN_2$	>3·0
$ZnSnAs_2$	—	—	—

TABLE 10
I-V-VI Compounds

	MELTING POINT (°C)	ENERGY GAP Q (eV)	MOBILITY ELECTRONS μ_e (cm² V⁻¹ sec⁻¹)	MOBILITY HOLES μ_h (cm² V⁻¹ sec⁻¹)
$CuSbSe_2$	452	—	—	—
$AgSbSe_2$	636	~0·7	—	1500
$AgSbTe_2$	556	~0·6	—	75
$AgBiSe_2$	762	—	—	—

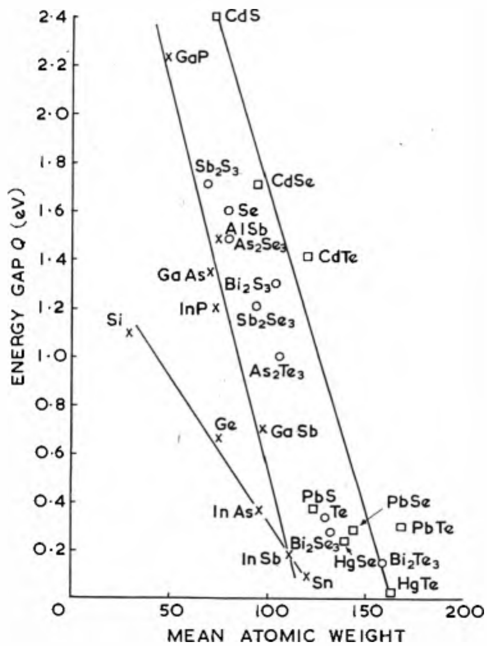

Fig. 3. Variation of energy gap with atomic weight

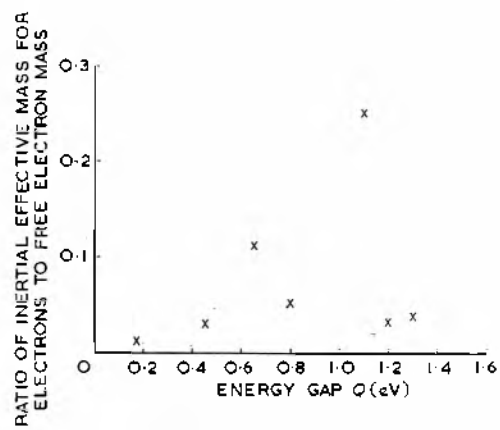

Fig. 4. Variation of effective mass with energy gap

General Relations

It will be noted that in each class of compound the energy gap Q falls as atomic weight M rises, Fig. 3. This is consistent with the decrease in bond-energy in a covalent bond as M rises. Variations in electronegativity-difference between the atoms lead to variations in the details of the relationship between Q and M, superposed on this general trend. According to the simplest theory of energy bands, the effective mass is expected to vary directly with Q, but in fact this is not generally the case, as shown in Fig. 4.

This would probably apply if the variation in Q arose directly from the relative motions of one valence band and one conduction band as M varied. However, the band structure is complex, and the conduction band in particular in the diamond-structure crystals has several components. As M varies, the identity of the component in which the minimum energy occurs may also vary. When this happens there will be a switch from one relationship between effective mass and gap-width to a different one, since this relationship is not the same for the different components. Correspondingly, according to the simple theory, μ should rise systematically as Q falls, and therefore M rises. Again this is not universally so, as shown in Fig. 5. It is true however that the highest mobilities are encountered in low-gap semiconductors, and that mobility rises as Q falls and M rises in each class of material. There are however other important factors, resulting in different values of μ for similar gap-widths, comparing different classes of compounds. These differences arise from the variation in the details of the bonding and crystal structure, and from the variation in electro-negativity of the ions in the compounds in a given type of bond.

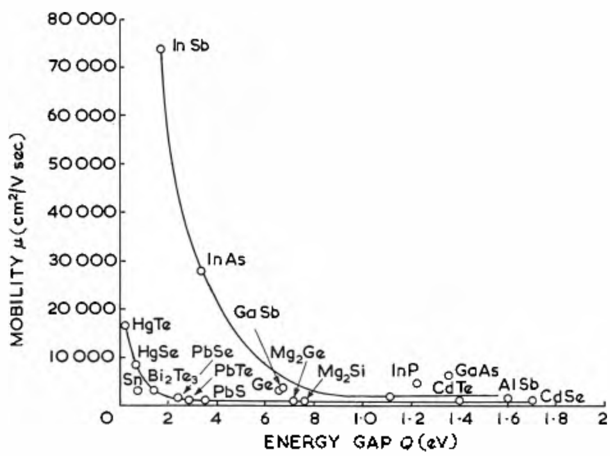

Fig. 5. Variation of mobility with energy gap

Thermoelectric Applications

Considerable interest has developed recently in the possibility of thermoelectric refrigeration* and generation of power. This subject requires separate mention, as considerations arise other than mobility and lifetime. For either application it is possible to define a figure of merit $z = \alpha^2\sigma/K$, where α is the thermoelectric power and K the thermal conductivity. These three quantities, α, σ and K are inter-related in a semiconductor, and it can be shown that z has a maximum value for any material provided it is 'doped' to give a thermoelectric power near $200\mu V/°C$. The thermojunction will be formed between n- and p-type material, so that the e.m.f. produced by the junction will be $400\mu V/°C$. Provided this is achieved, in the case of lattice scattering in a semiconductor with covalent bonds, $z \propto N/m_i K_o$. Here N is the number of 'valleys' in the band-structure; N is 4 for electrons in germanium, 6 for electrons in silicon and unity for electrons in InSb. m_i is the 'inertial' effective mass which determines the mobility. K_o is the component of thermal conductivity due to the lattice vibrations, as distinct from the component K_e due to electrons. The total thermal conductivity $K = K_o + K_e = K_o + A\ (k/e)^2\sigma T$. K_o is found to decrease as atomic weight M increases in any class of material, as shown in Fig. 6. It will be noted that the alkali halides have lower values of K_o than covalent-bonded materials of similar atomic weight. However, they are of no value for thermoelectric applications because of low μ and low dielectric constant, as discussed above. The value of N cannot at present be predicted for any semiconductor, and does not appear to vary in any systematic way with atomic weight.

Since in any family of semiconductors m_i and K_o both fall as atomic weight rises, for thermoelectric applications attention has been concentrated on compounds with high values of M, and the best thermoelectric materials known so far are bismuth telluride, Bi_2Te_3 and lead telluride, PbTe, and more complex materials made from these by substituting elements from the same columns of the periodic table, e.g., Sb for Bi, or Se for Te. Such substitution lowers the value of K_o, in many cases without lowering the mobility. In this type of application the electrical conductivity σ is required to be relatively high (the particular value desired is that which matches a thermoelectric power of $200\mu V/°C$), and relatively high carrier densities are necessary, of the order $10^{18}/cm^3$. Thus the extreme conditions of purity and crystal perfection required for transistor operation are not necessary for thermoelectric applications; while μ should be as high as possible, one has in fact to be content with the value characteristic of a heavily doped material, which as shown earlier is likely to be less than that which can be obtained in the pure material (unless indeed screening effects reverse this trend). Further imperfection in the crystal due to substitution of an atom of different size is likely to have little further effect in lowering μ. Moreover, this situation means that ternary compounds such as those listed in Tables 8, 9 and 10 are possibly of interest if K_o is low; it will be even more difficult to produce these in good quality crystalline form, stoichiometric and free from impurity, than is the case with binary compounds. This does not necessarily disqualify them however for thermoelectric applications, as it would for transistors. Research on these compounds is very active in many laboratories at present in case some of them prove superior to those based on bismuth telluride.

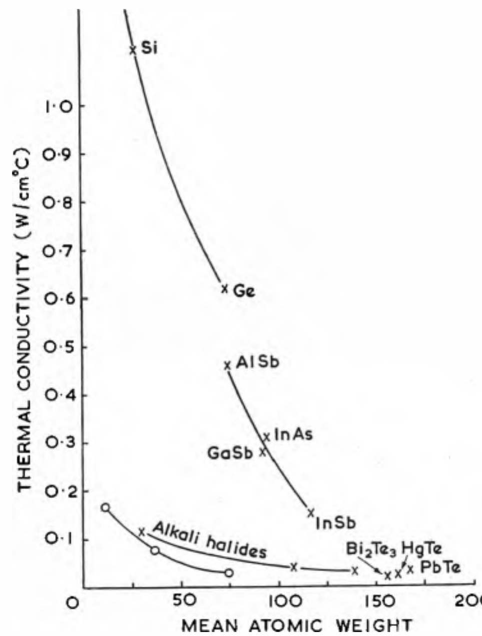

Fig. 6. Variation of thermal conductivity with atomic weight

* *See page 690, this issue.*

V.S.W.R. Indicators with Automatic Read-Out

By M. Kollanyi*, M.Sc., A.M.Brit.I.R.E., and R. M. Verran*, B.Sc.

An automatic indicator giving a direct read-out either on a meter or a digital display unit can replace the conventional v.s.w.r. indicator with its time taking adjustments. The circuits can be used with any slotted line or equivalent device. A single movement of the probe carriage is all that is necessary for the instrument to supply the correct v.s.w.r. reading.

The system described is based on a maximum-minimum detector that detects and stores the extreme values of the voltages encountered during the recording cycle. The ratio of these voltages can be calculated by various electronic methods which are also described.

A residual v.s.w.r. of <1·05 is attainable with a maximum range of v.s.w.r. measurement of 1 : 3 or more if required, although this range depends on the method chosen.

(Voir page 707 pour le résumé en français: Zusammenfassung in deutscher Sprache Seite 712)

MOST readers will be familiar with the tedious and time consuming nature of v.s.w.r. measurements made with a slotted line. First, the cavity of the slotted line must be tuned to the signal source, then the position of an antinode must be found and the gain of an amplifier used with the system adjusted precisely, and finally the position of a node must be found and the ratio of the two levels determined.

This article describes some of the methods devised and circuits constructed by the authors in an attempt to reduce this labour to a minimum. However, before proceeding with the description of these circuits, it is useful and instructive to consider some of the other methods for automating v.s.w.r. measurement that are at present in use.

A system which causes the probe carriage of the slotted line to be mechanically reciprocated has obvious advantages, but does not really get to the heart of the problem as the maximum and minimum values of the probe output must still be found and their ratio determined as before.

All that has been saved is the movement of the probe carriage to find the positions of a node and an antinode. Usually with this arrangement the probe output is displayed on an oscilloscope, and in this case the gain of an amplifier must still be adjusted to set a level to a predetermined mark.

A better approach to the problem is to feed the output from the probe into an amplifier with an extremely high degree of automatic gain control. This sets the maximum output from the amplifier to a fixed level, and to determine the v.s.w.r. it is sufficient to measure the minimum value. If this system is used in conjunction with a reciprocating probe carriage, then it is possible to obtain a persistent reading on a meter or other indicating device.

Other methods are based on the directional properties of various devices, such as the magic-T. With the magic-T, if arm 3 is connected to a matched generator and the impedance to be measured is connected to arm 4, then the v.s.w.r. can be computed from the voltages existing at arms 1 and 2.

The directional coupler, as its name implies, is capable of discriminating between two directions of propagation possible on a line, so that the forward and reflected components can be separated and their amplitudes measured separately.

The multiple probe system also has its shortcomings although it provides a conveniently readable display. It is difficult to maintain the setting of the probes and to keep the sensitivity of the detectors constant at various frequencies.

* *Marconi Instruments Ltd.*

The disadvantage of most of these systems is that they are not applicable to existing slotted lines, and in some cases they require a complex mechanical arrangement for their functioning. With the directional coupler and magic-T type of arrangement there is the alternative disadvantage that the phase information about the load is suppressed,

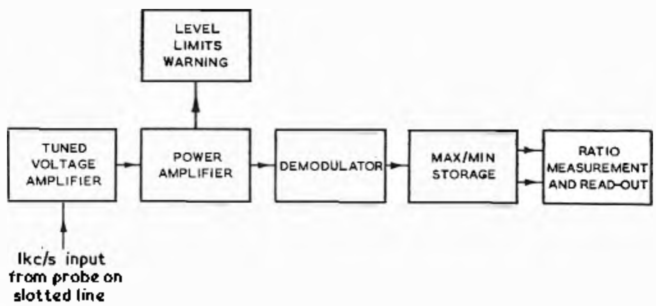

Fig. 1. The direct reading v.s.w.r. indicator

although this information is often as important as the information about the magnitude of the v.s.w.r.

The system to be described does not suffer from these disadvantages. It is readily applicable to any existing conventional slotted line, and the only modification that is necessary is to mount a switch on the probe carriage. One movement of the carriage backward and forward along the slotted line with this switch is sufficient for the instrument to compute the v.s.w.r. existing on the line and display the answer as a stationary reading on a meter. The instrument is also capable of functioning as a conventional v.s.w.r. indicator if for some reason this should be desired.

It is claimed that in many cases a more accurate answer is obtainable with this system than by conventional means, for the time taken to obtain a reading is so much reduced that there is correspondingly less time for the signal frequency or amplitude to drift.

Discussion

The system developed by the authors (Fig. 1) is based on a circuit which extracts and stores the extreme values of the voltages encountered when moving the probe along the slotted line. To be able to do this, it is first necessary to detect the envelope of the microwave signal from the probe and amplify it with a selective amplifier in the conventional way.

The envelope of this audio frequency signal is, in turn, obtained from a demodulator stage the time-constant of which is chosen so that its output signal can follow the amplitude of the microwave signal along the standing wave pattern. To be able to do this without significant error, the time-constant of the demodulator circuit must be less than 0·02sec. This requirement is based on the assumption that the speed of the probe carriage does not exceed half a wave length per second.

The output of this demodulator stage is used to operate the circuit which extracts and stores the extreme values. This circuit, shown in Fig. 2, consists fundamentally of rectifiers and capacitors arranged in such a way that during the recording process the signal from the demodulator charges C_1 through MR_1 to its peak value, while C_2 discharges through MR_2 into the demodulator to the lowest

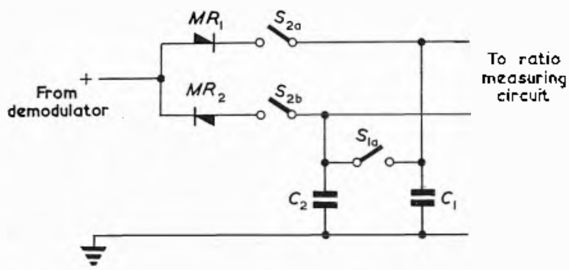

Fig. 2. Maximum and minimum storage circuit

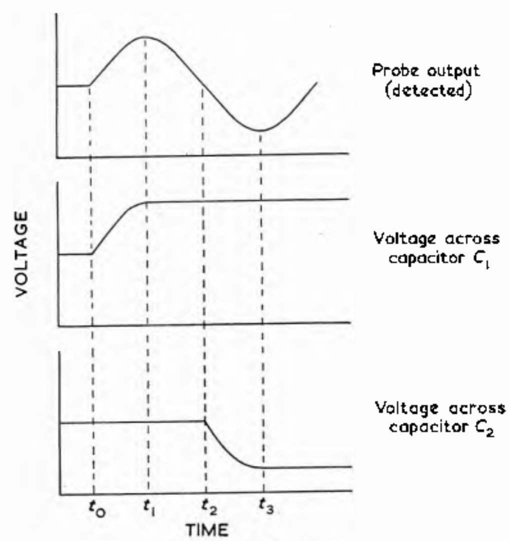

Fig. 3. Waveforms

voltage produced. To carry out the recording process the operator has merely to move the probe carriage once along the slotted line.

The purpose of S_1 is to provide an initial charge on C_2 which must always be greater than the expected minimum. This initial charge is supplied through MR_1, and consequently at the beginning of the recording period C_1 and C_2 will have equal voltages (Fig. 3). As soon as the switch is released (t_0) and the carriage is moved the voltages on the two capacitors will vary independently.

In Fig. 3, it has been assumed that the voltage change begins by increasing; consequently the voltage on C_1 increases also, while that on C_2 remains unchanged. At t_1 the voltage reaches its peak value, and from this time the voltage on C_1 will not change. Moving the probe further the potential falls below the initial value at t_2, and at this point MR_2 starts conducting and will discharge C_2 until the voltage across this capacitor reaches the minimum value at t_3, where it will remain.

Due to the square law of the microwave detector, this process gives two voltages proportional to the square of the extreme values on the line. The result is not affected by the speed of the carriage as long as the time-constants of the circuit are negligible. The two voltages are stored in C_1 and C_2 for further processing by the ratio measuring circuit.

Some difficulties were encountered in designing the demodulator, mainly owing to the conflicting nature of the requirements placed upon it. Not only must it be linear within wide voltage limits, but a low time-constant and a low amount of ripple are imperative. The values actually used ($C = 1\mu$F, $R = 33$kΩ) appear to be a good compromise between low ripple and low time-constant.

The wide voltage limits are important in order to avoid the necessity of adjusting the gain of the selective amplifier too often. A usable output ratio of 1:40 from the demodulator is desirable in order to cover the voltage standing wave ratio change $\sigma = 2{\cdot}5$ twice.

At first, germanium diodes were used in the demodulator, but although they had a relatively good performance at low levels, their output became non-linear at higher volt-

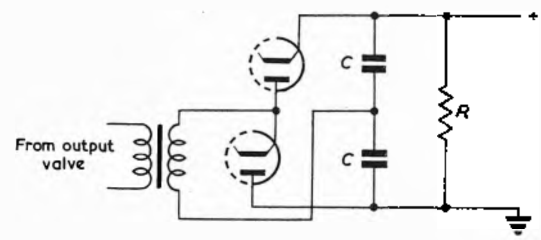

Fig. 4 Demodulator circuit

ages. Since it was impossible to reach the 120V upper limit aimed at they were discarded and replaced by thermionic diodes.

The diodes are connected in the form of a voltage doubler circuit, as shown in Fig. 4. The output from this demodulator is in perfectly linear relationship with the input voltage except at very low voltages, where the linearity is impaired by the 'splash' current of the thermionic diodes. This maintains a constant voltage at the output in the absence of any input and imposes a lower limit on the output voltage for a given limit of error.

To determine the magnitude of this lower limit, let the case be considered when the maximum signal voltage is a, the minimum signal voltage is b, and the splash voltage is e. In the absence of splash voltage the standing wave ratio (σ) measured would be

$$\sigma = \sqrt{(a/b)} \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (1)$$

(taking into account the square law of the first detector) With the splash voltage, however

$$\sigma' = \sqrt{\left(\frac{a+e}{b+e}\right)} = \sqrt{(a/b)}\sqrt{\frac{1+(e/a)}{1+(e/b)}} \quad \ldots\ldots \quad (2)$$

Ignoring second order terms

$$\sigma' = \sigma(1 + (e/2a))(1 - (e/2b)) \quad \ldots\ldots\ldots \quad (3)$$

and

$$\sigma' = \sigma(1 + (e/2a) - (e/2b)) \quad \ldots\ldots\ldots \quad (4)$$

Now the proportional error $P = \dfrac{\sigma - \sigma'}{\sigma}$ .... (5)

or, substituting,

$$P = (e/2)\left(\frac{a-b}{ab}\right) \ldots\ldots\ldots\ldots\ldots \quad (6)$$

Since $a = b\,\sigma^2$ this may be expressed as

$$P = (e/2b)\,\frac{\sigma^2 - 1}{\sigma^2} \ldots\ldots\ldots\ldots\ldots \quad (7)$$

If the maximum permissible error is fixed at 2 per cent, and $\sigma_{max} = 2{\cdot}5$ then

$$b_{min} = \varepsilon\sqrt{2P}\,\frac{\sigma^2 - 1}{\sigma^4} = 21\varepsilon \text{ volts.}$$

In the actual circuit the splash voltage was approximately 0·5V, setting the limit for the lowest voltage to approximately 10·5V.

The maximum voltage available from the demodulator is 150V, so the r.f. voltage ratio range actually covered is $\sqrt{(150/10{\cdot}5)} = 3{\cdot}8$. This gives a sufficient margin around v.s.w.r. $\sigma = 2{\cdot}5$ to make frequent gain control adjustments unnecessary.

It is advisable to incorporate a warning system that gives an indication to the operator when the voltage limits to be observed are exceeded. This can be operated from the power stage driving the final demodulator.

So far voltages proportional respectively to the square of the maximum and minimum voltages existing on the slotted line have been stored in two capacitors. During the read-out period these two voltages are processed in

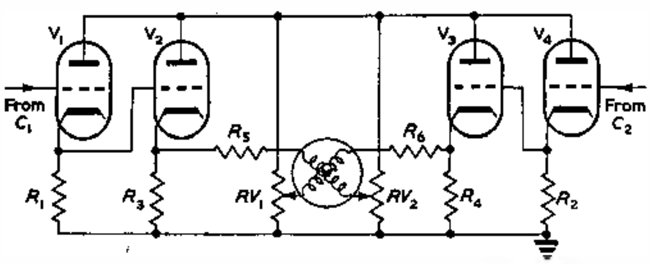

Fig. 5. Ratio measuring circuit using ratiometer

order to display their ratio either directly or in a logarithmic form.

It would appear that the most straightforward method would be to use a crossed coil ratiometer, which could be connected to the two capacitors via cathode-followers in order to obtain a permanent reading. However, some care is necessary in order to maintain acceptable accuracy. The output voltage of a cathode-follower is not zero for zero input voltage, and so some form of backing off device must be used. Furthermore, the output voltage must remain linear with the input voltage even when a current of several milliamperes is being supplied to the ratiometer coil.

These considerations have led to the final circuit, shown in Fig. 5. A double cathode-follower is used so that a high input impedance and low output impedance can easily be obtained. In the interests of simplicity the backing off device chosen was a potentiometer rather than the more elegant method of using a similar cathode-follower.

It might be thought that the limitations imposed on the voltage range over which the instrument operates as a ratiometer would limit its usefulness, but this was not found to be a problem as these limits are far wider than those of other parts of the system such as the demodulator and power output stages. A more serious objection is that once the ratiometer itself is designed nothing can be done to change the scale shape and it is difficult to modify simply the range over which the instrument will compute ratios.

Some other methods, relying almost exclusively on the exponential discharge of capacitors, were tried in order to overcome the limitations of the ratiometer. One of these systems measures the time necessary to equalize the voltages on the two capacitors, and is shown in Fig. 6.

Consider the two capacitors C_1 and C_2 charged to V_1 and V_2 volts respectively, V_1 being the greater.

If C_1 is discharged through a resistor R_1 to V_2, then the time required for this will be t according to the formula

$$V_2 = V_1 \exp(-t/R_1C_1) \quad \ldots\ldots\ldots\ldots \quad (8)$$

Since

$$\sqrt{(V_1/V_2)} = \sigma$$

Then

$$t = 2R_1C_1 \ln \sigma \text{ or } t = K \ln \sigma \quad \ldots\ldots\ldots \quad (9)$$

Thus it is seen that the time required to equalize the two voltages is directly proportional to the logarithm of the v.s.w.r.

The operation of the system shown in Fig. 6 is simply to discharge C_1 through a resistor until no difference can be detected between its voltage and that of C_2. The zero detector is isolated from the storage capacitors by two cathode-followers, and a chopper samples the voltage on the cathode of this valve. The output from the chopper

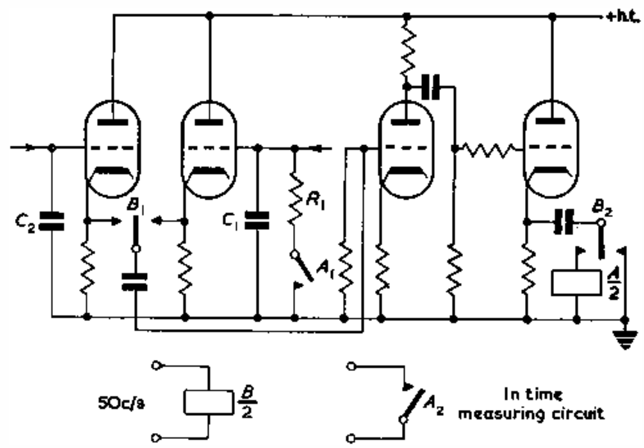

Fig. 6. Voltage equalizing circuit

is, amplified then rectified by a second chopper and used to operate the relay which discharges C_1.

The accuracy with which the zero may be detected by this method depends greatly on the mechanical chopper used and the gain of the amplifier following it. In the circuit built the zero was detected with an inaccuracy of only a few millivolts, which gave a very low error in the indicated v.s.w.r.

The time measuring system used is shown in Fig. 7. An initially charged capacitor is discharged through a resistor for the same time as is necessary to equalize the maximum and minimum voltages, and the change in voltage is measured by a valve-voltmeter circuit incorporating a differential system to compensate for the effects of leakages.

The range of v.s.w.r. measurement may be altered by adjusting the potentiometer RV and so altering the initial voltage to which the capacitor is charged, at the same time adjusting the compensating voltage.

A different system, also based on the exponential discharge of the storage capacitors but which obviates the need for the chopper is shown in Fig. 8. In this system the two capacitors C_1 and C_2 are connected to separate discharging resistors after the recording period.

The voltages across these two resistors are used to control two Schmitt trigger circuits, so that the ratio of the two initially stored voltages is converted into a time difference that can be measured by some method such as the one previously described.

The top end of the discharging resistor is connected to the grid of V_{1a} (Fig. 9), and the various circuit potentials are so chosen that when the voltage across this resistance is zero, V_{1a} is non-conducting.

When the resistance is connected to its associated storage capacitor after the recording period, the increased voltage on V_{1a} grid makes it conduct, while V_{1b} ceases to conduct, and the circuit remains in this condition until the voltage on the capacitor drops below a predetermined level, when V_{1b} again conducts and V_{1a} becomes non-conducting. A negative pulse is derived from V_{1b} anode when this happens, and this pulse together with the similar pulse from the other trigger circuit operates the time measuring circuit.

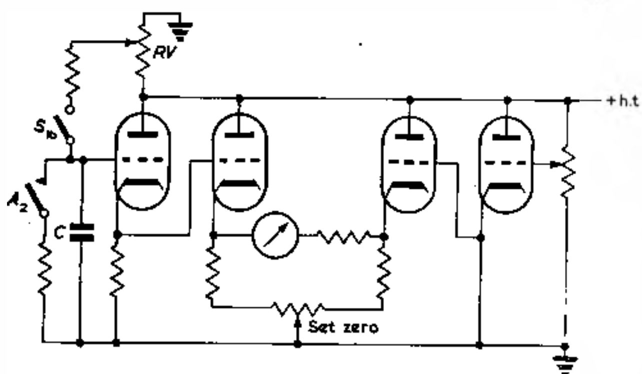

Fig. 7. Time measuring circuit

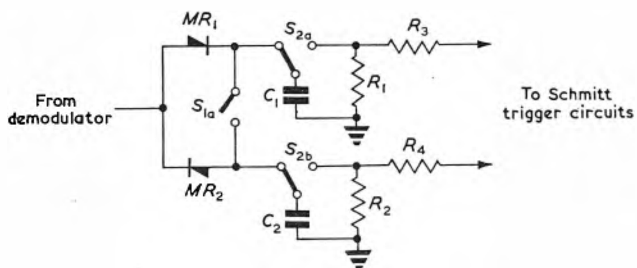

Fig. 8. Simultaneous discharge system

The clamping valve V_{2a} reduces the voltage swing on the cathode of V_1 and so reduces the hysteresis in the operation of the trigger circuit.

To consider the quantitative relationships involved in this method, suppose that two capacitors C_1 and C_2, charged respectively to V_1 and V_2 volts are discharged through resistors R_1 and R_2 to the same voltage v. The times t_1 and t_2 for the voltages to fall to this value will be given by

$$v = V\, e^{-t/RC}$$

or

$$t = RC\,(\ln V - \ln v) \qquad (10)$$

If now $R_1C_1 = R_2C_2 = T$

Then

$$t_1 - t_2 = T\,(\ln V_2 - \ln V_1) \qquad (11)$$

or

$$t_1 - t_2 = T\,\ln(V_2/V_1) = \tau \qquad (12)$$

This methods lends itself to the production of a read-out in digital form by using the pulses derived to start and stop an electronic counter working with a fixed frequency oscillator. By suitable choice of the discharge time-constants and the oscillator frequency, the counter may be made to indicate the ratio of the two voltages (the v.s.w.r.) directly in decibels.

If V_1 is the maximum stored voltage and V_2 the minimum one, then v.s.w.r. on the line is $\sqrt{(V_1/V_2)}$.

Expressed in decibels, this is

$$20 \log \sqrt{(V_1/V_2)} \qquad (13)$$

or

$$\sigma\text{dB} = 10 \log e \,.\, \ln(V_1/V_2) \qquad (14)$$

It has already been shown above that the time difference τ measured when discharging V_1 and V_2 with equal time-constants will be

$$\tau = T\,\ln(V_1/V_2) \qquad (15)$$

If the oscillator is of frequency fc/s, then in a time τ the counter will make $\tau\, f$ counts.

For accurate readings, the counter should make 100 counts per 1dB v.s.w.r., and so substituting in equation (15):

$$Tf\ \ln(V_1/V_2) = 100\ (10 \log e \,.\, \ln V_1/V_2) \qquad (16)$$

or

$$Tf = 1\,000 \log e \qquad (17)$$

So, if the oscillator frequency is known, the discharge time-constant may be suitably fixed to make the counter direct reading.

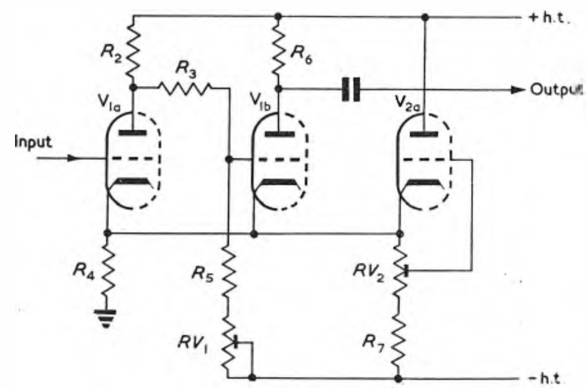

Fig. 9. Schmitt trigger circuit

Apart from the errors already discussed above, there will be further errors in this system if there are differences between the threshold voltages of the two Schmitt trigger circuits or in the time-constants of the two discharging circuits.

Suppose that the two threshold voltages are v and $v+\delta v$ and that the two time-constants are T and $T + \delta T$.

The maximum error will occur when the longer time-constant is associated with the lower threshold voltage, so that if both discharges start from a voltage V, then for one circuit

$$v = V \exp\left(-\frac{-t_1}{T + \delta T}\right) \qquad (18)$$

and for the other circuit

$$v + \delta v = V \exp(-t_2/T) \qquad (19)$$

From these two equations

$$\ln \frac{v + \delta v}{v} = \frac{t_1}{T + \delta T} - t_2/T \qquad (20)$$

or

$$t_1 - t_2 = (T + \delta T)\ln\frac{v + \delta v}{v} + t_2\,\delta T/T \qquad (21)$$

This equation shows that, as would be expected, the tolerance on the threshold voltage must be small compared with this voltage, and that δT must be small compared with T.

The system as described depends greatly on the stability of the changeover voltage of the Schmitt trigger circuits, a stability which can only be achieved by the use of stable components and well stabilized supplies, including the negative supply.

It is possible to greatly simplify the circuits in both the simultaneous discharge and the voltage equalization methods by using the well defined cut-off of modern crystal diodes.

The replacement for the Schmitt trigger circuits is shown in Fig. 10. The voltage existing across the storage capacitor C_1 and its associated discharging resistor R_1 is applied to the diode MR_1 through current limiting and isolating resistors R_2 and R_3.

During the time that the voltage on both the storage capacitors is greater than the biasing voltage V_b, only MR_1 is conducting and so the a.c. signal does not reach the output. As soon as the voltage on C_5 falls to V_b, MR_2 will provide a path to the output for the a.c. signal. When C_1 falls to V_b volts, the diode MR_1 stops conducting and cuts off the a.c. signal.

The time duration of the a.c. signal at the output may either be measured by counting the number of cycles transmitted, or by some other convenient method such as charging a capacitor.

A similar system can also be utilized with the voltage equalizing circuit, as is shown in Fig. 11.

As long as the voltage of C_1 is higher than that of C_2

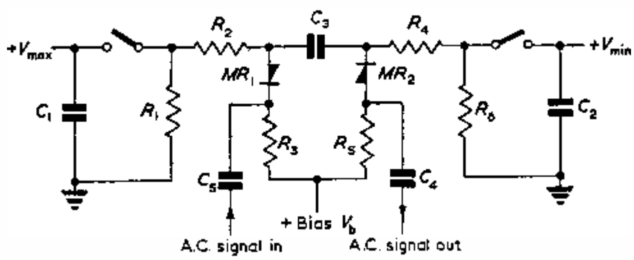

Fig. 10. Simultaneous discharge system with diodes

the diode is biased in the reverse direction, and the a.c. signal is not passed to the amplifier.

When C_1 has discharged through R_1 to the same voltage as C_2, the diode MR_1 starts to conduct, the a.c. signal is passed to the amplifier and having been rectified operates the relay A, ending the discharge. The time required to equalize the voltages can be measured by any of the time measuring circuits mentioned earlier, which can conveniently be actuated by additional contacts on relay A.

With equal voltages on the cathodes of V_1 it is necessary to connect the diode MR_1 as shown so that it receives the slight positive bias necessary before it will start to conduct.

The Mullard OA202 silicon junction diode was found to be very suitable in this application. The amplitude of a.c signal used was 50mV, and the necessary change in d.c. bias between the conducting and non-conducting states was found to be less than 0·1V under these conditions.

Great simplifications could be achieved in the indicating system by the use of logarithmic circuits, and Fig. 12 shows one system that could be used. If the tuned amplifier of the standard system can be replaced either fully or partly by a tuned logarithmic amplifier, then the demodulator will supply voltages proportional to the logarithm of the probe voltage. The maximum minimum detector stores the two extreme voltages, and it is now sufficient merely to measure their difference to measure the magnitude of the v.s.w.r. This can be done either by a 'floating' valve-voltmeter circuit, or by a pair of cathode-followers with a meter between the cathodes. It is even simpler to discharge the voltage difference between the two storage capacitors through an integrating galvanometer. This galvanometer would integrate the discharge current, which integral is proportional to the voltage difference between the two provided that their capacitances are equal. A reading proportional to the logarithm of the v.s.w.r. is then obtained in a steady form.

This circuit automatically reduces the extent of voltage variations at the demodulator, thus allowing greater limits of operation without adjusting the gain of the amplifier.

Practical Details

All the circuits described can only function correctly if the voltage from the demodulator is held between certain limits during the measurement—albeit fairly wide limits. A simple warning circuit was developed to give the operator visual warning if these limits are exceeded (Fig. 13).

This circuit consists of a grid leak detector as the low level indicator and a valve normally biased to cut-off as the high level indicator.

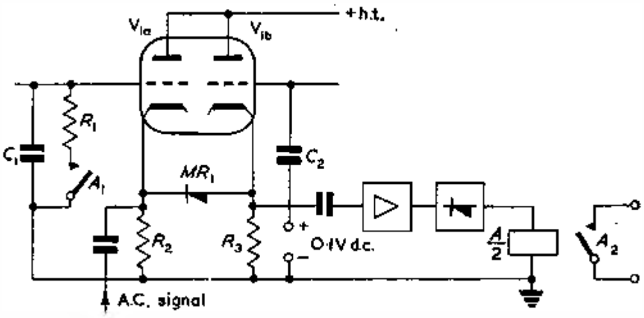

Fig. 11. Voltage equalization system with diode gate

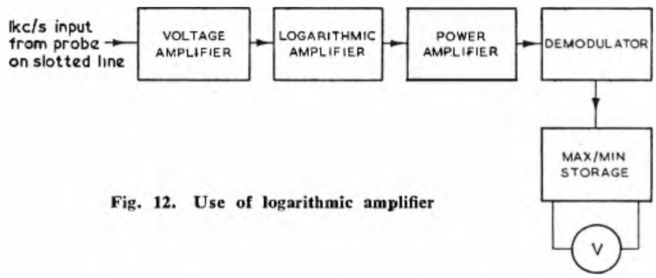

Fig. 12. Use of logarithmic amplifier

In Fig. 13, V_A is the grid leak detector, which gets its signal from the anode of the power stage of the selective amplifier. The grid of V_A is initially biased slightly positive, but above a certain signal voltage (set by RV_1) the valve is cut off as a result of grid rectification, and the neon Lp_1 extinguished. If the voltage of the signal is reduced below this threshold level, the valve starts conducting and the neon lights up giving a warning on the front panel of the instrument.

V_b is biased into cut-off to such an extent that it only starts conducting when the signal level reaches the maximum obtainable from the power output stage before distortion sets in. This maximum value can be set by RV_2, and a second neon Lp_2 on the front panel is lit when the maximum level is exceeded.

The system using a ratiometer as described previously was built around a meter which indicated a maximum ratio of currents in its coils of 10 : 1. Preliminary tests on the instrument showed that the acceptable ranges of currents for the two coils were 0·1 to 1·0mA and 1·0 to 10·0mA respectively, and also that much better discrimination was obtainable at high than at low ratios. This led to a cathode-follower arrangement which with equal voltages on the storage capacitors, would cause the meter to read full scale.

In order to get the necessary low output impedance from the cathode-follower to be able to supply a maximum of

10mA to the high current coil, and still keep the output voltage linear with the input, it was necessary to use five half sections of 12AT7 valves in parallel as V_2 in Fig. 5. Fortunately, as the other coil of the meter only required a maximum of 1mA one half section of a 12AT7 was suitable as V_3.

The input impedance of the system was kept very high by making the cathode resistors R_1 and R_2 of the first valve of each pair 1MΩ. This high input impedance meant that the meter indication would remain quite steady for many minutes.

The virtues of the cathode-follower-ratiometer system are the simplicity of the switching arrangements necessary and the instantaneous reading obtained on the meter.

The switching sequence is initiated by a control switch mounted on the probe carriage of the slotted line. This is a three position switch, spring loaded to return to the central position and is a convenient method of moving the probe carriage. In this particular case it was arranged that with

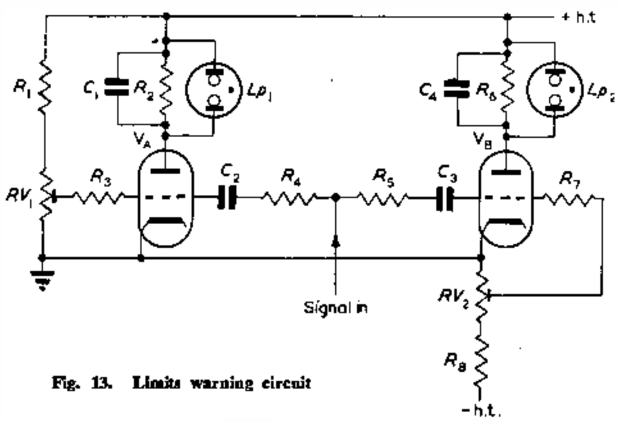

Fig. 13. Limits warning circuit

the control switch in the central position, both S_1 and S_2 (Fig. 2) were open. With the switch in the left position S_1 and S_2 are closed and with the switch in the right position S_2 only is closed. The sequence of operations to take a reading is first to move the control switch to the left, and then having moved the switch to the right, to push the probe carriage along the slotted line and, finally, to release the control switch.

It was arranged for the ratiometer to function as a conventional v.s.w.r. indicator by passing a constant current through one of its coils and connecting the other to the demodulator circuit.

With the system using Schmitt trigger circuits to time the simultaneous discharge of the two storage capacitors, it was obvious from the first that the practical problems would be associated with getting two circuits to work, and more important, to stay working in a precisely similar manner. Very well stabilized positive and negative h.t. supplies were found to be a prime requirement, as were high stability, close tolerance resistors for R_2, R_3, R_5 in Fig. 9. RV_1 and RV_3 allow for differences between valves in the initial setting up procedure.

In this setting up, it is first necessary to ensure that the threshold voltages of the two Schmitt trigger circuits are the same and then to check that the discharge time-constants are equal. Both of these tests are best carried out by observing the residual v.s.w.r. reading on the read-out device when unity v.s.w.r. is measured. (This is best obtained by not moving the probe). With careful adjustment the reading can be kept below 1·02 when this test is carried out. What is more important, it was found that over a period of many weeks the operation of the circuit was such that no further adjustment was needed.

The switching arrangements for this system were rather complex, and to get correct operation a certain sequence had to be followed.

The control switch had first to be moved to the left, which closed S_1, connected the storage capacitors to their charging circuit via MR_1 and MR_2 and connected the meter across the output of the demodulator, where it functioned as a conventional v.s.w.r. indicator. In the time measuring circuit S_{1b} charged the capacitor C to a voltage determined by the potentiometer RV.

On returning the control switch to the central position, all the relays remained operated, but on moving the switch to the right the meter was reconnected to the time measuring circuit and S_1 opened. This is the recording period when C_1 and C_2, the storage capacitors, are charged to their respective voltages. Returning the control switch to the central position connects the storage capacitors to the discharging resistors and initiates the computing period. After a short delay (1 to 2 seconds) the result appears on the meter.

The limit to the accuracy that can be achieved with the voltage equalization system is the accuracy with which the equality of the two voltages on the storage capacitors can be measured. Two stages of voltage amplification and a cathode-follower stage to operate the discharging relay were found to be sufficient.

The equalization of the two voltages could be carried out to within 50mV without any particular refinements to the circuits. Only a simple chopper was used, and no attempt was made to remove hum from the amplifier. The accuracy of the v.s.w.r. measurement achieved was satisfactory, giving a residual v.s.w.r. of approximately 1·05.

Conclusion

Systems have been developed which can take much of the labour from v.s.w.r. measurements using slotted lines and still give access to the important phase information if this is required. Their accuracy is quite sufficient for all normal requirements, and the persistent meter reading gives less chance of reading errors than with the conventional system.

It is felt that these systems might with advantage be adapted for use in other applications where the ratio of two quantities is to be measured. Applications that come to mind are in modulation meters, and in the measurement of noise factor of receivers.

Acknowledgments

The authors wish to thank Marconi Instruments Ltd for permission to publish this article.

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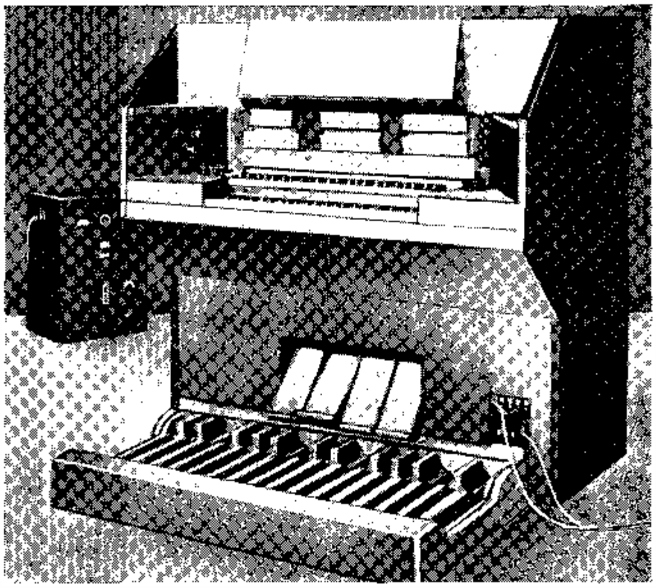
Gas Tubes as Music Generators

By Alan Douglas, M.I.R.E., M.Amer.I.E.E.

The obvious attractions of two electrode gas tubes as relaxation oscillators would suggest that many problems in the generation of electrical music could be economically solved thereby; but, in fact, no commercial instrument uses such devices. Some recent circuits make the best of the inherent physical limitations of inexpensive gas tubes.



(Voir page 707 pour le résumé en français: Zusammenfassung in deutscher Sprache Seite 712)

As long ago as 1928, W. Kock and O. Vierling designed a two-manual organ using two electrode neon tubes as sawtooth oscillators. At that time, such tubes had reached a higher state of development in Germany than in any other country. An instrument of this kind was then a great novelty, and its vagaries were accepted as a matter of course.

Regardless, however, of the prime source of oscillations, the formation of acceptable tone colours by synthesis was not at that time understood, and because of this we find a lapse of nearly ten years before further developments took place. By this time the hot cathode three-electrode gas tube was well established, and the time appeared ripe for another approach to gas tube oscillators. But such tubes are too costly for polyphonic instruments and this article refers only to inexpensive two electrode cold-cathode tubes.

Because of the basic fact that all gas tubes, on ionization, can pass a considerable current which may be used to discharge a capacitor, it is comparatively easy to obtain a linear sawtooth wave from such oscillators. If the charging current I is held constant, the charge on the capacitor C at any time t after the tube goes out is

$$Q = Q_o + It$$

where Q_o is the charge on C at the instant of tube extinction.

The voltage across C and the tube is

$$V = (Q/C) = (Q_o/C) + (It/C) = (V_e + (It/C))$$

Thus the voltage across C rises linearly during the charging period, and this is important because any departure from linearity will affect the distribution of harmonics; which, in turn, will affect the fidelity of synthesized tones. Should I, for instance, not be constant, then the voltage across C will rise exponentially, and if the line voltage is low, the curvature may be quite serious musically.

While the full charging voltage is usually obtained at musical frequencies, there is generally some uncertainty as to the instant at which the discharge through the tube commences, and even more about the instant at which this ceases. The most important factor controlling this is the configuration of the external circuits associated with the tube. Other things being equal, if the current is so great that the whole of the cathode is covered by the glow, the cathode drop will be high; on discharge, the voltage may fall as A in Fig. 1. If, however, that current is limited in some way so that the cathode is not completely covered by the glow, the cathode drop will not affect the voltage (B in Fig. 1). Although large voltages are not essential in a music generator, there must be a much higher voltage across the tube than the actual ionizing potential of the gas used. This is because the mechanism by which the glow discharge is built up and maintained depends basically on secondary emission from the cathode, due to positive ion bombardment. As this is a relatively inefficient process, it is necessary for each electron leaving the cathode to produce many ionizations en route to the anode. Thus the discharge running voltage will be at least several times the ionization potential. The best voltage for cheap tubes appears to be about 55V.

If, then, the glow does not completely cover the cathode, the voltage drop depends on the gas used, separation of the electrodes, and the work function and purity of the cathode material; and not on the gas pressure. But the current density per unit area of the cathode does depend on the gas pressure; and, of course, the purity and composition of the gases used affects the performance.

It is well known that cold-cathode tubes of this type have an upper frequency limit. This is because ionization is not instantaneous, requiring a time depending on the applied voltage and the amount of original ionization. The mass of the positive ions call for time to cross to the cathode and accelerate sufficiently to produce ionization by collision. The rate of ionization can be increased with an increase in applied voltage, but since the cathode heats up to some extent with the bombardment, the degree of ionization remaining

The above photograph shows the Kock and Vierling two-manual organ.

when the capacitor discharges may be very variable. If the frequency is high enough, there may be ions which have not had time to recombine. The time for deionization is much greater than that for the current to fall to zero in an external circuit having negligible inductance. Therefore the dynamic characteristic of the tube is not quite the same as the static characteristic. At very low frequencies, however, it is more difficult to control the discharge and in the case of an oscillator in which the tube is used to charge a capacitor, it may be difficult to start ionization. As free electrons can be liberated by incident light, one method exposes the oscillator tubes to a weak light source such as a flashlight bulb. Even so, the rate of ionization may be erratic from one instant to another.

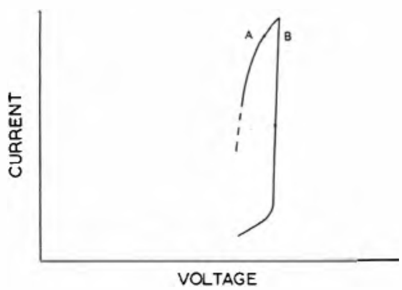


Fig. 1. Current/voltage curve for gas tube

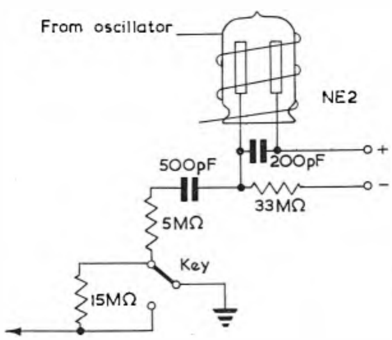


Fig. 2. Method of controlling ignition potential

A method of controlling the ignition potential with some certainty is to provide a magnetic field round the tube. This confines the path of electrons crossing to the anode, and greatly regularizes the action of the discharge. An example of this method is shown in Fig. 2, where an inductance supplied from a previous oscillator is used to increase the locking in a synchronized gas tube frequency divider using very inexpensive tubes (NE2). Such glow tubes have symmetrical electrodes, but it is worth noting that in a tube in which the cathode element is the smaller, the ignition potential is usually lower than if the connexions are reversed. This is a particularly desirable tube on account of its low cost, but discharge is uncertain because of the large physical variations between specimens, especially irregularities over the cathode surface resulting in minor changes in cathode drop. This is a result of cathode material impurity. A particularly irritating effect from this is 'splash' or a random waveform. These give rise to transients, and may occur even with good quality tubes. Severe radio interference can be caused thereby.

All these relatively minor discrepancies add to make it virtually impossible to replace one glow discharge tube of an inexpensive type with another without considerable readjustment of the external circuits. As the impedance of the tube is very low, it is susceptible to external loading and this may distort the waveform. Such loading may take the form of another cascaded divider, for example. It is difficult to assure synchronization over two or three semitones with a cheap gas tube for any length of time, the waveform becoming distorted towards the upper and lower frequency limits. A recent American circuit overcomes this difficulty and also greatly increases the apparent impedance[1].

Fig. 3 shows two gas tubes in series, with series timing capacitors C_5 and C_6. This capacitive voltage divider isolates the output so that loading does not affect the frequency. The two tubes have a high impedance to earth. Capacitors C_3 and C_4 in another divider are 1/100th of the values of C_1 and C_2. As their midpoint is connected to that of V_3 and V_4, the proportion of the output of the first stage at the second stage is nearly proportional to the values of the two capacitors.

Assuming that the negative flyback pulse of the first stage appears before the charge on C_5 and C_6 is high enough to

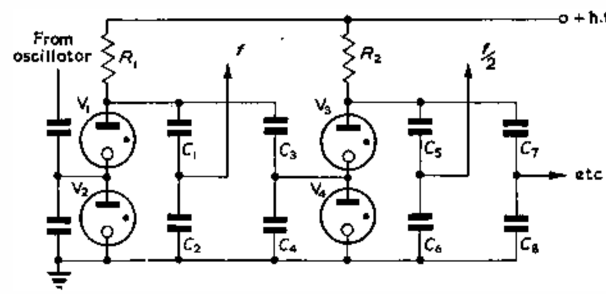


Fig. 3. Gas tubes in series with capacitive voltage dividers

ionize V_3 and V_4 and is applied to the junction of V_3 and V_4; the lower electrode of V_3 then becomes sufficiently negative to ignite it. A higher positive potential then appears on the upper electrode of V_4 and that tube ignites. Capacitors C_5 and C_6 quickly discharge and extinguish V_3 and V_4. Capacitors C_5 and C_6 begin to recharge to produce the rising portion of the second stage output wave. So the first stage triggers the second. Any divided signal fed back to the output of the first stage (a common failing with single tube dividers) must pass through the high impedance voltage divider (C_3 and C_1 in series) and the low impedance shunt divider C_2. Very great attenuation of the fed-back signal is achieved.

Notwithstanding these precautions, a good deal of artificial ageing of cheap neon tubes is necessary to stabilize their performance. Thereafter their behaviour depends to a great extent on how often they are used. Long periods of dis-use frequently result in inability to start and means must be provided to adjust the free-running frequency of such oscillators.

It can be seen that most of the limitations in performance are physical ones, and if the tubes are to remain of an inexpensive type, it would seem impossible to improve their performance to any measurable extent. For the experimenter who does not object to some instability, many ingenious circuits and applications have been devised by Nicholas Langer[2]; and for music generators using three electrode tubes, see Oskar Sala, German patent 917470.

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A Delta Modulation System Using Junction Transistors

By B. E. Williams*, B.Sc.

A delta modulation system, operating at a digit frequency of 14kc/s, is described which provides a simplex speech link with good intelligibility. The equipment has only one manual control, the press-to-talk switch, other adjustments being preset or made automatically. The receiver regenerates the incoming signal by voltage and time slicing before feeding the decoder. The equipment will work satisfactorily over a wide range of ambient temperatures and is suitable for use over equalized lines or radio links. A power supply of between 6 and 12V is required.

(Voir page 707 pour le résumé en français: Zusammenfassung in deutscher Sprache Seite 712)

THIS equipment was designed to provide a simplex speech link using a binary code output signal. The pulse repetition rate was to be as low as was consistent with good intelligibility and the number of controls available to the operator was to be a minimum. The equipment would have to work satisfactorily over a wide range of ambient temperatures.

Simple delta-modulation was chosen as the encoding system because it provides better quality at low digit rates than either pulse code or delta pulse code modulation, both of which have the added disadvantage of requiring a frame synchronizing system. A digit rate of 14kc/s was chosen as a compromise between intelligibility and bandwidth required.

Fig. 1 shows a block diagram of the system. The output of the microphone is equalized before being fed to the input of the constant volume amplifier which ensures that the input level to the coder is near optimum for a wide range of microphone output levels. The output of the coder is converted into an alternating-binary (or three level) signal before transmission. This form of signal has no d.c. component and requires less bandwidth than plain binary code.

The input to the receiver is amplified by the constant level amplifier and then fed to a voltage slicer which removes the noise and degradation which may be produced in the transmission path and also converts the signal back to the usual binary form. Pulses derived from the voltage slicer output are used to lock the receiver timing circuits to the frequency of the sender. The slicer output is time sampled with pulses from the receiver timing circuits to reproduce accurately the coder output, which is then decoded and fed to the telephone earpiece.

All the transistors used in this equipment are of the high frequency junction type although in some circuits low frequency transistors could be used.

The Constant Volume Amplifier

The circuit of the constant volume amplifier (c.v.a.) and microphone equalizer is shown in Fig. 2 and consists of the transistors X_1 to X_9 together with the associated components. The microphone used has a frequency response which rises at the rate of 7dB/octave over the band 300c/s to 3kc/s. The equalizer consists of the components R_2, C_1 which have an insertion loss that increases by 6dB/octave so that the overall response of the microphone and equalizer is approximately flat from 300c/s to 3kc/s.

The parameter used to control the gain of the c.v.a. is the input impedance of a grounded base amplifier X_3. The input impedance is inversely proportional to the emitter bias current. The bias current flowing in R_5 is shared

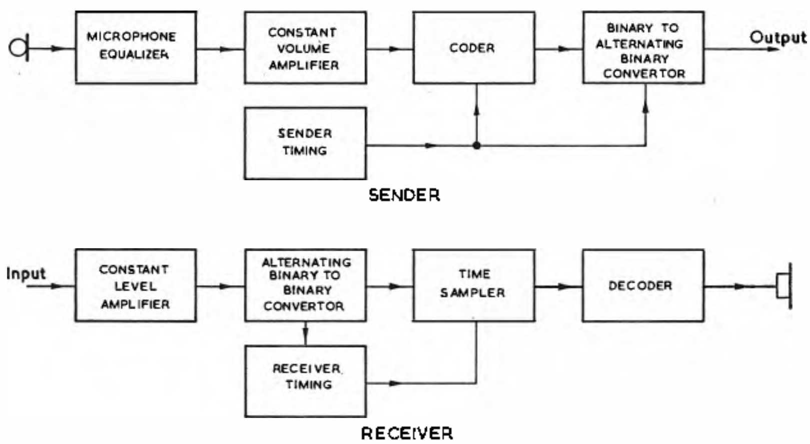

Fig. 1. Block diagram of the delta-modulation system

between the transistors X_1 and X_2 according to the relative potentials of their bases; if the difference is greater than about 0·15V all the current in R_5 flows into the transistor whose base is the more negative. The base of X_2 is held at a fixed voltage by the potentiometer chain R_7, R_8 which is decoupled to audio frequencies by C_6. The input impedance of the control stage, X_3, will then vary over its entire range as the base potential of X_1 is varied by $\pm$0·15V about that of X_2. The output of the c.v.a. is rectified, smoothed and fed back to provide a negative d.c. bias voltage for X_1. For an input signal within the control range the rectified output will be within 0·15V of the reference voltage on the base of the control stage, and should the input level rise, the d.c. bias on X_1 will go more negative and the bias current in X_2 will fall so that the signal input current remains almost constant. The source impedance of the input to the control stage is comparatively low so that the input current is inversely proportional to the input impedance. If the reference voltage is large compared with the 0·3V change in voltage required to control the gain, the difference in amplifier output for signals at extremes of the control range will be small.

The variation in microphone output level is less than 40dB and for this application a 3dB change in output can be tolerated. To obtain the required stability of output the reference voltage should be equal to or greater than 1V and

* *Communication from the Staff of the Research Laboratories of The General Electric Company Limited, Wembley, England.*

for a 40dB control range the ratio of maximum to minimum current in X_2 will be 100. 1mA was chosen as the maximum current so that the minimum will be 10μA. The signal input current, which is constant, must be a fraction of the minimum bias current if non-linearity is to be avoided. The output level from the control stage is therefore low and is amplified by three grounded emitter stages in cascade, X_3 to X_5, giving an overall voltage gain of 10 000 to 40 000,

The source impedance of the rectifier is low compared with R_{27} so the charge time depends on the smoothing components R_{27} and C_{16}. The decay time is determined by C_{16} and R_{28} together with the base current of X_1 which may flow in either direction according to the parameters of the particular transistor used. At high temperatures the collector leakage current flowing in the base of X_1 would, in the absence of R_{28}, charge C_{16} to such a potential that

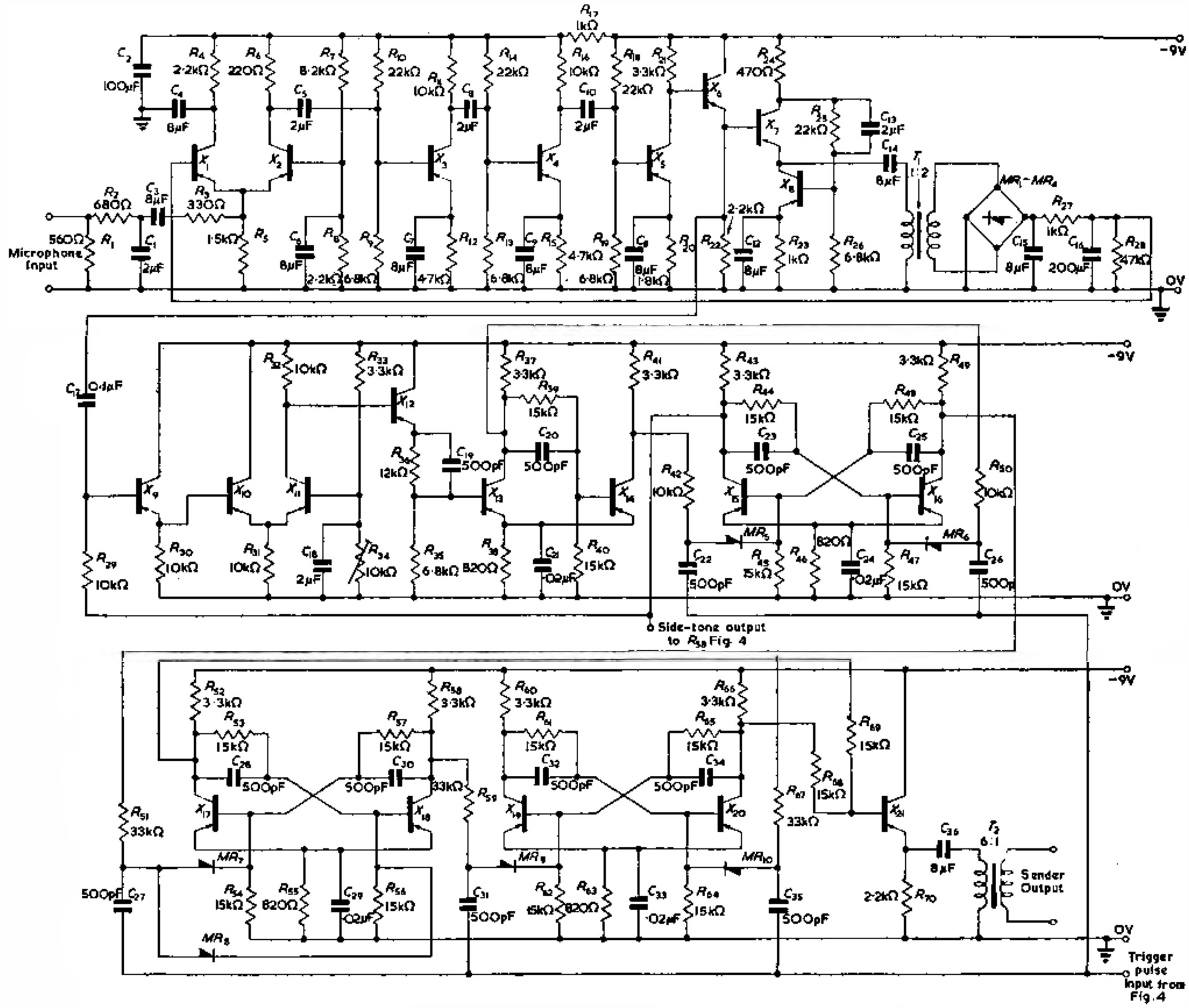

Fig. 2. Delta-modulation sender

depending on the current gains of the particular transistors used. The output of X_5 is emitter-followed by X_6 and fed to the coder and also to a stacked emitter-follower circuit, X_7X_8. This circuit can provide high peak output currents for both positive and negative input signals and so is suitable for driving the full wave bridge rectifier (MR_1 to MR_4). A full wave rectifier is used because the component of the ripple at the input frequency is very small, being due mainly to the differences in forward characteristics of the bridge diodes, with the result that stability is obtained with less smoothing than is required by a half-wave rectifier. The rectifier circuit should have a differential time-constant; that is the d.c. output should be able to increase rapidly to adjust for an increase in input but should decay slowly, in the absence of an input, to bridge over the silent periods between words and sentences.

the rectifier would cease to operate. The resistor R_{28} will prevent this occurring for values of leakage current less than 20μA.

The control stage is unbalanced, i.e. when the gain alters, an output, corresponding to the change in control current through R_6, is fed to the amplifier stage X_3. To minimize the effect of this unbalance the control stage load has a low resistance (220Ω) so that the d.c. drop across it will be small.

Since the input to the control stage is also applied to the emitter of X_1 the signal current in each transistor will be in the ratio of their bias currents. At high input levels the signal current fed into X_1 will be many times that in the control stage and so the collector of X_1 is decoupled to prevent the input signal by-passing the control stage via X_1 and the supply line.

Referring to Fig. 3 an estimate of the control range of the c.v.a. can be made as follows:

For an input in the control range the output current of X_1, and since $\alpha \simeq 1$ the input current also, is almost constant and equal to V/AR_L amperes peak-to-peak. From linearity considerations let the maximum peak-to-peak input current be limited to $1/N^{th}$ of the control stage bias current whose minimum is then $= NV/AR_L$ amperes. If the input impedance of $X_1 = re = kT/qI_e = 25/I_e$ at room temperature with I_e in milliamperes[1].

$$\text{The maximum input impedance} = \frac{25 \,.\, A \,.\, R_L}{1000 \,.\, N \,.\, V}\,\text{ohms}$$

$$\text{maximum input (volts peak-to-peak)} = V/AR_L$$

$$\frac{25 \,.\, A \,.\, R_L}{1000 \,.\, N \,.\, V} = (25/N)\text{mV peak-to-peak}$$

If I_M (mA) is the maximum control current, then control range

$$= \frac{\text{Maximum control current}}{\text{Minimum control current}} = \frac{I_M \,.\, A \,.\, R_L}{1\,000 \,.\, N \,.\, V}$$

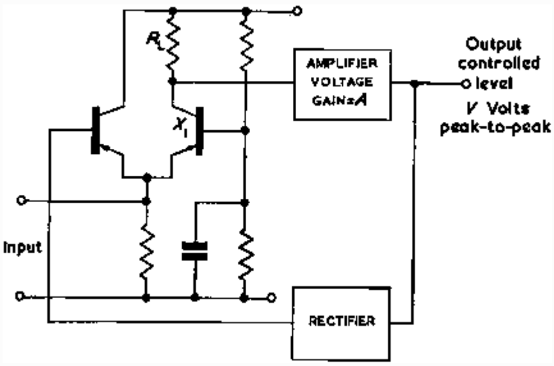

Fig. 3. Constant volume amplifier diagram used to estimate the control range

The assumption is that $re = kT/qI_e$ is accurate for emitter currents greater than about $2I_{co}$. In this circuit at high temperatures I_{co} may be as large as 20μA and the input impedance can be 20 per cent different from the value calculated. This error reduces the control range for a fixed current ratio by 2dB. It is interesting to note that the maximum input voltage depends only on the degree of distortion that can be tolerated. The quality of low digit rate delta modulation systems is such that the distortion produced when $N = 2$ cannot be detected in the decoder output.

The Coding Circuit

The form of delta modulation used in this equipment is a simple system for conveying information, in this case speech, in the form of a train of binary code pulses. The polarity of the pulses indicates whether the input signal is greater or less than an internally generated comparison signal at each digit period. All code pulses represent the same change of input signal amplitude so that this system does not require a frame synchronizing signal. The comparison signal is generated by integrating the binary code output. The comparison and input signals are compared and the difference voltage is used, after amplication, to control the polarity of the next code pulse so that the difference will be reduced. To recover the information at the receiver the binary code is regenerated and then integrated to produce an output, identical to the comparison signal in the coder, which is passed through a low-pass filter and then fed to the telephone earpiece. The d.c. levels in the coder feedback loop are so arranged that in the absence of an input signal the coder output consists of alternate positive and negative pulses. The comparison signal therefore varies about the mean of the binary code voltage levels; this is a suitable operating point because it produces the highest frequency idling pattern and allows equal positive and negative excursions of the comparison signal. A lower frequency idling pattern would produce an audible note in the telephone earpiece. The maximum rate of change of input voltage that the comparison signal can follow is determined by the time-constant of the integrating network and the binary code amplitude. The maximum input amplitude that can be coded at any frequency therefore falls at the rate of 6dB/octave. This slope limitation is no great disadvantage when coding speech because the power spectrum of speech falls at about 6dB/octave above 800c/s so that providing the 800c/s component does not overload the coder the higher frequencies are not attenuated. The accuracy with which the higher frequencies are coded falls because fewer digits are available to define each cycle.

The decoded receiver output contains, in addition to the information transmitted, quantizing noise which is produced in the coding process. The amount of quantizing noise in the output tends to be constant and independent of the amplitude of the signal being coded. The signal-to-noise ratio therefore improves as the input to the coder is increased. It has been found experimentally that with low digit rates it is preferable to slope overload the coder on signal peaks to obtain the best signal-to-noise ratio, in spite of the peak clipping that occurs. The input level at which the modulator begins to code is in general the peak-to-peak amplitude of the comparison signal idling pattern; for inputs smaller than this the coder is undisturbed. The result on the comparison signal of integrating one binary code pulse is known as the delta-step and represents the smallest change in signal amplitude that can be coded.

Referring to Fig. 2 the modulator consists of the transistors X_9 to X_{15} and the associated components. The binary code produced at the collector of X_{15} is fed to the coder integrating network $R_{29}C_{17}$. The integrating capacitor is returned to earth through the output impedance of the emitter-follower X_6 in the c.v.a.; this impedance is sufficiently low for the charging current of C_{17} to produce a change in voltage across it much smaller than the delta-step.

The sum of the audio output of X_6, the comparison signal acros C_{17} and a d.c. level is emitter-followed by X_9 and directly connected to the input of an emitter-coupled slicer X_{10} and X_{11}. The slicer output across R_{32} is emitter-followed by X_{12} and fed into a Schmitt trigger circuit X_{13} and X_{14}; this circuit produces outputs at its collectors which correspond to the input from the slicer but which are of opposite polarity and of a suitable amplitude and d.c. level to be used as steering potentials for the Eccles-Jordan bistable flip-flop X_{15} and X_{16}. The flip-flop is fed with a train of positive going trigger pulses through the capacitors C_{22} and C_{26} and produces the binary code at its collectors. The flip-flop is triggered or not according to its present state and the polarity of the steering potentials produced by the Schmitt trigger circuit. The slicer compares the input to X_9 with the bias potential at the base of X_{11} and according to the sign of the difference voltage the flip-flop is triggered in such a way that the error is reduced during the next digit period. For example, suppose the input to X_9 is more negative than the slicing level, X_{11} is then cut off and the base of X_{13} approaches the supply potential, X_{13} therefore conducts heavily and X_{14} is cut off. The steering potential on MR_6 is then several volts positive with respect to that of MR_5 and a trigger pulse is fed into the base of X_{16} switching it off and X_{15} on. The next binary code pulse is therefore positive and so reduces the error voltage.

Provided the coder is not slope overloading the method used to add the audio input and comparison signals results in almost constant current charging of the integrating capacitor because the potential of X_9 base only varies by a single delta-step. The time-constant of the integrating circuit has been chosen to produce a delta-step of about 0·4V peak-to-peak; the slicer gain is then sufficient to switch the state of the Schmitt trigger with an input of less than half of the delta-step. Also, the delta-step then covers the slicing range even at high temperatures when due to the change in base current of X_3 the mean level of its input alters.

Binary to Alternating Binary Convertor

The output from the coding circuit, which is a binary pulse code signal, is converted into 'alternating binary' form before being transmitted to the receiver. This form of signal has three equally spaced voltage levels, the centre level corresponding to one binary state and both positive and negative levels corresponding to the other. In this case the negative binary state corresponds to the centre level. In the alternating binary signal consecutive positive binary pulses are transmitted as alternately positive and negative pulses. It is clear that while the binary signal may remain in one state for a number of digits the maximum pulse duration of the alternating binary signal is one digit period. It follows that the low frequency response of transmission circuits required to handle this code can be far worse than if the circuits were to handle binary code. The coder slope overloads on peaks of the audio input and so groups of twenty or so similar pulses in the output are common. The -3dB point in the low frequency response of circuits to handle binary code with groups of twenty similar pulses and produce less than 5 per cent droop would be 5·5 c/s whereas with alternating binary code the corresponding figure is 110c/s. The high frequency response required by both types of signal is the same and so alternating binary provides a saving in bandwidth of over four octaves.

To convert binary code into the alternating binary equivalent, the binary signal is used to gate digit rate pulses into a single stage binary counter. The output of the binary counter is delayed by one digit period, inverted in phase and then added to the original counter output. The result of this addition is the alternating binary signal and this conversion is illustrated by Fig. 5.

The circuit of the convertor is shown in Fig. 2 and consists of the transistors X_{17} to X_{21}. The binary code at the collector of X_{16} is used to steer trigger pulses into the counter. If the binary code is in the more negative state there is a reverse bias on the diodes MR_7 and MR_8 greater than the amplitude of the trigger pulses so that the counter is not triggered. If the binary code is positive, trigger pulses will be fed through MR_7 and MR_8 to trigger the counter which then changes state on every pulse. The operation of the counter is as follows; suppose X_{17} is conducting and X_{18} is cut off the voltage across C_{30} will be several times that across C_{28}, the polarity of the charges being the same and such that the plate connected to the transistor base is the more positive. The trigger pulse is fed to both bases and turns both transistors off. The time-constants of the capacitors C_{28}, C_{31} are long compared with the trigger pulse width so that when the pulse ends the charge on each is unaltered, and since there is less charge on C_{28} than on C_{30} the base of X_{18} will be more negative than the base of X_{17}. The circuit therefore settles down with X_{18} conducting and X_{17} cut off—the opposite condition to that existing before the trigger pulse was applied.

The outputs at the collectors of the binary counter are fed through the steering networks $R_{59}C_{31}$ and $R_{67}C_{35}$ to control the d.c. level of the trigger pulses fed to a similar bistable stage consisting of the transistors X_{19} and X_{20}. The trigger pulses are fed into the base whose steering potential is the more positive. The time-constant of the steering networks is about a third of a digit period so that the d.c. levels of the pulses applied to MR_9 and MR_{10} depend on the state of the binary counter during the previous digit period.

This stage therefore produces outputs identical with the binary counter but delayed by one digit period.

One output is taken from the counter stage and another from the delay stage such that the waveforms are out of phase and these outputs are combined in a resistive adder R_{68}, R_{69} to form the alternating binary code. The output of the resistive adder is emitter-followed by X_{21} into the output transformer T_2 which will produce about 1V peak-to-peak in a 150Ω load.

The Constant Level Amplifier

The voltage slicer circuit in the receiver requires a fixed input level for optimum performance and to provide this constant level an amplifier is used which is nearly identical with the constant volume amplifier already described. In this case the control range required is 30dB and the accuracy and stability of the output level should be $\pm 0{\cdot}5$dB. Since high frequency transistors are used the high frequency response of the constant volume amplifier is more than adequate to handle the alternating binary signal; the low frequency response has been improved by increasing the capacitance of the emitter by-pass capacitors C_5, C_7, C_9 so that the droop produced on the input signal is negligible. The linearity of the constant level amplifier must be better than that of the audio amplifier and so the maximum input applied to the control stage is reduced

The circuit of the constant level amplifier is shown in Fig. 4 and consists of the transistors X_1 to X_8. The form of the circuit is identical with that of the constant volume amplifier except that in this case the input from the transformer T_1 is fed in series with the emitter bias resistor R_2 to avoid using a large coupling capacitor. The reference voltage on the base of the control stage is about 3V to provide the required stability of output level.

The Alternating-Binary to Binary Convertor

The output of the sender unit is an alternating-binary pulse code signal and before it can be decoded it must be converted back into binary code. The received signal may have noise added and its waveform degraded by the transmission path between sender and receiver. The binary code signal used for decoding must consist of pulses of equal duration and amplitude or the signal-to-noise ratio of the audio output will be impaired. The system used in this equipment to regenerate the binary code is to feed the alternating-binary signal at a constant amplitude into voltage slicers which slice the positive and negative pulses and combine them to form a binary output in which, due to noise and degrading of the signal, the pulses may not be integral multiples of the digit period. The slicer output is integrated and then time sampled by digit rate pulses, produced by the receiver timing circuits, which are locked in phase to the incoming signal. The time-sampled binary code is then fed to the decoder and audio amplifier. The receiver timing circuits are locked to the frequency of the input signal by pulses formed by differentiating the slicer output.

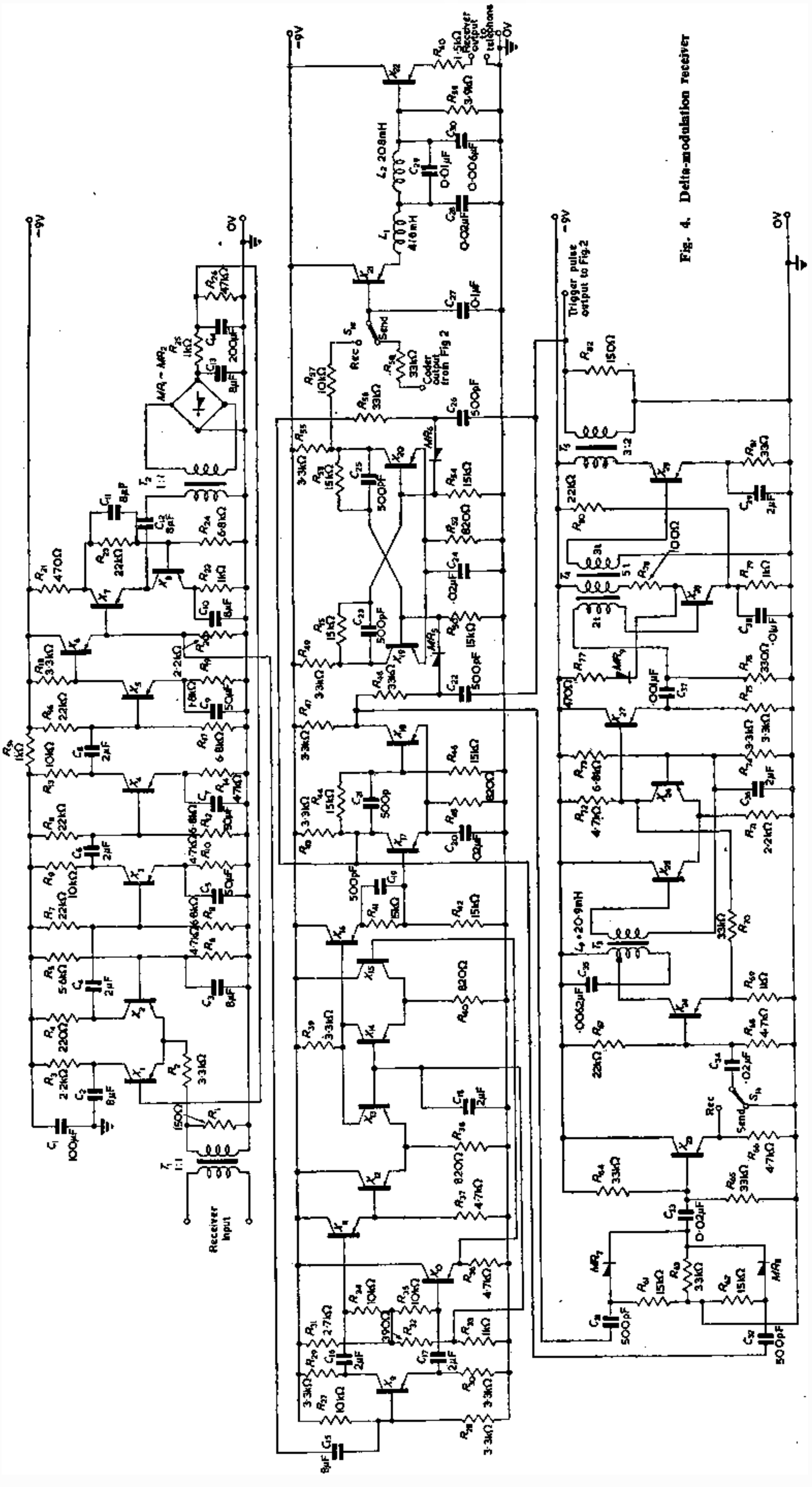

Fig. 4. Delta-modulation receiver

Referring to Fig. 4, the constant level amplifier delivers the received signal at its output at an amplitude of 4V peak-to-peak; this output is fed to a phase-splitting stage X_9, which produces outputs of equal amplitude and opposite phase. Each phase is then emitter-followed to the input of an emitter coupled slicer. Both slicers have the same circuit configuration and component values and have a common output collector load, R_{39}. The bases of the slicer transistors X_{14} and X_{15} are connected together to a point on the potential divider R_{31}, R_{32}, R_{33} which is positive with respect to the d.c. level of the slicer inputs by the voltage drop across R_{32}. The voltage across R_{32} is arranged to be 25 per cent of the peak-to-peak amplitude of the slicer's inputs which are therefore voltage sliced at a point midway between the centre and the more positive levels. Slicing at this point gives the maximum discrimination against noise in the slicer output. The output across the common slicer load R_{39} is the binary equivalent of the slicer input signal since when the centre level is being received neither slicer draws current through R_{39}: at other times one or other of the slicers draws current. It is not possible for both slicers to draw current through R_{39} simultaneously since the signals at their inputs are in anti-phase. The operation of the slicers is illustrated in Fig. 5.

The binary code produced across the slicer's common load resistor is emitter followed into a Schmitt trigger circuit formed by the transistors X_{17} and X_{18}; this circuit gives outputs across its collector loads corresponding to the binary code but of opposite polarity. The outputs are of suitable amplitude and d.c. level to be used as steering potentials for the time sampling circuit. The receiver timing circuits generate short duration positive-going trigger pulses at the digit frequency which occur about 60 per cent of a digit period after the cross-overs in the binary code output of the voltage slicers. These pulses are fed through the capacitors C_{22} and C_{23} to the diodes MR_5 and MR_6 which are connected to the bases of an Eccles-Jordan bistable flip-flop X_{19} and X_{20}. The d.c. levels of the trigger pulses fed to MR_5 and MR_6 are determined by the Schmitt trigger outputs which are fed through the resistors R_{48} and R_{56}. The amplitude of the trigger pulses is such that they overcome the reverse bias only on the diode whose steering network is connected to the more positive Schmitt trigger output.

The Schmitt trigger therefore controls the trigger pulse input to the flip-flop so that the binary code generated by the sender is accurately reproduced at its collectors. The output of the flip-flop is then fed to the decoding circuit and audio amplifier.

Decoding Circuit and Audio Amplifier

To reproduce the coder comparison signal at the receiver the regenerated binary code is integrated; this signal contains the information fed into the coder and also the quantizing noise component which is produced because the comparison signal is only an approximation of the coder input. The output of the integrating network is fed to the telephone earpiece after passing through a low-pass filter which removes the components of the quantizing noise that lie outside the pass-band of the filter. When the equipment is used to transmit the decoder provides a side-tone signal by decoding the local coder output; this enables a check to be made on the operation of the constant volume amplifier and the coder circuits.

Referring to Fig. 4 the switch S_{1a} selects either the regenerated receiver input or the coder output and feeds the integrating capacitor C_{27}. The level of the sidetone is required to be less than that of the receiver output so the value of the integrating resistor R_{58} used to produce the sidetone is about three times that of R_{57}. The output of the integrator is emitter-followed by X_{21} into the low-pass filter which has a cut-off frequency of 3kc/s and consists of a constant-k section followed by an m-derived section, with an infinite attenuation at 3·5kc/s. The filter is terminated by R_{59} which also provides a d.c. return for the emitter current of X_{21}; the effect of the bias current on the filter coils is to reduce their inductance by less than 1 per cent and can therefore be neglected. An m-derived filter section is used because it is possible for the coder to produce a stable idling pattern of two positive pulses followed by two negative pulses; this pattern has a fundamental component at 3·5kc/s and can cause an unpleasant background whistle. The filtered output is emitter-followed by X_{22} and fed directly to the telephone earpiece which is a low impedance unit and together with R_{60} forms the emitter load of X_{22}.

The Timing Circuits

The digit rate of the sender is determined by the frequency of the train of trigger pulses fed to it and in the receiver a source of trigger pulses locked in phase to the sender output is required to regenerate accurately the transmitted signal. This requires that the digit frequency of the sender should be accurate and stable if the receiver is to have no controls. The sender trigger pulses are generated by a blocking oscillator whose frequency is controlled by a stable LC oscillator. The equipment is a simplex system and so the same oscillator and pulse generator can be used for the receiver if a synchronizing signal is fed in.

The timing circuits are shown in Fig. 4 and consist of the transistors X_{23} to X_{29}. The oscillator circuit[2] is formed by the transistors X_{24} to X_{26}. In the sending condition, the capacitor C_{34} is earthed and X_{24} operates as a class-A grounded base amplifier whose output is fed to a tap on

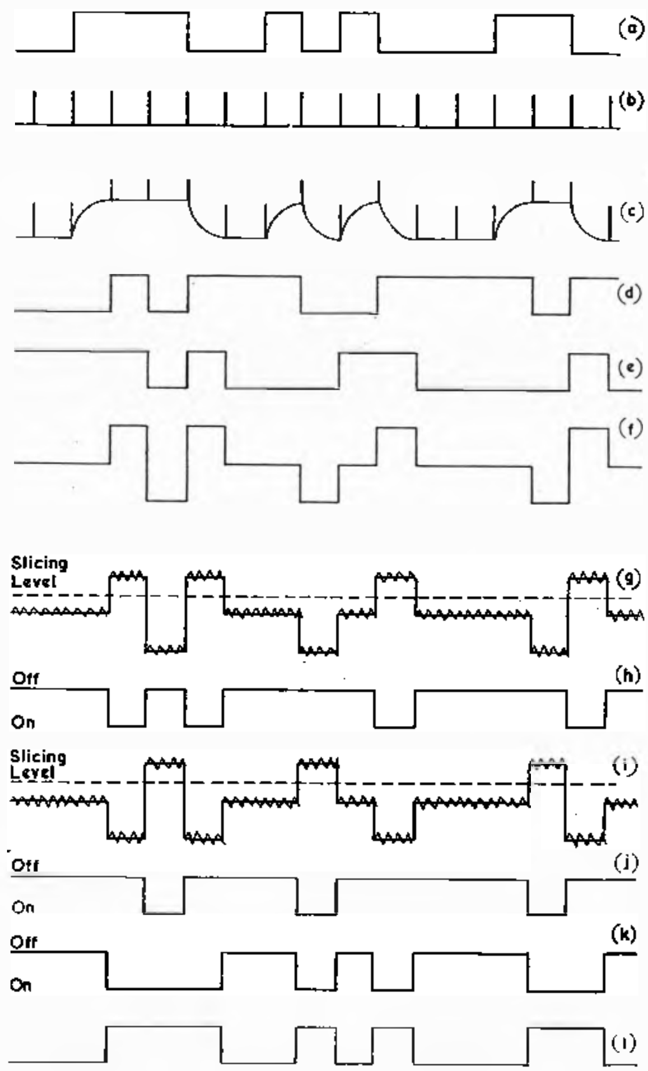

Fig. 5 (a)-(f) Waveforms in binary to alternating binary convertor (Fig. 2)
(g)-(e) Waveforms in alternating binary to binary convertor (Fig. 4)

(a) Coder output from X_{16} (b) Trigger pulses (c) Junction of R_{61} and C_{27} (d) Counter output from X_{17} (e) Delay stage output from X_{20} (f) Alternating binary code at X_{21} (g) Input to base of X_{15} (j) Current in collector of X_{14} (k) Current through R_{39} (l) Binary code across R_{39}

the primary of a tuned transformer T_3. The secondary of the transformer is connected between the bases of an emitter-coupled slicer X_{25} and X_{26}. The output of the slicer if fed back through R_{70} to the emitter of the grounded base amplifier and with the correct connexion of the secondary winding of T_3 oscillations can be maintained in the tuned circuit at an amplitude depending on the feedback current in R_{70}. The value of R_{70} is chosen so that the peak-to-peak input to the slicer is about 2V. Under these conditions the output of the slicer is a square-wave which is very suitable for triggering the blocking oscillator. Also, the mean input impedance of the slicer is high and so does not damp the tuned circuit appreciably. The output impedance of the grounded base amplifier causes very little damping of the tuned circuit because the collector is fed to a tap on the transformer primary. The frequency is adjusted initially by selecting the value of a small trimming capacitor which is in parallel with C_{35} and thereafter will drift by less than 30c/s as the ambient temperature is varied over the range $-20°C$ to $+60°C$.

The square-wave output of the oscillator is emitter-followed by X_{27} and then differentiated by C_{37} and R_{76}, the negative-going pulses being used to trigger the blocking oscillator, X_{28}. The blocking oscillator is prevented from free-running by the small negative bias on its emitter produced by the potential divider R_{79}, R_{80}. The blocker therefore produces an output pulse for every negative pulse from the differentiating network. The output of the blocker is taken from a winding on the transformer T_4 and fed to a pulse amplifier X_{29} which is driven between cut-off and saturation. The pulse amplitude at the collector of X_{29} is therefore clearly defined and about equal to the supply voltage. The required pulse amplitude is about 2/3 of the supply voltage and so a step down transformer T_5 is used; the transformer secondary is connected to produce positive-going output pulses which are of about 0·5μsec duration at half height. A small negative bias is developed across C_{39} and R_{81} in the emitter of the pulse amplifier to ensure that the stage is thermally stable.

In the receiving condition, the same oscillator and pulse generator circuits are used, but instead of C_{34} being earthed it is connected to the emitter of X_{23} which feeds synchronizing pulses into the oscillator. The outputs of the Schmitt trigger circuit X_{17}, X_{18} are fed to the differentiating networks $C_{31}R_{61}$ and $C_{32}R_{62}$ and the positive going pulses only from each network are fed through the diodes MR_7 and MR_8 to the emitter-follower X_{23}. A positive synchronizing pulse is therefore fed into the oscillator for every cross-over in the binary code output of the voltage slicers. In this equipment the receiver will lock to the sender frequency and regenerate the received signal accurately provided that the difference between the sender frequency and the free running frequency of the receiver does not exceed 1 per cent.

Performance

The system described provides a digitized speech channel of good intelligibility which is extremely simple to operate and will work satisfactorily over a wide range of ambient temperatures. The lower limit of $-20°C$ is fixed by freezing of the electrolyte in the electrolytic capacitors and the upper limit of $+60°C$ is set by the maximum allowable junction temperature of the transistors. No trouble was experienced due to collector leakage currents varying with temperature: high frequency transistors were used throughout.

A power supply of between 6 and 12V is required, the power consumption at 9V being 700mW.

By telephone standards the quality of the audio output from the decoder is very poor, the signal-to-noise ratio being about 14dB. The criterion for this equipment, however, is intelligibility rather than quality and in this respect 14kc/s delta modulation provides almost 100 per cent intelligibility.

The receiver will regenerate the incoming digital signal in spite of noise added in the signal path. In an experiment white noise in the band 0 to 20kc/s was added with an otherwise clean input signal and it was found that with a signal-to-noise ratio of 6·6dB the number of incorrectly regenerated digits was 0·1 per cent.

High stability resistors are used throughout with an initial tolerance of ± 5 per cent. The capacitors used are metallized paper with a tolerance of ± 20 per cent, with the exception of the electrolytics and the oscillator tuning capacitor. (Fig. 4, C_{35}).

Acknowledgment

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A New Research Laboratory

A new research laboratory block, which is part of the Research and Development Division of Tube Investments Ltd. has recently come into full use at Hinxton Hall, near Cambridge.

Other establishments within the Division are a Technological Centre and a Computer Unit at Walsall, and a Department of Technical Information and an Operational Research Unit at Birmingham.

Research work was started at Hinxton Hall in 1954 and was concentrated mainly on solid state physics. With the new building a total area of 35 500ft² has now been provided and facilities include a lecture theatre.

Among the new instruments developed by the laboratory are an X-ray scanning microanalyser and a tube eccentricity meter.

The first prototype microanalyser was developed at the Cavendish Laboratory, Cambridge, but in view of its potentialities in metallurgical research, development at Hinxton Hall was concentrated on exploiting its use in the metallurgical field.

In the Hinxton Hall microanalyser the specimen surface is scanned at slow speed by an electron beam focused to a diameter of less than one micron. X-rays are emitted by the surface atoms when irradiated by this beam, and an analysing crystal is used to select the emission from any chosen element present. This emission is then detected and the signal used to modulate the brightness of a second electron beam scanning a fluorescent screen. Since both beams are scanned in synchronism, a picture is 'painted' on this screen, exhibiting the distribution of the element selected over the area being scanned on the specimen.

This X-ray image can be compared with an electron image formed simultaneously by making use of some of the electrons that bounce back off the specimen. This reflection electron image reveals the surface topography very clearly and is displayed side by side with the X-ray image for comparison. Point quantitative analyses and a linear distribution display are also possible.

An electrical method of measuring tube eccentricity has been developed at Hinxton Hall on which wall thickness variations are assessed by their electrical screening effect.

Transmitting coils inside the tube are electromagnetically excited to produce a field and to induce signals in two external coils placed diametrically opposite. If the wall thickness of the tube being tested varies, its screening effect on the electromagnetic field passing through the tube wall causes a different signal to be developed in the two external coils. This signal can be arranged to indicate as a measure of the wall thickness difference.

Arising out of this work, a sensitive micrometer device has been developed. It also operates on a variation of the electrical screening principle. The instrument can be used for recording small displacements in length measurements (e.g. as in creep testing), as a sensitive position indicator, or as a limit switch where an extremely small movement of the operative part must give reliable operation of the switching circuit. In applying the new device, only a very small attachment need be made to the moving part of the device being studied. Its versatility is such that it can be used to study movements of half an inch or a few millionths of an inch.

A Small, High Voltage, Regulated Power Supply with Variable Output

By J. D. O'Toole*

An increasing field of application exists for high voltage power supplies with stable outputs, without any requirement for large amounts of power. The needs of this field appear to have been imperfectly satisfied due to apparent limitations of available small valves. A basic form of circuit which overcomes the apparent limitations is described, and a typical circuit of this type is given with its performance figures.

(Voir page 707 pour le résumé en français: Zusammenfassung in deutscher Sprache Seite 712)

THERE are few small valves which can normally be considered as suitable for the construction of high voltage power supplies with variable output voltages, in the range of one to four kilovolts and current capacity of the order of ten milliamperes. The valves which are available do not immediately make the design of such supplies a straightforward process, because conventional lower voltage circuits require the feedback amplifier valves to have voltage ratings of the same order as the output voltage. A regulating circuit has been devised using the 3D21A (CV2659) beam tetrode as a series regulator and 12AX7 double triodes in the feedback circuit. This basic design reduces the rating requirements and offers a general design approach.

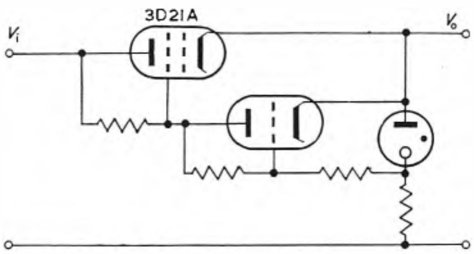


Fig. 1. Shunt stabilized, screen grid supply circuit

Series Regulator

The immediate problem posed by the use of the 3D21A beam tetrode is the low maximum screen grid voltage rating of 850V, compared with the anode voltage rating of 4·5kV maximum under certain conditions of low screen voltage. The use of a second power supply for the screen grid would require this supply to float at the h.t. line potential; because of the insulation requirement, and economy, this arrangement is very undesirable. A simple series resistor in the screen grid supply would not stabilize the grid voltage, because of the large variation of screen current which occurs, and a potentiometer chain across the valve would require an excessively large current to stabilize the screen under all load conditions. An arrangement using a shunt, screen grid, stabilizer circuit, has consequently been chosen. The basic design is shown in Fig. 1, where the triode is the shunt stabilizer, itself stabilized by the gas-filled tube maintaining voltage. The triode anode current at minimum output voltage approximately equals the screen grid current at maximum output voltage and maximum load current. The voltage across the triode remains essentially constant for all load conditions.

Feedback Amplifier

It is easy to see that a triode or pentode amplifier in the feedback circuit, with variable power supply output voltage, controlled by variable amplifier control grid bias, requires approximately 1kV rise of anode voltage for 1kV rise of output voltage. This requires high anode voltage ratings, and therefore special valves, but if a circuit of the type shown in Fig. 2 is used, the large anode voltage swing does not occur. This circuit is basically a cathode coupled, differential amplifier, formed by triodes V_{2a} and V_{2b}, with a high value of fixed cathode resistor R_1 and a variable resistor R_2. Consider the case when $R_2 = 0$; then rise of output voltage V_o as a result of load variation raises the grid voltage of V_{2b}. The anode current rises, but the

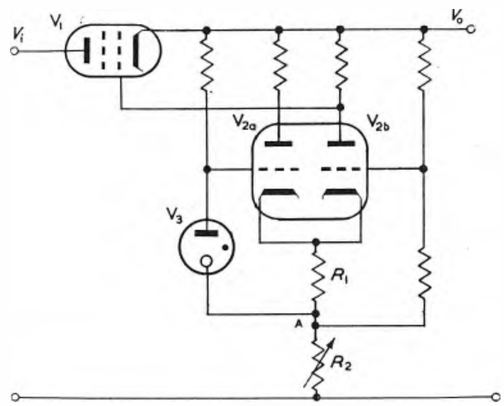


Fig. 2. Feedback amplifier circuit

cathode voltage remains constant, stabilized near the stabilized grid voltage of V_{2a} by a fall of anode current in this valve. The anode voltage of V_{2b} falls and the grid to cathode voltage of V_1 consequently increases negatively, reducing V_o to compensate for its initial change.

Consider next an increase of R_2, raising the potential of point A. The voltage across R_1 remains constant, approximately equal to the stabilizing voltage of V_3, therefore the current through R_1 must remain constant and the cathodes of V_{2a} and V_{2b} must rise in potential by the same amount as point A. A fall of V_{2b} anode current must take place as long as the potential of its grid remains constant; its anode voltage must therefore rise. The grid to cathode voltage of V_1 must fall and V_o consequently rises. Equilibrium is restored when the grid to cathode voltage of V_{2b} has returned to its original value, that is when V_o has risen by the same amount as point A. It can be seen that within quite wide limits the triodes V_{2a} and V_{2b} always operate under essentially the same conditions and the amount by which the output voltage may be raised is limited only by the input voltage V_i.

This circuit has incidental features which are noteworthy with regard to output voltage stability at any particular voltage setting:

* *The Plessey Co. Ltd, formerly Standard Telephones and Cables Limited.*

(1) The balanced twin triode arrangement reduces the effects of valve ageing and heater voltage variation.

(2) Current through the stabilizer or reference tube V_3 is constant under all conditions. Changes of the output voltage setting are therefore not followed by slow drift of the output voltage, such as may occur if the current through V_3 is also changed.

3D21A Series Regulator Circuit

The circuit which has been used with the 3D21A is shown in Fig. 3. It is a particular design which gives the following performance:

Input voltage	3·5kV
Output voltage	800V to 2·7kV
Output current as in Fig. 4	16mA at 2·7kV
*Output impedance	25Ω
*Stabilization ratio (small signal)	$1{\cdot}8\ 10^4$

* *Output voltage 2·5kV, load current 10mA.*

The feedback amplifier is modified to a cascode circuit, by using one half of the double triode used for regulating the series valve, screen supply. Considerable reduction of output impedance and increase of the stabilization ratio thereby result from the increased gain. There is additionally practically no loss of the characteristics of (1) and (2). The shunt regulator valve V_{4a} is stabilized by the voltage across R_4, which is in turn stabilized by the voltage across V_3. Such an arrangement does not give quite such a well regulated screen supply as would be obtained by using an additional stabilizer valve, as in Fig. 1.

Resistor R_{14} is required to limit the grid current of V_{3a}, during the interval when V_i has been applied, but V_3 has not ionized, and to ensure that sufficient voltage will be available at the anode of V_3 for ionization to take place. It additionally serves with capacitor C_1 as a filter for stabilizer noise. Capacitor C_2 improves the high frequency response of the feedback amplifier. Capacitor C_3 provides a low impedance return path for signals which appear on the h.t. line, above the amplifier cut-off frequency.

Increased output voltage may be obtained by increasing the applied voltage. It is, however, necessary to realize that, with, say, 4·5kV input and 800V output, a much larger current will be conducted by V_{4a} than when only 3·5kV is applied. In designing this type of supply the current carried by this valve and by the screen grid of V_1 must be carefully considered at the extremes of output voltage. For this purpose the general form of the screen grid and control grid characteristics of the 3D21A are given in Fig. 5.

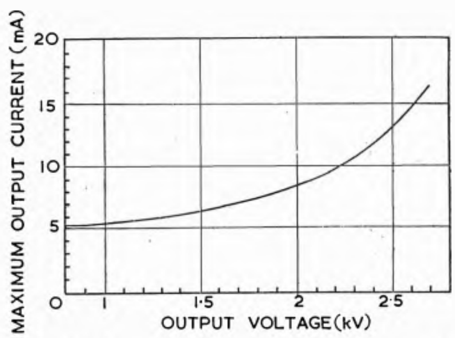


Fig. 4. Maximum output current versus output voltage

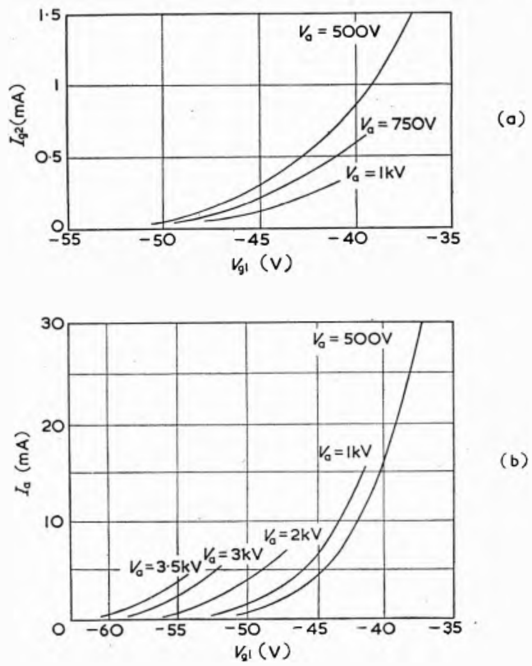
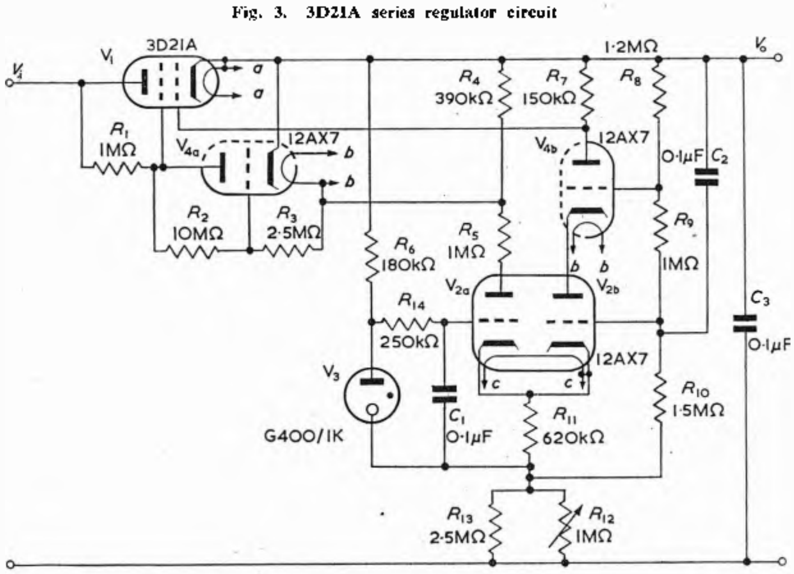


Fig. 5. Screen grid and control grid characteristic of 3D21A.
Screen grid voltage 400V

Delayed switching of V_1 would be advisable with this circuit; it may be obtained by thermal delay devices. It should also be noticed that three heater supplies are indicated, but this requirement may be reduced to two if the trustworthy valve type 6057 is used in place of the equivalent 12AX7 valves.

Fig. 3. 3D21A series regulator circuit



Conclusion

In conclusion some formulae are given for assessing the performance of this type of power supply. The small signal stabilization ratio of a circuit where the series valve is shunted by a resistor R is:

$$\frac{R}{R + r_a}\ [(r_a/R_L) + \mu\beta A + (r_a/R) + 1]$$

where r_a = series valve anode impedance

μ = series valve amplification factor

βA = feedback amplifier circuit overall gain.

The output impedance is:

$$\frac{r_a}{1 + \mu\beta A}$$

The cascode voltage gain A, using identical triodes is:

$$\frac{\mu' R_L'}{(\mu' + 2)r_a' + R_L'}(\mu' + k + 1)$$

where μ' = feedback amplifier valve amplification factor

r_a' = feedback amplifier valve anode impedance

R_L' = feedback amplifier valve load resistance

For the circuit of Fig. 3, $\beta = \dfrac{R_{10}}{R_8 + R_9 + R_{10}}$

$$k = 1 + R_9/R_{10}$$

Acknowledgements

The author would like to express his thanks to Standard Telephones and Cables Limited, for permission to publish this article.

A Versatile Multi-Frequency Network Analyser

Some five years ago, the Electrical Research Association formulated its plans for a network analyser to be built at its Research Laboratory at Leatherhead, Surrey.

The association has, over a long period, been actively concerned with many studies in electrical engineering such as the flow of harmonics in power systems and rates of rise of restriking voltages and these problems were among the many that were taken into consideration when the design of the analyser was considered. It was recognized too that internal research would not occupy the analyser full time and in order to make the analyser available to individual members of the association a flexible and versatile machine would be required.

An electronic network analyser meeting these requirements has now been constructed and is in full operation.

Its principal parameters are as follows:

Impedance 100Ω (10mH and 1μF)
Power 2·5mW (500mV and 5mA)
Frequency 1·592kc/s (ω 10^4rad/sec)
Frequency range 0·159 to 15·9kc/s ($\omega = 10^3$ to 10^5rad/sec)

The resolution and repeatability of the measuring equipment is as high as 0·1 per cent of base amplitude and 0·1° of angle for amplitudes greater than 10 per cent.

The impedance and power level chosen results in a signal of 100μV for the metering of a current at 100 per cent level. The desired amplitude resolution calls therefore for a sensitivity of 0·1μV in the current measuring equipment, while the phase resolution requires sensitivity of 0·016μV.

The complete design comprises eight wings of racks housing the network elements which are connected to four interconnecting units or plugboards and the central measuring equipments for the steady state and transient studies respectively which are placed in two control desks at opposite sides of the array. All the network elements are available for use in conjunction with either control desk on large problems.

However, the wings of racks and the plugboards may be suitably apportioned between the two desks, so that studies may be carried out simultaneously on two independent networks, thus giving a high utilization factor for the equipment.

Passive Network Elements

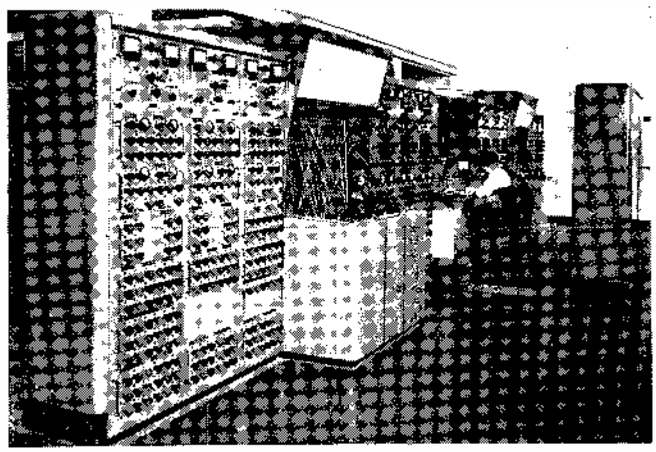
The passive network elements consist of two-terminal elements, composed of decade impedances in simple series or parallel arrangements, and three- or four-terminal elements such as π networks and transformers, together with some electronic amplifiers. The terminals of each two-terminal element are brought out and identified by colour (red or black) at the ends of flexible conductors at a plugboard. Three- or four-terminal units are treated as pairs of two-terminal units. Each unit has also four terminals available for the purpose of metering, i.e., two voltage terminals and two current terminals. The latter are the potential terminals of a 20mΩ shunt incorporated in the read lead of each unit.

Active Network Elements (Generators)

Eighteen generators of a conventional design have been constructed to act as steady state power sources. These 'generators' are in fact adjustable amplifiers, with an independent phase and amplitude control in each unit. The output amplifier, which ensures a low output impedance of 0·1Ω, has its output isolated from earth. Provision has been made for the insertion of built-in adjustable reactances and power and terminal voltage monitors, which will result in a considerable economy of time required to achieve the initial balance in load flow studies.

Central Measuring Equipment for Steady-State Studies

The main measuring system consists of three automatic a.c. potentiometers operating simultaneously to indicate the potential difference V (amplitude and phase) between the selected points in the network, the current I through the selected unit (amplitude and phase), the direct phase difference ϕ between V and I and the products $VI \cos \phi$ and $VI \sin \phi$. The last give the active and reactive load (watts P and V Ars Q) when V and I refer to voltage across and current through a unit or network. An independent direct voltmeter is also provided, mainly for the purpose of quick checks and for monitoring a voltage across one unit while adjustments may be made to other units.



The E.R.A. network analyser

The reference signal for the bridge measurement is derived from the same three-phase master oscillator that provides excitation for the generator units. The measurement of voltage (or current) is made by comparing the selected voltage signal from the network nodes (or from a shunt) with a reference signal derived from a potentiometer. The difference of these two voltages operates a servo-system which adjusts the slider position on the potentiometer and the phase of the signal supplied to the potentiometer until a balance is obtained.

The shafts of the potentiometers and phase shifters are coupled to dials on the control desk so that their angular positions display current and voltage in amplitude and in phase relative to a reference signal, together with the phase-angle between them.

The measurement of active and reactive load, which is usually difficult in variable frequency systems, has been made fairly easy by the use of an analogue method. The multiplying operations required are performed on a fixed frequency analogue signal by means of potentiometers and a sine-cosine resolver ganged with the voltage and current bridges. The accuracy of these operations is governed by the linearity of the amplitude potentiometers (which is 0·1 per cent) and the accuracy of the sine-cosine potentiometer which may introduce errors of 0·2 per cent.

The direct measurement of voltage is made by means of an amplifier-operated rectifier instrument. Amplitude only is measured, but phase-angle may be indicated on one of the monitoring cathode-ray tubes. Three such tubes are provided for permanent display of voltage waveforms at important points in the measuring system which can be selected at will, thus facilitating adjustment and giving indication of faults.

A System for
Providing a Precise Vector Voltage

By D. J. Collins*, B.Sc., A.M.I.E.E., and J. E. Smith*, B.Sc.

In many fields of electronics there is often a requirement to produce a voltage of known phase and amplitude relative to some reference level. If one neglects electro-mechanical devices, the accuracy to which this can be done is largely dependent upon the accuracies of the components of the phase shifter. The article deals with a possible system for producing a precise vector voltage which can be of any amplitude less than reference and of a phase relationship between 0° and 360°. Such a system is of great use in the calibration of phase shifters and phase measuring devices.

(Voir page 707 pour le résumé en français: Zusammenfassung in deutscher Sprache Seite 712)

IN the development of phase measuring equipment, it is quite often necessary to obtain a sinusoidal voltage of precisely known amplitude and phase relative to an initial reference voltage.

This can be done without undue difficulty, if it is required only at one precisely known frequency. However, if the voltage is required at a frequency not precisely known or over a band of frequencies, the problem is more difficult. The system detailed in this article indicates a method of tackling the problem over quite a wide frequency range.

The system consists of two stages. In the first stage, four vector voltages are established with phase relationships of 0°, 90°, 180° and 270°. The second stage utilizes these four vector voltages to select the precise vector voltage required.

Reference Vector Voltages

The basic reference voltage at the required frequency is conveniently obtained from a laboratory oscillator. It is necessary for the oscillator to provide a voltage with low distortion and adequate stability. A secondary requirement is that the voltage level shall be greater than the amplitude of the required voltage. This second requirement is not of prime importance, since the oscillator output can be suitably amplified. For the purpose of description it will be accepted that a 10V r.m.s. level is adequate, and this basic 10V reference will be termed the 0° phase.

The system now requires the generation of other reference vectors, but these are limited to vectors of 10V amplitude at the other cardinal points, i.e. 90°, 180° and 270°.

The 180° phase presents no problem, and at this level a simple image or -1 amplifier (previously described[1]) can be used. The quadrature phases are not so easily obtained, particularly if the frequency is not accurately known or a range of frequencies have to be covered. In this system, the basic integrator is used to provide the 90° phase. The circuit is shown schematically in Fig. 1, and it can be proved that if the gain M is greater than 10^8 and if $R \sim X_c$ at the chosen frequency, the angular error is of the order of 0·1°. If $R = X_c$, then the amplitude transmission is unity.

Fig. 2 shows how, by the use of the two operators units -1 and j, any pair of adjacent cardinal points may be derived from the 0° phase.

Vector Selection

Each cardinal pair embraces a quadrant and, for a certain vector requirement, the reference voltage configuration which covers the appropriate quadrant should be chosen. For convenience of analysis, it is assumed that the 0° to 90° quadrant is selected. There are thus available two reference voltages, E and jE where $E = 10$. Consider these voltages applied to a resistance network as shown in Fig. 3, and let the output voltage E_o be the required vector $A\angle\phi$.

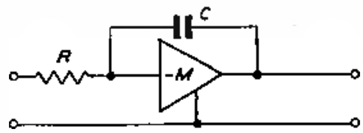


Fig. 1. Basic integrator

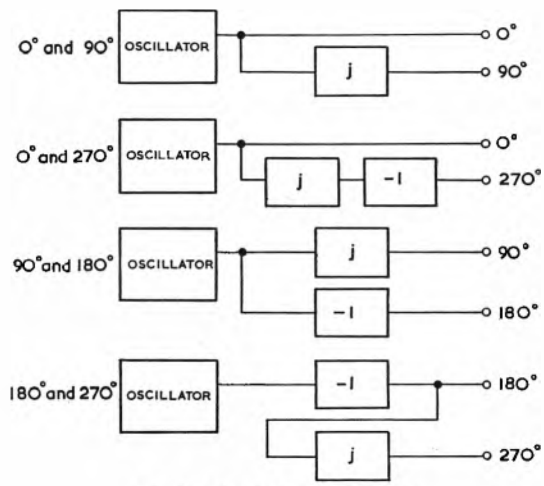


Fig. 2. Cardinal pair generation

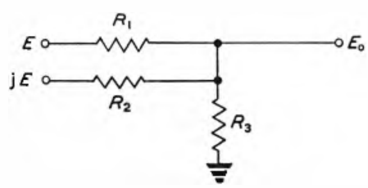


Fig. 3. Resistive phase-shift network

It can be shown that

$$A = \frac{R_3\sqrt{(R_1^2+R_2^2)}}{R_1R_2+R_1R_3+R_2R_3} \text{ and } \phi = \tan^{-1} R_1/R_2$$

Thus by suitable choice of the three resistor values, the desired vector can be obtained, and it is obvious that the vector parameters are unaffected over a wide frequency range.

Since it is likely that some load may be applied to this

* *Solartron Research & Development Ltd.*

vector source, a buffer amplifier can be interposed and R_3 can conveniently form the input resistance of the amplifier. R_1 and R_2 are then more suitable as decade resistance boxes.

The resistive phase shifter of Fig 3 can be utilized in more convenient form by arranging calculated networks at, say, 10° steps, and selecting the required network by a uniselector system coupled to a digital indicator. Use is made of this resistive phase shifter in the commercial instrument shown in Fig. 4.

Fig. 4. Instrument for utilizing the phase-shift network of Fig. 3.

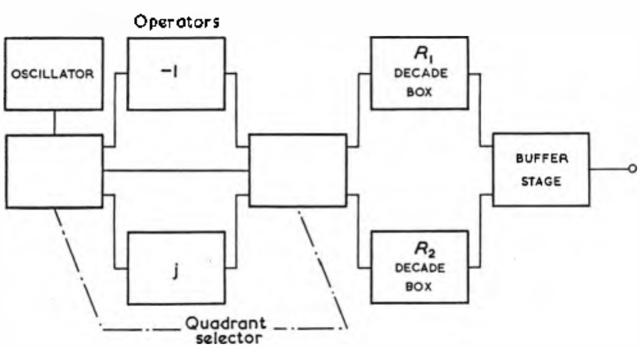


Fig. 5. A complete system

Practical Considerations

A complete system is shown schematically in Fig. 5, and a system of this type has been used in the laboratory to cover a frequency range of 1c/s to 1kc/s. With modification, it would be possible to extend the frequency range in both directions.

The oscillator, apart from having the required voltage output, should also be capable of driving the system load without increase in distortion. Its output impedance will be required to be low, usually less than 10Ω.

The '−1' operator amplifier should have reasonably high input impedance, at least 10kΩ, and again an output impedance of less than 10Ω. It should have negligible phase error, less than 0·1°, over the frequency range required, and should be capable of handling the basic voltage level. Its gain should be capable of slight adjustment about the unity condition.

The 'j' operator integrator should have the same order of input and output impedance as the '−1' unit. Its phase error can be made less than 0·1°, but its amplitude transmission is only unity if $R = X_c$. To maintain this condition over a wide frequency range will require a change of the value of C and possibly variations in R. It is thus convenient, when operating over a wide frequency range, to select spot frequencies and switch suitable CR values for these frequencies. An important point to note is that the integrator is liable to have a long term d.c. level drift due to grid current, and it may be necessary to include a 're-set' switch to discharge the capacitor.

The quadrant selector is merely a switch system which selects the appropriate vectors, and it may be convenient to incorporate on this switch monitor points at which the basic amplitude can be set. The level can be set by using an a.c./d.c. digital voltmeter and adjusting the oscillator output and the fine controls on the operator units. The reference voltage vectors will thus be obtainable at an amplitude accuracy of the order of 0·1 per cent and an angular accuracy of better than 0·2°.

The voltage selector system, comprising the decade boxes R_1 and R_2 and the buffer amplifier, will satisfy most requirements, but it may be more convenient to choose resistor pairs for R_1, R_2 at selected angles. In addition, at high frequencies, variation of stray capacitances in the decade boxes will rule them out. The buffer amplifier will be required to have negligible phase error, in addition to its dynamic characteristics. The voltage selection system, assuming a perfect reference input, can have errors less than 0·1 per cent amplitude and 0·1° angularity.

In the laboratory, it has been possible to produce, with this type of system, vectors to a phase accuracy of 0·2° and 0·2 per cent amplitude accuracy.

Acknowledgment

The authors wish to record their thanks to the Directors of Solartron Research & Development Ltd for permission to publish this article.

Appendix

(a) A Limited Check of Integrator Error

A phase-sensitive voltmeter can be used in the following manner to indicate any large error in the angular accuracy of the integrator.

Using the 0° phase as reference, the 90° phase as signal, it is possible from the meter indications to calculate an error angle ϕ_1. Then reversing the connexions, using the 90° phase as reference, the 0° phase as signal, a new error angle ϕ_2 is calculated.

The integrator error is then $\frac{1}{2}(\phi_1 + \phi_2)$

If the voltmeter has no error, ϕ_1 will equal ϕ_2. If, however, it has error, the magnitude is given by:—

$$\tfrac{1}{2}(\phi_1 - \phi_2)$$

This method, however, can only give indication of error if it exceeds 0·5°. It can only be used to ensure that there has been no gross error in the design or construction of the integrator.

(b) Use of the Integrator as a 'Backing Off' Source

Reference 1 describes a method of using a phase sensitive voltmeter to measure small errors in nominal 0° and 180° phase relationships. It utilizes a −1 or image amplifier as a source of 'backing off' voltage. The integrator provides a source of quadrature 'backing off' which can be used in measuring errors in nominal 90° and 270° phases by the same method.

By the same token, if it were possible to derive from the basic 0° phase a 270° phase to an accuracy of, say, 0·01°, the system could be used for checking the integrator itself!

References

1. Collins, D. J., Smith, J. E. Measurement of Small Phase Shifts with a Phase Sensitive Voltmeter. *Electronic Engng.* 30, 146 (1958).
2. Collins, D. J., Smith, J. E. Some Methods of Phase Measurement used in Transfer Function Analysis. *Electronic Engng.* 30, 182 (1958).

An Alternative Method for Comparing the Electrical Properties of Epoxy Casting Resins

By D. H. Thompson*

This article describes a method of assessing the dielectric constant, loss, and volume resistivity of epoxy casting resins over a wide temperature range and for frequencies up to 5Mc/s.

This coverage is achieved at the expense of reduced absolute accuracy when compared with the standard method of Hartshorn and Ward; however, when primarily used for comparison of resins and assessing the temperature stability of their electrical parameters, the method described here merits more widespread use.

In particular, absolute values may be quoted from this method provided care is taken to verify or normalize the results at laboratory temperature.

(Voir page 707 pour le résumé en français: Zusammenfassung in deutscher Sprache Seite 712)

It will be common experience in the electronic and electrical industry that the specialized information relating to the electrical properties of epoxy casting resins is seldom available from the manufacturers. This is understandable since the variety and nature of measurements ultimately required do not fall within the scope of resin manufacturers. Thus, it is felt that the details of methods developed for the speedy accumulation of electrical data, may be of interest to others in the electronic industry.

The problem arose of selecting an epoxy resin-filler combination possessing particularly stable electrical characteristics for a circuit encapsulation application. The selection of a suitable system involved measurements of a range of 60 resin-filler-hardener systems.

Two methods were available, one based on that of Hartshorn and Ward, utilizing a jig supplied by Marconi Instruments Ltd for use with their TF329G circuit magnification X meter (Q meter); the other, to use resin cast trimmers as test-pieces and the same Q meter.

Discussion of Merits of Available Apparatus

The Marconi jig was not readily adaptable or suitable for use at other than laboratory temperature, a disadvantage common to its parent method; difficulty also exists in allowing for edge effects.

The method using cast trimmers may be applied with care over a wide temperature range. In fact, some resins have been tested from $-60°C$ to $+140°C$ by this method. The edge effects are no longer serious but some doubt exists concerning any strains set up by the contraction of the resin during cure, which may leave the trimmer somewhat distorted, or the resin stressed. It was known that very high pressures tend to increase the dielectric constant of dielectrics; however, an initial survey showed such effects, if present, to be small in the test sample and errors in excess of one or two per cent. unlikely.

Preparation of Trimmer

Certain precautions were necessary in the preparation of the jig; these are best described in detail. A 100pF air dielectric, ceramic mounted, pre-set trimmer capacitor was used (Fig. 1(a)).

Stiff leads of 20 s.w.g. tinned copper wire were soldered to the free lug of the rotor and to one of the solder 'blobs' retaining the stator. To prevent surface effects, sleeves of high resistivity were placed over these leads to extend about 1cm beyond the centre boss of the capacitor (Fig. 1(b)). P.T.F.E. was used, as this material has high resistivity (volume resistivity approximately $10^{20}\,\Omega$-cm), low surface contamination and can withstand high temperatures. The capacitor was next degreased with re-distilled carbon tetrachloride and dried, finally, at 105°C.

Prior to the casting, the capacitor was hung by its leads in a suitable mould, typically a 2in diameter impact extruded aluminium can prepared with a release agent, and resin poured in to a level just covering the boss. Entrapped air was removed by placing the mould in a vacuum chamber and reducing the pressure to approximately 0·1mm Hg. Vacuum filling is, of course, to be preferred. The method cited here is intended to simplify apparatus requirements and is satisfactory if sufficient care is taken. After curing to the manufacturer's schedule and removing from the mould, a hole $\frac{1}{16}$in diameter was drilled $\frac{1}{4}$in deep, axially, in the boss of the trimmer, to accept a thermocouple junction (Fig. 1(c)).

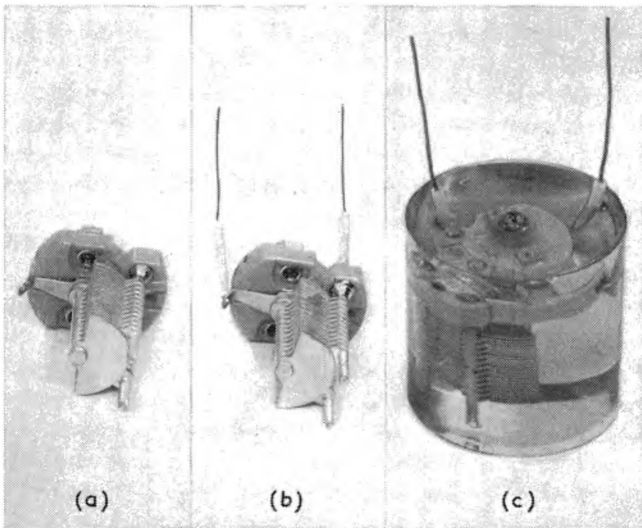


Fig. 1. Dielectric trimmer used for the measurement of electrical properties of epoxy resin systems

(a) As received
(b) Showing p.t.f.e. sleeved leads
(c) Potted in unfilled epoxy resin systems

Method of Using Trimmer

It is necessary to modify a suitable, thermostatically controlled, oven such that p.t.f.e. bushes are fitted through its walls to allow the capacitor leads to pass directly to a Q meter. These leads must be of the absolute minimum length, and must be standardized.

* *E.M.I. Electronics Ltd.*

With the capacitor thus mounted and using either the standard Marconi inductors or laboratory versions made to tune to desired frequencies, the capacitance of the jig may now be measured at any temperature within the range covered by the oven. Provided the rotor of the potted capacitor is connected to the Q meter 'earthy' capacitance terminal, the thermocouple may be left in situ when readings are taken.

A small, low temperature, housing was made to contain the specimen in a close-fitting can surrounded by a mixture of ethane diol, methylated spirits, and solid carbon dioxide. It is important to maintain the same lead length as before.

The method of measuring the capacitance with a Q meter is as described in the instruction manual, but above 2Mc/s calibration by substitution was necessary. The dielectric loss of the capacitor may be found directly, in terms of shunt resistances, by direct substitution. It is only necessary to plot graphs of Q against shunt resistance for each frequency used, and of capacitance against the capacitance dial setting for each frequency above 2Mc/s. Direct substitution is not necessary for each operation.

For measurements at higher frequencies only, the use of a smaller trimmer would be advisable.

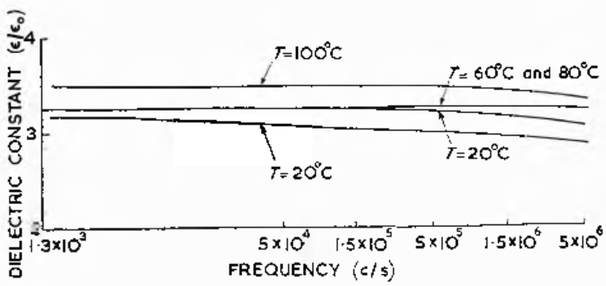


Fig. 2. Variation of dielectric constant with frequency and temperature for an amine-cured epoxy resin system

The capacitance was measured at audio frequency using a capacitance bridge; the resistance was measured at 500V d.c. The Cintel, wide-range bridge proved suitable in this application. Both measurements are straightforward but the use of p.t.f.e. insulation and a guard system for the ohmmeter is advisable.

Interpretation of Results

The dielectric constant K is given by :—

$$K = C_2/C_1$$

where C_1 = capacitance of unpotted trimmer.

C_2 = capacitance of potted trimmer.

The volume resistivity ρ_{dc} is given by :—

$$\rho_{dc} = C_1 R/\varepsilon_0$$

C_1 = capacitance in farads.

R = resistance of the trimmer in ohms.

$\varepsilon_0 = 1/36\ \pi \times 10^{-9}$ farads/metre.

ρ_{dc} = volume resistivity in ohm-meters

The above equation can be used to obtain a parameter which is the a.c. volume resistivity at pulsatance ω and which may be used to derive stray shunt resistances, e.g., given by the expression

$$R = \varepsilon_0 \rho\omega/C$$

The above equation expresses the shunt resistance introduced when a unit having a stray capacitance of C farads (measured before potting) is potted, provided this stray capacitance were in air. Thus, in general, a high value of $\rho\omega$ is preferable to a low loss tangent as the latter is also a function of K for the resin.

Accuracy

The accuracy will not be considered here in detail, but estimates made have shown the probable error not to exceed ±5 per cent for dielectric constant and ±10 per cent for volume resistivity. Typical electrical characteristics of an epoxy resin system measured by the techniques described are shown in Figs. 2 and 3.

Conclusion

This method of using trimmers has been successfully employed to evaluate sixty resin systems. These results have been compared to those obtained using the Marconi jig mentioned, with satisfactory agreement.

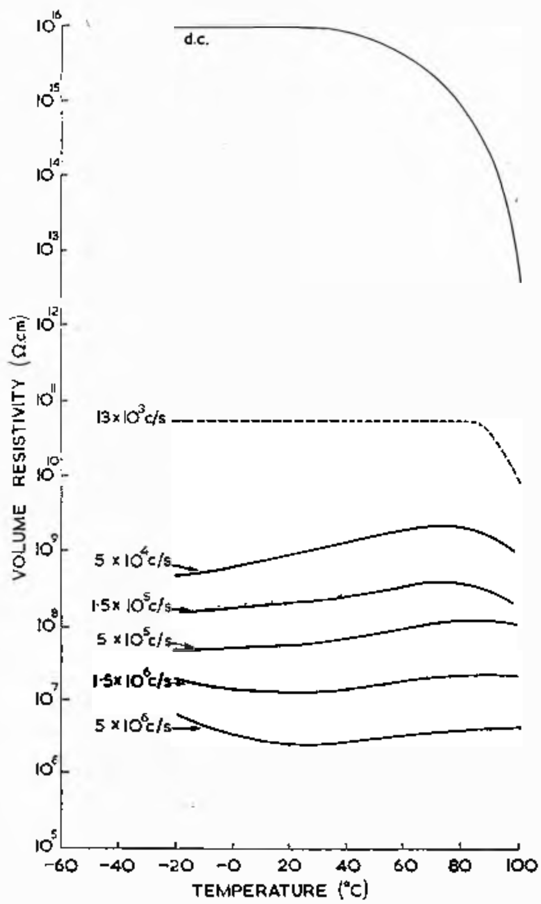


Fig. 3. Variation of d.c. and r.f. volume resistivity with temperature for amine-cured epoxy resin system

This jig is not original but it is not widely known and when known, its accuracy is suspected. The justification for this article is to attempt to remove some of this suspicion and establish the validity of the method for practical industrial purposes when comparative results are of primary importance. In these circumstances, this method can be used to save labour and obtain results more speedily than the standard method.

Acknowledgment

Thanks are due to the heads of departments involved and Company officials for permission and encouragement in publishing this article.

A Transistorized L.T. Regulator

By J. H. Deichen*

The methods of providing a stabilized d.c. supply for valve heaters are examined and their limitations are discussed. A simple circuit using a transistor and two Zener diodes is then given and its design and performance described.

(Voir page 707 pour le résumé en français: Zusammenfassung in deutscher Sprache Seite 712)

FOR a specific problem in hand, it was necessary to regulate the heater voltage of a signal generator oscillator valve and of a d.c. amplifier for an automatic level control circuit. The two valves were rated at 400mA and 200mA at 6·3V respectively. A number of previously published circuits were investigated but most of them were found to be deficient in one way or another.

The constant voltage transformer is undesirable as it is frequency conscious and too bulky to fit into a portable instrument.

The back-to-back Zener diode method as described by Kezer and Aronson[1] regulates only the peak voltage and not the r.m.s. value. It is only effective if a large part of the waveform is cut off, thus giving low efficiency. Like the single Zener diode for regulating d.c. as described by Pereli[2] the regulation is only effective for relatively small currents. Any large variations in the Zener diode current alters the diode voltage. Therefore, for good regulation, the current variation must be small and the impedance of the load must be large compared with the effective resistance of the diode.

The transductor[3,4] has the disadvantage of being bulky and heavy and is unsuitable for installing into small instruments. It is a constant average current device and to achieve constant r.m.s. regulation a compensating unregulated supply must be available for part of the control voltage, and this supply must not be derived from a source supplying a varying load. R.M.S. stability could not be achieved by the smoothing method described by Smith[3] as the input capacitor of the filter would charge to the peak voltage, which becomes greater with mains increase due to the peak current waveform of the a.c. fed to the rectifiers. Stability could, however, be obtained by a choke input filter but the choke must be very large so that its impedance is high compared to the output capacitor impedance. The large choke renders the system unsuitable for the application under discussion. A further disadvantage of the transductor when feeding more than one valve is that, being a current regulator, if one valve fails or is removed the other valve or valves are forced to take higher current.

A method investigated by the author was the use of a multivibrator, the output of which was fed via a centre-tapped, step down transformer to a rectifier and a smoothing circuit and thence to the heaters of the valves to be regulated. The stability proved to be very good but all the heater load plus the losses were derived from a stabilized high tension supply. This, of course, rendered the system expensive as a large high tension regulator was required and the multivibrator valve took a heater current as great as the regulated low tension required.

Transistorized Regulator

The transistorized regulator which has a similar circuit arrangement to the conventional high tension regulator but without a d.c. amplifier stage was investigated and adapted for the instrument in question. The circuit is, in fact, an emitter-follower and is shown in Fig. 3. The principle of operation can best be shown by starting with the circuit of Fig. 1. As this is a cathode-follower its cathode voltage is largely independent of the applied anode voltage. The regulation is dependent upon the output impedance for load variation and upon the gain for variations in input voltage. The improvement in voltage variation against input change for this circuit is not sufficient

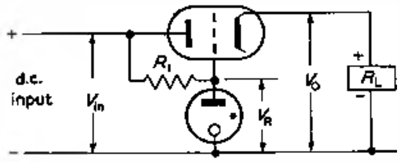


Fig. 1. A simple valve regulator

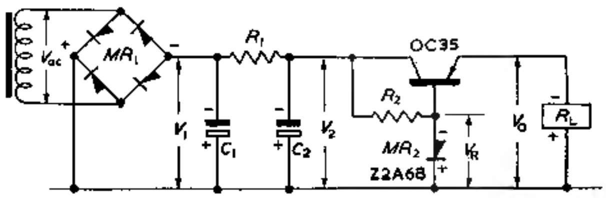


Fig. 2. A complete transistorized regulator, 10 to 1 improvement

for most applications. Therefore, to lower the effective output impedance and to increase the gain of the system a d.c. amplifier is added between the reference neon and the valve grid. The circuit shown in Fig. 2 is an equivalent circuit using a transistor and Zener diode instead of the valve and neon. The theory of this circuit is given by Brown and Stephenson[5] and is similar to the cathode-follower, but unlike the cathode-follower it caters for low voltage and high current. Very often for low tension supplies the degree of regulation can be lower than for high tension and, therefore, a d.c. amplifier is not necessary. The circuit in Fig. 2 does, in fact, give a 10 to 1 voltage change improvement at 600mA for 6·3V heaters. Referring to Fig. 2 the resistor, R_2, is designed to supply the correct nominal current through the Zener diode, MR_2, and for this particular design the nominal current was set at 25mA. The base-emitter current of approximately 15mA is also obtained via resistor R_2. Components C_1, C_2, and R_1 constitute the smoothing circuit for the bridge rectifier MR_1. The hum level obtained was less than 5mV r.m.s. and the power dissipated in the transistor was less than 3W at nominal input voltage.

To obtain greater stability a d.c. amplifier can be added between the Zener diode and the transistor base. However, a somewhat easier method, although with not as great an improvement in stability, is the use of two Zener diodes in cascade as shown in Fig. 3. The stability of a regulator depends on the stability of the reference voltage and, in fact, using the circuit of Fig. 2, the reference voltage does

* *Marconi Instruments Ltd.*

alter with mains voltage change due to the large change in current through the reference diode. The circuit in Fig. 3 uses a second Zener diode, MR_3, to regulate the current through MR_2. This circuit gives a voltage improvement of 30 to 1. Referring to Fig. 3, the components C_1, C_2 and R_3 constitute the smoothing circuit, giving a hum level of less than 1mV r.m.s. across the load, R_L. The value of resistor R_3 is selected to give a nominal current of 35mA through diode MR_3. As stated before, the load consisted of two valve heaters, one taking 400mA and the other 200mA. Removing the 400mA valve the voltage across the other rose by 0·1V. On the other hand, by making a 20 per cent increase in load the voltage dropped by 0·05V. It was found that the voltage drop between the base and emitter of the transistor was 0·3V, therefore the load voltage is 0·3 less than the reference. Using a 5 per cent tolerance reference diode of 6·8V nominal the load voltage will lie between 6·16 and 6·84. For many valve heaters this tolerance is permissible, therefore an adjustment for voltage level is not necessary. If a closer value must be obtained then a suitable resistor can be added in series with the heaters. The circuit does not provide for temperature compensation and does, in fact, vary by about +0·15 per cent/°C rise. The temperature coefficient of the Zener diode used is given at +0·04 per cent/°C, but by using two lower voltage diodes in series a negative coefficient of approximately −0·06 per cent/°C is realized. Therefore the temperature coefficient of the circuit would be approximately +0·05 per cent/°C rise.

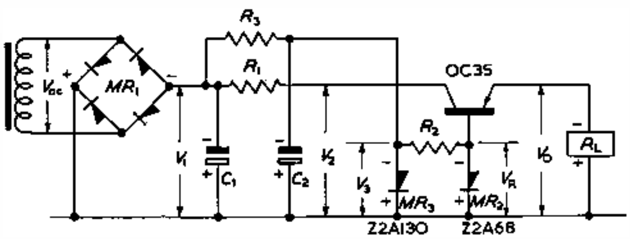


Fig. 3. A complete transistorized regulator with two Zener diodes in cascade, 30 to 1 improvement

Design Considerations

The design considerations are very simple, the most important being to not exceed the transistor dissipation. If greater load current is desired the transistor can be substituted by one of greater rating and they can, of course, be used in parallel. The collector voltage must be sufficiently high for regulation down to the lowest input visualized and the author found a value of 4·5 to 5V above the emitter voltage to be sufficient for ±10 per cent change in mains voltage. For obtaining the resistor values for Fig. 3 the following formulae apply:

$$R_1 = \frac{V_1 - V_2}{I_L}$$

where I_L is the load current.

$$R_2 = \frac{V_3 - V_R}{i_R + i_{be}}$$

where i_R is the nominal current through the reference diode and i_{be} is the base-emitter current.

$$R_3 = \frac{V_1 - V_3}{i_3 + i_R + i_{be}}$$

where i_3 is the normal current through the Zener diode MR_3.

The calculation of resistor values for Fig. 2 is similar. Component and voltage values for Fig. 3 are as follows:

V_{ac}—15V r.m.s.	R_1—8·5Ω
V_1—16·5V	R_2—150Ω
V_2—11·5V	R_3—33Ω
V_3—13V ±5 per cent	C_1—1 000μF
V_4—6·8V ±5 per cent	C_2—500μF
V_o—6·5V ±5 per cent	

It should be noted the polarity is normally negative with respect to earth.

Acknowledgments

The author wishes to thank the management of Marconi Instruments Ltd. for permission to publish this article and to Mr. J. Golding for his helpful suggestions.

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Extensions to Nucleonic Instrument Factory

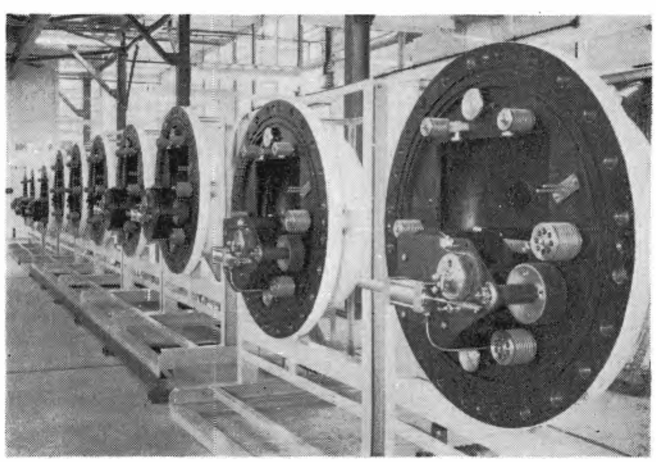
The extensions opened recently by Sir John Cockcroft have nearly doubled the factory area of Plessey Nucleonics Ltd of Northampton.

The main programme of the Company has been the development of a system for the detection of failures in fuel element cans in the reactors of nuclear power stations. To this range of burst slug detection equipment has been added flux scanning equipment, radiation detectors, process control and health instrumentation.

The Company's Burst Slug Detectors were installed at Calder Hall and Chapelcross and improved types are now being assembled for installation at Hinkley Point and elsewhere.

Among the new instruments developed at Northampton is a low energy gamma ray monitor which has application as a replacement for alpha monitors for the detection and location of fissile materials such as ^{239}Pu and ^{235}U. Other radiation monitors include types developed for food, milk and water monitoring

Ten burst slug detection precipitator units, manufactured by Plessey Nucleonics Limited for the nuclear power station at Hinkley Point, here being assembled in the clean area (dust free) of the Northampton factory.



THERMOELECTRIC COOLING*

RECENT developments in thermoelectric cooling are the results of applying new semiconductor materials, evolved in studies of solid state physics, to certain principles discovered in the classical physics of the early nineteenth century.

Seebeck observed, in 1821, that a potential difference develops when the junction of two dissimilar conductors is heated. He misunderstood the meaning of his results, but his arrangement of materials in a thermoelectric series is substantially the same as that recognized today.

A second thermoelectric effect was discovered by Peltier in 1834. He showed that heat is absorbed or generated when a current passes across the junction between two dissimilar conductors, and the direction of flow of the current determines whether the junction becomes cold or hot. This effect is distinct from the normal Joule resistive heating effect, but is superimposed upon it. Again the results were misinterpreted and it was thought that Joule's law did not apply when the current was small.

Thomson (later Lord Kelvin) sought to define a relationship between the Seebeck and Peltier effects in terms of thermodynamic theory. He concluded that a third thermoelectric effect existed and that there is heating or cooling in a single conductor, along which there is a temperature gradient, when a current is passed through it. This is known as the Thomson effect.

In 1838 Lenz succeeded in freezing water by passing current through a bismuth antimony junction, the ice being melted when the current was reversed. It is possible that the success of his experiment owed something to the fact that it was carried out in the low ambient temperature of a St. Petersburg winter.

Very little practical application was found for these discoveries but the Seebeck effect has long been used in thermocouples for temperature measurement and in thermopiles for radiation detection. In both these applications high sensitivity compensates for low efficiency but, until recently, even inefficient thermoelectric refrigeration based on the Peltier effect was impossible with the materials then available.

However, a basic theory of thermoelectric refrigeration was derived by Altenkirch in 1911. He was able to express the coefficient of performance in terms of a figure of merit calculated from the thermal e.m.f. coefficient and the thermal and electrical conductivities of the materials of the thermocouple. Practical refrigeration was not possible, however, until the new semiconductor thermocouples became available in very recent years.

Theory

It can be shown[1] that the maximum value of the performance ϕ is

$$\phi_{max} = \frac{T(\sqrt{(1+\theta^2)}-1)}{\Delta T(\sqrt{(1+\theta^2)}+1)} - \tfrac{1}{2} \quad \ldots\ldots\ (1)$$

where ΔT is the temperature difference ($T_H - T_C$) between the hot and cold junctions

T is the mean temperature, $\frac{1}{2}(T_H + T_C)$

and ϕ is the figure of merit of the thermocouple, given by

$$\phi = \frac{(\alpha_p - \alpha_n)T^{\frac{1}{2}}}{(\kappa_p/\sigma_p)^{\frac{1}{2}} + (\kappa_n/\sigma_n)^{\frac{1}{2}}} \quad \ldots\ldots\ (2)$$

and α is the thermal e.m.f. coefficient

κ represents thermal conductivity

σ represents electrical conductivity

and the subscripts p and n refer to the positive and negative elements of the thermocouple, which must have absolute thermal e.m.f. coefficients of opposite sign.

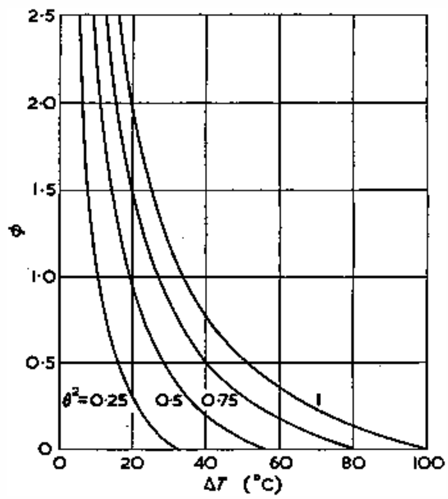


Fig. 1. Coefficient of performance plotted against temperature difference for a mean temperature of 290°K

As $\theta \to \infty$, ϕ tends to $(T - \frac{1}{2}\Delta T)/\Delta T$ which is the coefficient of performance of an ideal thermodynamic machine. By setting $\phi_{max} = 0$ the maximum temperature difference may be found from

$$\frac{\Delta T_{max}}{T} = 2\,\frac{\sqrt{(1+\theta^2)}-1}{\sqrt{(1+\theta^2)}+1} \quad \ldots\ldots\ (3)$$

Fig. 1 shows the variation of the coefficient of performance with temperature difference for various values of figure of merit. In Fig. 2 the maximum temperature difference is shown plotted against the figure of merit.

When the temperature difference is appreciable, considerable variations of α, σ and κ occur over the length of the couple. It is found to be desirable in making calculations to use the mean values of these three quantities over the range of temperature concerned.

Cascade arrangements of thermocouples may be used to obtain greater differences of cooling than are possible with a single stage. The first stage is then used to provide a cold sink for the hot junctions of a second stage, which

* *Communication from the G.E.C. Research Laboratories, Wembley.*

may, in similar fashion, provide an even colder sink for a third stage, and so on. For practical reasons it is considered advisable to limit the arrangement to two stages.

Semiconductor Thermocouples[2]

Equation (1) shows the high values of θ, the figure of merit, are required in the application of thermoelectric refrigeration, and Fig. 1 indicates that values of θ approaching unity are essential. In equation (2) it would be more convenient to have separate values of θ defined for each element in terms of

$$\theta_p = \alpha_p \sqrt{(\sigma_p T/\kappa_p)}$$
$$\theta_n = -\alpha_n \sqrt{(\sigma_n T/\kappa_n)}$$

If σ/κ is the same for each element, then θ is the arithmetic mean of θ_p and θ_n, and this is effectively true for the best thermocouples. The ratio σ/κ is substantially the same for all metals at a given temperature and the best choice for thermoelectric applications are those with the highest values of thermal e.m.f. coefficient. Such combinations are found

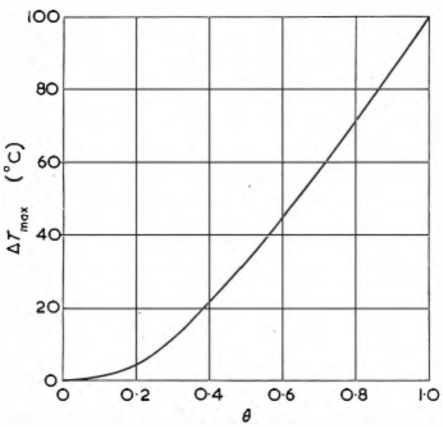


Fig. 2. Maximum temperature difference plotted against figure of merit at a mean temperature of 290°K

among the alloys of bismuth and antimony. In semiconductors the ratio σ/κ is usually much smaller than for metals but this apparent disadvantage, from the thermoelectric point of view, is often more than counterbalanced by a very much higher value of thermal e.m.f. coefficient, which may well rise to millivolts per °C instead of units or tens of microvolts per °C for metals.

Thermoelectric coefficients in semiconductors are high when the carrier concentration is low and low when the carrier concentration, and therefore the electrical conductivity, is high. There is thus a conflict between the requirements, for a high figure of merit θ, that both the thermal e.m.f. coefficient and the electrical conductivity should be high. Both these characteristics can be modified by varying the concentration of impurities in the semiconductor, and the optimum value of θ is obtained when the thermal e.m.f. coefficients are about $\pm 200\mu V/°C$. The figure of merit depends upon other factors which are capable of a degree of control, such as carrier mobility and the effective mass of the charge carriers, but prediction of the value of the functional combination of all the factors which defines the end result is complicated and difficult. In a given series of semiconductors, however, such as the Group IV elements, diamond, silicon and germanium, or the intermetallic compounds of Groups III and V, it is found that improvement in thermoelectric performance generally goes with rising atomic weight. The best compound semiconductors developed so far for thermoelectric applications are lead telluride and bismuth telluride, both formed from elements of high atomic weight. The best thermocouple combination of this type found so far is formed from p-type and n-type bismuth telluride, Bi_2Te_3, with an overall figure of merit θ of 0·76 and giving, at a mean temperature of 290°K (17°C), a maximum temperature difference obtained by Peltier cooling of 65°C.

If, however, the semiconductor compound is alloyed with an isomorphous material the thermal conductivity of the alloy may be reduced with no change in the carrier mobility, or at least only a relatively small reduction in this parameter. Such alloying may therefore improve the thermoelectric properties of the material. It has been found possible to form solid solutions between bismuth and antimony tellurides (Bi_2Te_3 and Sb_2Te_3), bismuth selenide (Bi_2Se_3) or bismuth sulphide (Bi_2S_3). Some of the best

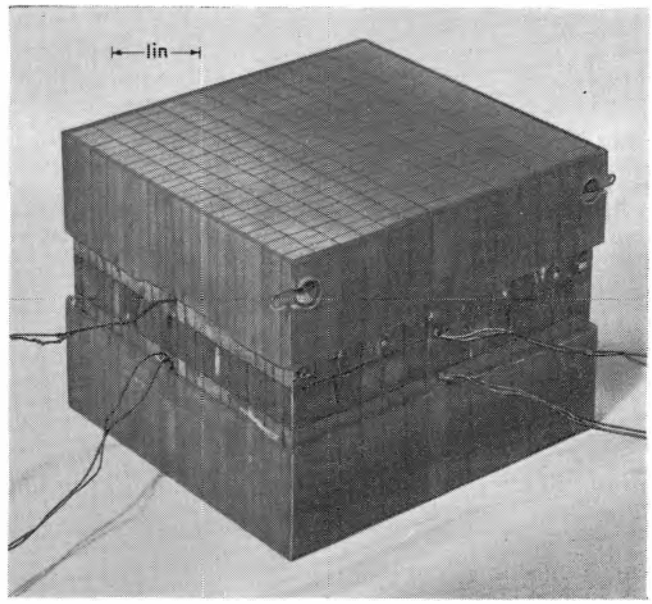


Fig. 3. Thermocouple elements soldered together and assembled in units

combinations of the resulting alloys made in the Research Laboratories of The General Electric Company at Wembley have given an overall figure of merit of 0·86 and a maximum temperature difference of 80°C at a mean temperature of 17°C. The improved results followed a reduction of thermal conductivity with no substantial change in the electrical conductivity or thermal e.m.f. coefficient.

Although the preparation of thermoelectric semiconductor materials is simpler than that of the preparation of semiconductors for transistors, the control of composition of the compounds, and alloys of compounds, required does present difficulties. Nevertheless, polycrystalline material is acceptable rather than the single crystals necessary for producing transistors and the impurity requirements are less stringent by a factor of 10^4 or 10^5. The thermocouple elements, cut from ingots or sintered from powder pressings, are soldered together and assembled in units such as those illustrated in Fig. 3.

Practical Applications

Although a temperature difference of 80°C is possible with the best semiconductor thermocouples it is achieved only when the unit is doing no work. When operating with reasonable efficiency, differences of temperature across the thermocouple of some 30 to 50°C are obtained. The efficiency of operation increases with smaller temperature differences and decreases as the difference of temperature is made larger.

When the thermocouple is used to cool the air in an enclosure further temperature gradients are set up between the thermoelectric junction and the air. The drop in air temperature which can be achieved then becomes 20 to 30°C. If a greater degree of cooling is required a second stage thermoelectric unit must be added. This will need to have a greater capacity and loading than the first stage.

A single thermocouple designed to work with a moderate direct current, in the range 5 to 10A, has a cooling power limited to about $\frac{1}{2}$W. Cooling units of many watts capacity are made from a large number of thermocouples, usually mounted side by side and connected electrically in series. The thermoelectric cooling principle offers the facility of designing proportionally smaller units down to very low levels of cooling power.

The use of thermoelectric cooling for domestic refrigeration is an obvious possibility which offers the advantages of silent operation and avoidance of the use of chemical refrigerants, but requires a small rectifier/transformer unit to furnish the direct current. The exploration of this potential application is receiving very careful study, in which technical problems and technical performance are being assessed alongside true economics. Note has to be taken of the very high degree of economy in manufacture which the conventional refrigerator industry has achieved.

There are many fields of instrumentation and engineering in which thermoelectric refrigeration appears to offer more opportunities. Instruments which have already been developed and described include an automatic dew point hygrometer, a cooled baffle for an oil-diffusion vacuum pump and a thermostat for electronic equipment. There are many interesting possibilities opened up in the prospect of being able to maintain control units, for example oscillator crystals, at stabilized temperatures below ambient. The power handling capacity of transistors and other components may be increased if local cooling by thermoelectric means can be applied.

Other possible applications of thermoelectric cooling arise in the fields of medicine and biology. Its use in surgery is perhaps somewhat problematical because the space available for the insertion of a cooling unit in the patient during an operation is limited and the amount of heat which must be extracted, for example to cool down arteries, is much larger than would at first sight be expected—the human body is remarkably effective in its power to maintain a flow of heat to a locally chilled area. There is a possible use for cooling measuring equipment and microtomes, where constant low temperature is important, and for cold-storing specimens in the portable equipment in which they are conveyed, for example, from the operating theatre to the pathology laboratory. A thermoelectric cool chest would be an excellent solution to the problem of storing biotics and antibiotics in hospitals and in doctors' surgeries.

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A New Data Processing System

The English Electric Company has placed on the market a new data processing system, known as the KDP.10, which is believed to be the most comprehensive and up-to-date system available anywhere in the world today.

The experience gained from the application of 'Deuce' to specific sectors of the commercial data processing field led the company to undertake intensive studies to find the ideal form which such data processing systems could take.

The new data processing system which is to be built at the company's Kidsgrove works, is based on a design by the Radio Corporation of America, modified to suit manufacturing techniques and operational requirements in the United Kingdom and Europe.

It is a complete and fully transistorized data processing system embodying the latest techniques of advanced computer design and is capable of handling alpha-numeric data and employs magnetic tape, punched paper tape and punched card input-output as well as printed output. These facilities allow the system to be integrated with any existing mechanized office and, by the use of magnetic tape, provide for the daily scrutiny and up-dating of records and files in organizations handling masses of detailed information.

Its magnetic core storage is expandable to provide for a very large capacity and its outstanding characteristics include an extremely high rate of transfer of information to and from the magnetic tape storage. The system can record and transfer information on tape at a speed of 33 333 characters/second in the quantity required—unlike the conventional systems which provide for standard lengths of tape whatever the amount recorded.

The KDP.10 system consists of a number of integrated units covering high speed store, programme control unit, tape selection and buffer unit and a console with associated monitor printers and paper tape punch, a paper tape reader and high speed printers.

The high speed store is a random access magnetic core device which provides storage for programmes and data. This memory can be easily expanded to meet system requirements within the range of 16 000/260 000 characters.

A tape selection and buffer unit *A* allows for the connexion of from one to eight tape stations to the computer and controls reading from and writing to magnetic tape on these stations. Programme instructions indicate the appropriate magnetic tape unit for input or output. It is important to note that the number of magnetic tape units directly controlled by the computer may be increased to 62 by connecting the tape selection unit *B* to each of the lines of unit *A*.

Paper tape punch and tape verifiers are used for the original preparation and verification of the system coded 7-hole punched paper tape operating at a rate of 1 000 characters per second for subsequent input into the computer.

The main output is from a high speed line printer operating under the control of the computer and producing up to six copies at 600 lines per minute; each line may have up to 120 characters.

Included in the peripheral equipment complementary to KDP.10 are a card transcriber, transcribing card punch, tape reader verifier and random access file. It is important to note here that in many cases it is convenient to use punched cards for input and output data; off-line card/magnetic tape and magnetic tape/card convertors operating at 400 and 150 cards/minute respectively are available.

The system is readily adaptable to meet the requirements of both small and large organizations; for a smaller concern, wishing to allow for future growth, the KDP.10 can be built up, unit by unit in proportion to the volume of work demanded of it without altering the basic installation.

Short News Items

Premiums for Technical Writing. The scheme for awarding annual premiums for articles on electronics, organized in the past by the Radio Industry Council is now under the joint sponsorship of the Council and the Electronic Engineering Association. Articles published in 1959 will be considered by the panel of judges early in the New Year. Six premiums of 25 guineas each are offered to writers not earning their living wholly or mainly by writing. The judges, headed by Professor H. E. M. Barlow, Professor of Electrical Engineering, University College, London, are given the greatest possible freedom in deciding which articles merit awards, but they principally take into consideration the value of the article in making known British achievements in radio and electronics, originality of subject, technical interest, presentation and clarity. Writers are reminded that articles for consideration should be in the hands of the Secretary of The Electronic Engineering Association, 11 Green Street, Mayfair, London, W.1, before the end of the year. Eight copies of the journal or relevant pages, proofs or reprints with a signed declaration of the eligibility of the writer or writers should be submitted. This latter ruling does not apply to authors of articles published in the following journals as the members of the panel of judges automatically receive copies of these: British Communications and Electronics; Electronic Engineering; Electronic and Radio Engineer; The Post Office Electrical Engineers Journal; Radio Constructor; Wireless World.

'Lead the Enduring Metal' is the title of a film recently produced by the Lead Development Association. The film, which may be borrowed from the Association, has a running time of 28min and describes the production and uses of lead. It is interesting to note that the cable making industry accounts for about one-third of this country's consumption of this metal.

The British Radio Valve Manufacturers' Association announces that in the near future it will hand over to a new Association, the Electronic Valve and Semi-Conductor Manufacturers' Association (VASCA), its responsibilities for semiconductors and industrial valves and tubes. The B.V.A. will continue its separate interest in domestic valves and television tubes. Founder members of the new Association are: English Electric Valve Co. Ltd; M.O. Valve Co. Ltd/General Electric Co. Ltd; Mullard Ltd; Siemens Edison Swan Ltd/British Thomson-Houston Co. Ltd; Standard Telephones & Cables Ltd. Sections, virtually autonomous, will deal separately with: (i) Transmitting, including microwave, valves. (ii) Semiconductors, including rectifiers up to (rating to be defined by the Section). (iii) Industrial receiving valves. (iv) Gas-filled devices. (v) Industrial cathode-ray tubes, photo-electric and photo-sensitive devices. Among the objects of VASCA are the provision of facilities for discussion and co-operation among members; promotion of standardization; encouragement of and participation in research and experimental engineering work and tests; co-operation with the Government, the Services, B.S.I., I.E.C. and I.E.E.; maintenance of a high standard of quality, design and workmanship; dissemination of information to members; compilation and dissemination of statistics; and the promotion of exports. The Secretary of VASCA is Mr. W. R. West, the Chairman, Mr. G. A. Marriott, and the offices are at Piccadilly House, 16 Jermyn Street, London, S.W.1.

Willesden and Hendon Productivity Committee are organizing a Conference to be held at the Headquarters of the F.B.I. at 21 Tothill Street, Westminster, London, S.W.1, on 27 January, 1960. Both large and small organizations should find the Conference of interest to all levels of management as well as people concerned with purchasing. The chairman will be Mr. M. Seaman who is Director and General Manager of British Oxygen Engineering Co. Admission fee will be two guineas. Requests for details and reservations, together with remittance, should be sent to Mr. D. I. Duncan, A.E.I. Switchgear Division, 20 Neasden Lane, Willesden, London, N.W.10.

The first comprehensive all-British Exhibition and Trade Fair to be held in the U.S.A. will be held at the Coliseum Exhibition Building, New York, in June 1960. It has the full backing of the British Government, The Dollar Exports Council, the Federation of British Industries and the British-American Chamber of Commerce in New York and is being organized by British Overseas Fairs Ltd (B.O.F.). The main aim of the Exhibition is to increase the sales of British goods in the U.S.A. and also to publicize and boost the prestige of Britain and British Industry throughout the North American continent. A New York Liaison Committee, under the Chairmanship of Sir Alexander Brackenridge, K.B.E., M.C., meets regularly in New York and is in continuous contact with the organizers and a similar Liaison Committee in London. Further particulars can be obtained on application to British Overseas Fairs Ltd, 21 Tothill Street, London, S.W.1.

The International Congress on Microwave Valves will be held in Munich from 7 to 11 June 1960. Papers may be read in German, English or French. Contributors are invited to write for further particulars to: Vortragsausschuss der Internationalen Tagung, Mikrowellenröhren, z. Hd. Herrn Prof. Dr. W. Kleen, München 8, Balanstrasse 73.

The Cambridge Instrument Co. Ltd have opened new research laboratories at Chesterton Road, Cambridge; the opening ceremony being performed by the Rt. Hon. the Lord Adrian, O.M., F.R.S., on 14 October. The main laboratory block consists of four floors approximately 100ft long and 40ft wide. Each floor has a separate function and the experimental or working area of each is laid out on an 'open vista' plan. On the ground floor is the Mechanical Engineering Laboratory and a constant temperature room. The first floor is devoted to Physics and Electronics while the second floor houses the Physical Chemistry Laboratory. The top floor is occupied by the Design and Drawing Office. It is anticipated that these new research facilities coupled with increased production facilities will provide a broader base and materially increase the Company's capacity to meet the requirements of science and industry.

E.M.I. Electronics Ltd have received an order from the Ministry of Pensions and National Insurance for an all-transistor EMIDEC 2400 computer, costing over half a million pounds to manufacture and install. The computing centre will be set up at the Ministry's Newcastle establishment to deal with the huge volume of data processing connected with the new Graduated Pensions Scheme. It has been calculated that there are some 25 000 000 insured persons, with over a million additions or changes each year. The computer will 'store' full statistics such as name and address, amount of each contribution and sum paid to date, of each insured person. Each day, the records of every insured man and woman in the United Kingdom will be processed in order to extract the required information for persons reaching retirement age, and other changes in personal particulars. The computer is capable of completing this task in under four hours. The number of forms which will have to be processed yearly to enable the scheme to operate at top efficiency will total over 30 000 000. Another EMIDEC 2400 to be delivered will be for the Royal Army Ordnance Corps. This will maintain full records of the Army's stocks of motor transport spares, the compilation of accounts for the stores, and the calculation of requirements for stock replenishment and the control of stock levels.

LETTERS TO THE EDITOR

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

Cell Effect in Capacitors

DEAR SIR,—During recent experiments with transistor d.c. amplifiers a puzzling drift was found to be due to a relatively pronounced cell effect existing in a 3 000μF 6V electrolytic capacitor. The capacitor, which was used to reduce short term fluctuations in input current to the circuit, showed a reluctance to follow changes in circuit voltage and acted like a cell which required charging or discharging before static conditions could be obtained.

Although most engineers are familiar with the cell effect, it may not be generally known that capacitor manufacturers have determined its cause and evolved methods of reducing it to negligible proportions.

Slight impurities in the aluminium electrodes are responsible for the effect, and making both electrodes from the same piece of material is a great help. If cell effect is already present in a capacitor, it may be reduced by shorting the electrodes for about 3 weeks. A combination of a common electrode material and shorting out will produce the best results.

Yours faithfully,

P. K. TOMKINS,
Nucleonics Laboratory,
Burndept Limited, Erith.

A Simple Anti-Coincidence or Coincidence Circuit

DEAR SIR,—A simple circuit, which can function as an anti-coincidence or a coincidence circuit whenever desired, was constructed using a 6AS6 valve. The 6AS6 is a pentode having two control grids, and is suitable for fast signal operation. The input pulses to the two grids were made square by 12AT7 double-triodes functioning as univibrators. The output pulse from the 6AS6 was fed to a 6AK5 cathode-follower, and subsequently it can be fed to a binary scaling circuit. This arrangement has been used in operating a gas-flow G.M. counter in anti-coincidence with a ring of cosmic ray guard-counters for low level tracer work*.

Fig. 1 shows the circuit diagram. The left-hand sides of the 12AT7's are conducting, negative pulses are fed to the first grids of the two valves, and square pulses, whose width can be adjusted with the potentiometers RV_1 and RV_2, develop at the anodes. Normally the 6AS6, having an anode potential of 180V, is not conducting as a negative bias is applied to both the grids and this can be adjusted from 0 to 25V by the potentiometers RV_3 and RV_4. For the anti-coincidence arrangement the grid g_1 is given a -20V bias and receives the positive square pulses from the line of the central counter, while the grid g_2 is given zero bias and receives the negative square pulses, suitably widened, from the line of the outer guard-counters. The pulses from the guard-counters only and also the simultaneous pulses from the central and one of the guard-counters will not be recorded, whereas, pulses from the central counter only will give sharp negative output pulses. For the coincidence arrangement the grid bias g_2 is changed to -15V and fed with the positive output pulses from the univibrator by changing the position of the switch S. The final output pulses can be registered by a suitable scaler and recorder.

The author wishes to express his thanks to Prof. B. D. Nag for suggesting the problem.

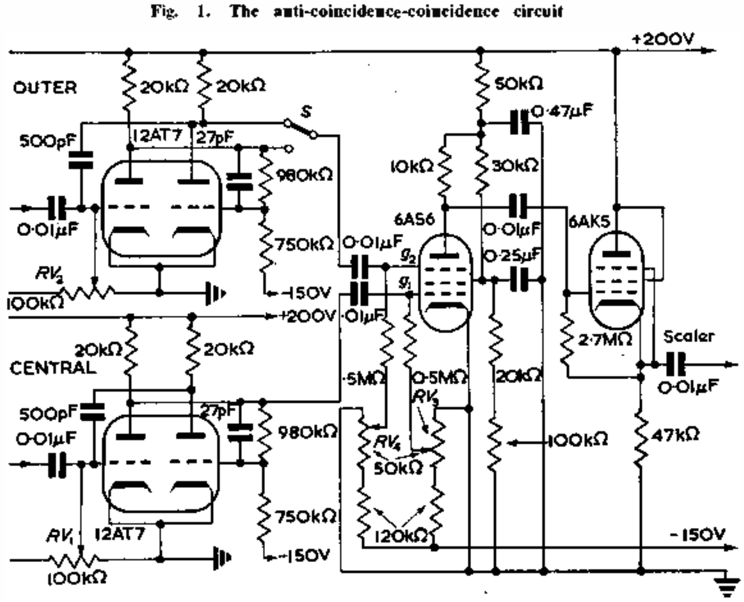
Yours faithfully,

F. HOSAIN,
Saha Institute of Nuclear Physics,
Calcutta 9.

REFERENCE

* HOSAIN, F. *Naturwissenschaften* 45, 107 (1958).

Fig. 1. The anti-coincidence-coincidence circuit



Current Creep in Transistors

DEAR SIR,—Mr. R. E. George, who reports (ELECTRONIC ENGINEERING, March 1959) on an approximately exponential drift in transistor I_{co}, may be interested to compare my experience of an unaccountable drift in oscillator frequency using certain specimens of transistor. It was reported in *Wireless World*, September 1957, p. 443-4 (see also October, p. 487).

This frequency drift, presumably due to individual transistor characteristics, for some other samples showed no signs of it, also had an approximately exponential characteristic with a time-constant of the order of hours. Similarly, the possibility of a thermal origin had been ruled out because of the low power and long time-constant. Unlike the effect reported by Mr. George, however, the drift continued for a short while after switching off, and the transistors concerned took at least several hours to return to their original state. The makers were unable to shed any light on the affair.

If Mr. George would like further details I would be happy to oblige.

Yours faithfully,

M. G. SCROGGIE,
Bromley, Kent.

Current "Creep" in Transistors

DEAR SIR,—The slow growth of reverse current through pn junctions reported by Mr. R. E. George (ELECTRONIC ENGINEERING, March 1959) is similar to the effects reported previously by Plummer[1]. In these experiments slow growths of current were produced by the adsorption of moisture on the surface of a semiconductor diode. In a series of measurements on pn junction diodes under controlled humidity, we[2] established that the slow growth of current and the 1/f noise were closely connected with the adsorbed water on the surface of the diodes. With very low humidity, both of these effects disappeared. Therefore it seemed probable that an explanation of the slow growth of current would also include an explanation of the noise spectrum.

We proposed a model in which the water film was insulated from the semiconductor by an oxide layer and separation of ions in the water film would occur when a voltage was applied across the junction. The separated ions would diffuse across the surface of the oxide layer generating electrostatically a surface 'channel' of the type discussed by Brown[3], and the slow extension of this channel would account for the growth of current. The diffusion constants of ions ($\sim 10^{-5}$ cm^2 sec^{-1}) are of the correct order to account for the long growth times. In this model, current would not flow through the water film and electrolysis would not occur[4].

An analysis of the current fluctuations, produced by fluctuations of the density of diffusing surface ions, which modulated the resistance of a surface channel, was made using the method described by Richardson[5], and the spectrum obtained varied as $1/f^{1.5}$. The measured exponent was $-1{\cdot}25$.

Measurements made by Law[6] of the adsorption characteristics of germanium oxide surfaces indicate that the problem of completely excluding water from a transistor is formidable. Present manufacturing technique makes use of the adsorption of a few monolayers of water to reduce the surface recombination velocity in pnp transistors and thereby to increase the current gain. In applications where low frequency noise and current 'creep' become troublesome, it may be worth sacrificing some current gain by rigorously excluding adsorbed water, and it would be interesting to hear the opinion of transistor manufacturers on this point.

Yours faithfully,

J. E. PALLETT,
University of Edinburgh,
Department of Engineering.

REFERENCES

1. PLUMMER, A. R. F. Observations on the Growth of Excess Current in Germanium pn Junctions. *Proc. Phys. Soc., Lond., 69B*, 539 (1956).
2. PALLETT, J. E., METTZER, B. Proc. International Conference on Solid State Physics in Electronics and Telecommunications, Brussels, 1958. (To be published by Academic Press).
3. BROWN, W. L. n-Type Surface Conductivity on p-type Germanium. *Phys. Rev. 91*, 518 (1953).
4. ERIKSEN, W. T., STATZ, H., DEMARS, G. A. Excess Surface Currents on Germanium and Silicon Diodes. *J. appl. Phys. 28*, 133 (1957).
5. RICHARDSON, J. M. The Linear Theory of Fluctuations Arising from Diffusional Mechanisms —An Attempt at a Theory of Contact Noise. *Bell Syst. Tech. J. 29*, 117 (1950).
6. LAW, J. T. The Absorption of Gases on a Germanium Surface. *J. Phys. Chem. USSR. 59*, 543 (1955).

The Correspondent Replies:

DEAR SIR,—Following my letter[1], my attention has been drawn[2] to a paper by Plummer[3] who reported the slow increase of reverse current across a germanium pn junction as attributable to water contamination. Although the rate of increase of current in the presence of the electric field across the junction was very slow, he found that on removal of the field, the effect had almost disappeared in one second. This creep phenomenon is confirmed by Mr. Pallett in his recent letter, who also finds it associated with excessive flicker noise. During measurements on transistor noise, I have noticed that some of the transistor types producing excessive noise tended to show current creep, but no systematic measurements were undertaken to establish this.

While carrying out simple routine checks on a batch of r.f. transistors, I had already discovered one with creep in I_{co} checked at 4·5V as described previously. I tried this sample in Mr. Scroggie's circuit[4]. Over a period of 28min, the frequency decreased by 0·1 per cent, while the current drawn increased by 30 per cent. The dissipation did not exceed 9mW. In order to test whether there was any relation between frequency drift and creep, the current drawn was reduced to its original value by increasing the base feed resistance. Other transistors of the same type under the same conditions showed far less drift than the specimen showing creep. It would appear therefore that the two effects might be related.

There can be no doubt that transistor manufacturers are aware of this problem. One manufacturer is already marketing low noise transistors, which I have found to be comparatively free from creep.

Yours faithfully,

R. E. GEORGE,
Electronics Section,
Physics Department,
Guy's Hospital Medical School,
London.

REFERENCES

1. GEORGE, R. E. Current 'Creep' in Transistors. (Letter to the Editor.) *Electronic Engng. 31*, 172 (1959).
2. ROBERTS, F. F. Private Communication.
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4. SCROGGIE, M. G. Transistor oscillator Stability. *Wireless World 63*, 444 (Fig. 2(b)) (1957).

Transistor Circuit Temperature Stability Analyses

DEAR SIR,—In a recent letter by Ryan[1] he indicated that the frequency selective amplifier he described achieved bias stabilization by means of negative feedback. The amplifier shown in Fig. 1 is

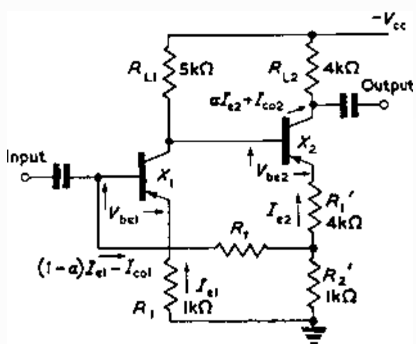


Fig. 1. The circuit described by Ryan

the configuration described by Ryan, where R_f was a twin-T filter. The actual stability of this or any other transistor circuit can be predicted by an analytical procedure as follows.

Let any bias parameter be a function of the main temperature dependent terms in the circuit. For this example assume the base-emitter voltage (V_{be}) and the reverse saturation collector-base current (I_{co}) cause the main temperature instability. Assume the temperature variations of I_{e2} is of interest, then.

$$I_{e2} = f(V_{be1}, V_{be2}, I_{co1}, I_{co2} \ldots\ldots \quad (1)$$

By differentiating equation (1) with respect to temperature:

$$dI_{e2}/dT = \partial I_{e2}/\partial V_{be1} \cdot dV_{be1}/dT + \partial I_{e2}/\partial V_{be2} \cdot dV_{be2}/dT + \partial I_{e2}/\partial I_{co1} \cdot dI_{co1}/dT + \partial I_{e2}/\partial_{co2} \cdot dI_{co2}/dI \ldots \quad (2)$$

Assuming

$$dV_{be1}/dT = dV_{be2}/dT \text{ and}$$
$$dI_{co1}/dT = dI_{co2}/dT$$

then equation (2) can be written;

$$dI_{e2}/dT$$
$$\left[\frac{\partial I_{e2}}{\partial V_{be1}} + \frac{\partial I_{e2}}{\partial V_{be2}}\right] dV_{be}/dT + \left[\frac{\partial I_{e2}}{\partial I_{co1}} + \frac{\partial I_{e2}}{\partial I_{co2}}\right] dI_{co}/dT \quad (3)$$

The partial differential terms enclosed in brackets depend on the circuit configuration and can be found by taking suitable mesh equations and differentiating with respect to each of the temperature dependent terms.

Defining

$$D_v = \partial I_{e2}/\partial V_{be1} + \partial I_{e2}/\partial V_{be2} \ldots \quad (4)$$

and

$$D_i = \partial I_{e2}/\partial I_{co1} + \partial I_{e2}/\partial I_{co2} \ldots\ldots \quad (5)$$

it follows simply (although somewhat tediously) that:

$$D_v \quad R_{l1} - R_1/D_c \ldots\ldots\ldots \quad (6)$$

$$D_i = R_{l1}(R_f + R_1) + R_1R_1'/D_c \ldots \quad (7)$$

where

$$D_c = R_1(R_1' + R_2') + R_1' R_{l1}$$

Using circuit values as indicated in Fig. 1 and assuming (from experimental observations) that V_{be} varies linearly at $-0{\cdot}002$V/°C and I_{co} at any temperature is given approximately by 7·03 exp $(-8400/T)$ for alloy junction low power pnp transistors), it follows that:

$$D_v \quad -0{\cdot}004$$
$$D_i - 0{\cdot}6$$

and by integrating equation (3)

$$I_{e2}(T_2) \quad I_{e2}(T_1) + 0{\cdot}004 \times 0{\cdot}002[T_1 - T_2] + 0{\cdot}6[I_{co}(T_1) - I_{co}(T_o)] \ldots\ldots\ldots \quad (8)$$

Assuming $T_1 = 65$°C and $T_o = 20$°C, the predicted shift in I_{e2} is 0·1mA. This is in excellent agreement with experimental observations.

Yours faithfully,

B. A. BOWEN,
University of Syracuse,
Syracuse, N.Y.

REFERENCE

1. RYAN, W. D. Tuned Transistor Audio Amplifier. *Electronic Engng. 31*, 103 (1959).

Correction

In 'The Correspondent Replies' by R. H. Evans on page 558 of the September issue, the last line of the final equation should read:

$$-\frac{V^d(0)\cos[\omega t + \phi_m = \phi_n - (d+1)\theta]}{(a^2+\omega^2)^{(d+1)/2}} e^{at} \Bigg\}$$

BOOK REVIEWS

Digital Computer Components and Circuits

By R. K. Richards. 511 pp. 70 figs. Medium 8vo. D. Van Nostrand Co. Inc., New York, London. 1958. Price 64s.

DR. R. K. RICHARDS is already well known to engineers and others associated with digital computers for his earlier book, 'Arithmetic Operations in Digital Computers', which contains an excellent account of the logical design of the arithmetic units of various types of computers. Dr. Richard's new book deals with the circuits and components that are used in computer logic and storage. It has been written as a companion volume to the earlier book but it can be read as a self-contained account of the subject. Although the reader is assumed to have some knowledge of electronics, the book should prove of interest not only to engineers both in computing and allied fields, but also to the many users of computers, who, despite a very limited experience of electronics, would like to appreciate more of the engineering side of the machines that they use.

The techniques being used in computers are changing so rapidly that it is a formidable task to describe them all in one volume, and indeed it is impossible to ensure that all are included since new ones will be appearing even while the manuscript is with the publishers. However Dr. Richards has made a bold attempt to be as comprehensive as possible within these limits and, although some of the information is of only historical interest, most of the techniques that will be present in working computers for many years are described. Despite all the current news of new techniques it should be remembered that by far the majority of commercially available computers which are installed and working reliably use thermionic valves and that the next round of machines will be using transistor circuits. Both of these techniques are dealt with at length in this book.

For electronic engineers in other fields an introductory chapter outlines the general concept of a digital computer. This introduction also contains a short dissertation on reliability and marginal checking, sides of computer circuit design that cannot be too heavily emphasized. A chapter is devoted to diode switching circuits and a long chapter each to use of thermionic valves, transistors and magnetic cores in circuit logic. A brief account of the physics of transistors and cores is given. Many different examples of these various techniques are given, not all of which have found (or will find) their way into successful computers. However, where particular circuits have been used in practice the computers are mentioned and the speeds of operation achieved are quoted. Not unnaturally almost all these references are to American computers; from a knowledge of these machines the reader can judge the success of the appropriate circuit.

A wide range of storage devices is described, from the now almost extinct cathode-ray tube and mercury delay-line stores to the magnetic core store which is the store most used in currently planned computers. A chapter is devoted to storage on magnetic surfaces (drums and tape) and accounts of the various methods of recording are described.

Further chapters cover decimal counting circuits and analogue-to-digital and digital-to-analogue convertors. Under miscellaneous components the cryotron is described. As the author says it may seem ridiculous to consider operating a computer at temperatures very near to the absolute zero but advantages of simplicity, small size, high-speed, low power consumption and low cost, in fact all the attributes that a computer engineer looks for in his components, may well make it the computing component of the future.

The reader should find something about every type of computer component in this book, and even if the reference may be too short the excellent bibliography will indicate where further information may be obtained. In all, this is a useful book and one suspects that Dr. Richards will be called upon to produce further editions in order that it may keep abreast of developments in the field.

C. H. DEVONALD.

Electromagnetic Radiation from Cylindrical Structures

By J. R. Wait. 199 pp. 69 figs. Demy 8vo. Pergamon Press Ltd. 1959. Price 50s.

THE author of this book is one of the leading authorities on the theory of radio aerials and has written many papers and reports on the subject. This book is partly a collection of his publications, some of which are incorporated with only minor modifications. It is useful to have this material in one volume, especially as it also includes a review of papers from the U.S.S.R.

The book will be intelligible only to the most accomplished mathematicians. After eight pages of introductory material the author plunges into the mathematics, making free use of Bessel and Hankel functions. After that there is hardly a page without these or some other transcendental function. The statement on the dust-cover that the reader . . . "requires knowledge of complex functions and electromagnetism at the intermediate level" is misleading, for only readers with good final honours mathematics will understand this book.

A review of the basic concepts of electromagnetism is given in an appendix. This is too brief to be recommended to students new to the subject, but the reader experienced in electromagnetic theory will find it a useful summary and will note some novel features of the presentation—he may not agree with them all.

The main aim of the book is to show how to compute the radiation patterns of slot radiators and dipole or stub aerials in or near wedges, cylinders and similar structures, and there are many diagrams of radiation patterns. There are also chapters on scattering from a dielectric cylinder and on diffraction by a cylindrical surface.

There is a long bibliography, mostly of mathematical papers, but no index—an omission which greatly reduces the value of any book. The bibliography does not include any of the series of pioneering papers in vol. 93, part IIIA of the Journal of the Institution of Electrical Engineers. Some of these are still of value to the practical engineer, but were perhaps thought to be of insufficient mathematical interest.

K. G. BUDDEN.

Electronic Engineers Reference Book

Edited by L. E. C. Hughes. 1 588 pp. 9 584 figs. Crown 8vo. 2nd Edition. Heywood & Co. Ltd. 1959. Price 84s.

A LARGE part of the text of this new edition has been revised and a considerable amount of new material added, notably new sections on non-destructive testing, components, special components, radiation detection, digital computer applications, simulators and electronic telephone exchanges.

This reference book endeavours to put before engineers in industry and in development laboratories, some of the latest knowledge and techniques which might not be easily available to them. It provides suggestions and possibilities to be taken into consideration when problems are examined from various points of view, physical, chemical, production, safety, reliability and maintenance.

The Fourier Integral and Certain of its Applications

By N. Wiener. 199 pp. Demy 8vo. 2nd Edition. Dover Publications Inc., New York, Constable & Co., London. 1959. Price 12s.

THIS new paper covered edition is in substance an elaboration of a course of fifteen lectures on the Fourier Integral and its Application, given at the University of Cambridge during 1932. The author's development of certain theories concerning the Fourier Integral—notably those of Plancherel and Tauber—have made it possible to apply analytical techniques to areas in physics and engineering which are basically statistical in nature.

Logical Design of Digital Computers

By M. Phister Jr. 408 pp. 60 figs. Medium 8vo. John Wiley & Sons Inc., New York. Chapman & Hall Ltd, London. 1958. Price 84s.

DR. PHISTER has produced a book on the logical design of digital computers which tackles the problem in a far more algebraic way than most writers have done previously. One expects that such a book will contain many diagrams of adders, multipliers, control circuits and so on, where each diagram comprises 'black boxes' representing the logical elements, the shape of the unit defining the particular function that the element carries out. Turning the pages of this book, however, one finds that there is a marked absence of diagrams and in their place are equations and it is these Boolean equations that define the computer.

The design of a computer can be divided conveniently into three parts which are to a large extent independent. First, the computer has to be specified in general terms (size of internal storage required, speed of operation, types of input and output equipment) so that it can do the job for which it is intended. Secondly, the electronic circuit elements from which this system is to be built must be designed. Finally, the arrangement of the inter-connexions between these units must be worked out so that the resulting machine meets the system specification. It is this last stage that is the logical design of the computer in the sense of Dr. Phister's book, and the ultimate product of this design is the wiring schedule from which the computer is constructed. As Dr. Phister says in his introduction the system specification and electronic design work call for flair and inspiration, but much of the detailed work of logical design can be reduced to rules and hence this step can more and more become an activity that can be automatically handled by a suitably programmed computer. The algebraic approach of this book should be particularly suited to this form of mechanization.

The early chapters outline the basic parts of a computer and the electronic units that go to make up these parts. These circuit elements are defined in terms of Boolean functions and the remainder of the book is concerned with the building of the different parts of a computer considered as a problem in Boolean algebra. Only serial, synchronous computers are dealt with, but if the design problems of such computers are understood other types of computers can be readily mastered. In order that the Boolean operations involved can be appreciated by the reader (who may well be unfamiliar with this subject) an extensive account of Boolean algebra is given, including several methods of simplifying Boolean functions. Since each term in these Boolean functions corresponds to a logical element in the resulting computer, these simplifications tend to reduce the equipment involved. Brief descriptions are given of various storage devices, such as magnetic drums and tape, and input and output units such as paper tape and card readers and punches, and the ways in which these fit into the computer design are outlined.

To one more familiar with the diagrammatic approach this treatment is at first difficult to understand. Dr. Phister claims that in his experience the algebraic approach has many advantages to offer as computers become more and more complicated. One feels, however, that while this method may appeal to mathematicians (and many logical designers are mathematicians by training) there is still a place for the logical diagram to assist the maintenance engineer whose background is usually of a more practical nature.

Dr. Phister indicates that his book is intended to form the basis for a university course and it is well supplied with exercises. However, no solutions are given and one feels that a student's supervisor would be kept busy in producing the answers—one exercise, for instance, begins, 'Design a serial decimal general-purpose computer having the following order code: ....'

With so many books appearing on various aspects of computers it is refreshing to find a novel approach and Dr. Phister is to be congratulated on this account. The presentation is good and in such a difficult text there is a surprising absence of misprints. As is usual these days an extensive bibliography concludes each chapter.

C. H. DEVONALD.

Electron Physics and Technology

By J. Thomson and E. B. Callick. 527 pp. 70 figs. Demy 8vo. English Universities Press Ltd. 1959. Price 50s.

AS the title implies, this book attempts to give a comprehensive account of the wide range of electron physics from the basic theory up to the technology of modern devices. This is a formidable task in which the authors have not quite succeeded. The main defects lie in the first two sections of the book covering the basic theory and entitled 'Free and bound electrons,' and 'Electronic devices employing space-charge variation.' These sections are too much condensed to form an adequate introduction to electron physics *per se* although they may form a sufficient basis for an understanding of the rest of the book. In addition there are a number of minor errors, such as symbols in diagrams differing from those in the text, and also some which are rather more serious. For example in the chapter on electronic conduction in a vacuum the equations for the motion of an electron in combined electric and magnetic fields (3·1) are valid if, for the charge on an electron, the substitution $e = -1{\cdot}6 \times 10^{-19}\ C$ is made, whereas for the following equations to be correct the magnitude of e $(+1{\cdot}6 \times 10^{-19}\ C)$ must be used. There is a similar confusion in the analysis leading to the equivalent circuit of the triode.

The following sections entitled 'Electron Inertia,' 'Microwave Devices' 'Special Purpose Tubes,' and 'Materials and Construction,' are much better and indeed the book is written essentially from the point of view of the technologist. A considerable amount of material on the technology of electron devices is presented in a very readable form although occasionally technical terms are used without explanation.

Electronics degree courses are so overcrowded that scarcely any mention of technology is possible. This book helps to bridge the gap between theory and practice and may be recommended to students of electronics as an introduction to the technological aspects. In its present form it is not very suitable as an introduction to electron physics, and in order to achieve this aim the initial sections should be expanded appreciably and rather more carefully written.

W. A. GAMBLING.

Jets and Rockets

By A. Barker, T.R.F. Nonweiler and R. Smelt. 245 pp. 121 figs. Demy 8vo. Chapman & Hall Ltd. 1959. Price 35s.

THE development of the rocket and jet unit is closely linked with the advance of high-speed aerodynamics and this fact might almost be described as the theme of this book.

The eighteen chapters include discussions on the momentum theory of jet and rocket propulsion, the rocket motor, performance calculations, jet propulsion, air resistance and aerodynamic design at high speeds, engine performance and the performance of rocket projectiles, ending with a chapter on future developments, appendices and a bibliography.

ELECTRONIC EQUIPMENT

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

(Voir page 567 pour la traduction en français: Deutsche Übersetzung Seite 572)

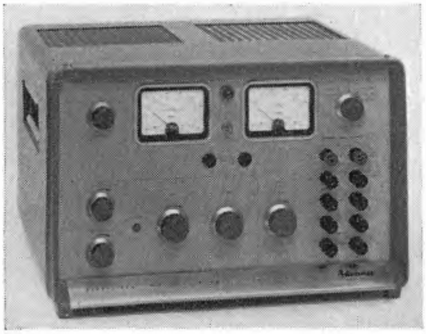
STABILIZED POWER SUPPLY UNIT

(Illustrated below)

Advance Components Ltd, Roebuck Road, Hainault, Ilford, Essex

The Advance stabilized power supply type L.101 is a special purpose instrument of considerable flexibility which provides a constant voltage source of 600V positive and two stabilized negative 150V d.c. supplies, one of which is a variable line having a high impedance source. Two stabilized 6·3V 4A a.c. supplies are also provided.

A stabilized variable 0 to 150V negative line of high source impedance is provided as a grid bias supply or for use in circuits drawing a small current



load of not more than 5mA. A 100 to 800V unstabilized line is also provided for general use.

The two positive lines have a maximum total load rating of 300mA. The positive 600V and negative 150V d.c. outputs may be used in series to provide a total 750V stabilized d.c. output with a maximum rating of 30mA.

Voltage variation on the positive 600V and negative 150V lines is better than 0·03 per cent of output for a 10 per cent change of input voltage, and the variation in output from the two 6·3V a.c. lines is less than 0·02V for the same condition.

The output impedance is less than 0·1Ω at d.c. and less than 0·5Ω at 50kc/s on all d.c. lines. The variable negative 0 to 150V d.c. line has a source impedance of 12kΩ. Ripple content is less than 3mV r.m.s. on all stabilized outputs at 150V and below. Above 150V the ripple content is less than 0·002 per cent r.m.s. and on the unstabilized 100 to 800V d.c. it is about 3V r.m.s. at 800V with 300mA load.

Adequate metering facilities are provided by two first grade meters, the voltmeter being calibrated from 0 to 250V and 0 to 1kV, and the ammeter from 0 to 50mA and 0 to 500mA.

Other facilities include control of the h.t. stabilized d.c. line enabling the output to be set to any desired level from 0 to 600V together with a source resistance control over the range 0 to 40Ω. Hum can be superimposed on the stabilized d.c. h.t. line and may be varied between 0 and 6V r.m.s. to give a good quality sine wave at the mains input frequency.

All outputs are floating with respect to the chassis, the positive and negative lines being connected through the neutral terminal.

EE 15 751 *for further details*

PULSE OSCILLOSCOPE

(Illustrated below)

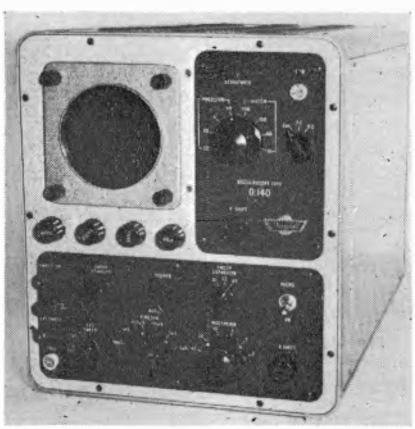
Furzehill Laboratories Ltd, 57 Clarendon Road, Watford, Hertfordshire.

The Furzehill 0·140 oscilloscope, now in full production provides facilities for pulse and r.f. examination at a competitive price. It embodies a time-base which can be used in either triggered or free-running condition over a range of 0·1μsec/cm to 0·5sec/cm. A useful feature is the fine control of the triggering which enables the horizontal sweep to be started at any desired point on the waveform under examination. On recurrent setting positive and immediate synchronism is provided by a separate sync amplifier which may be fed either from the internal amplifier or from an external source.

The deflexion amplifier is direct-coupled and is provided with 8 calibrated sensitivities, namely 50, 15, 5, 1·5, 0·5 and 0·15V/cm level to 10Mc/s plus two further sensitivities of 50mV/cm to 3·3Mc/s and 15mV/cm to 1Mc/s. Because the instrument is intended for pulse examination, the amplifiers are critically damped to obviate ringing.

The calibration accuracy on both time and voltage is better than 5 per cent but an internal calibrating wave of 1 per cent accuracy enables a check to be made at any time.

A four-inch p.d.a. flat-ended tube is used with astigmatism control, while the



graticule is provided with variable illumination which facilitates rapid and accurate measurement.

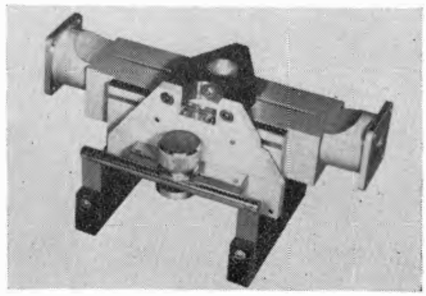
EE 15 752 *for further details*

WAVEGUIDE SLOTTED SECTIONS

(Illustrated below)

Distributed by: Leland Instruments Ltd, Abbey House, Victoria Street, London, S.W.1.

These instruments, manufactured by the Polytechnic Research Development Co. of New York, provide accurate standing wave and impedance measurements in the microwave region from 2·6 to 40kMc/s. They may be used both in design and routine tests for obtaining v.s.w.r. and



phase data on waveguide components or employed in conjunction with tuning devices to eliminate transmission line mismatches. In addition they may be used as permanent parts of radar and communication systems for checking antenna reflection characteristics. The frequency range is covered in eight standard waveguide sizes.

The slotted section consists of a length of waveguide, precisely machined from a solid block of metal, with a narrow longitudinal slot in the centre of the top wall. A carriage is provided to move the pick-up probe along the length of the slot. This carriage is supported on three ball bearings which roll in precision-ground and hardened grooved runways attached to the body of the unit. Adjustment of the motion of the carriage with respect to the body is provided so that the probe tip can be made to move accurately parallel to the inside surface of the waveguide.

Probe carriages are designed to provide smooth and accurate carriage motion, without backlash or play. The absence of sliding contacts reduces wear to a minimum and instrument accuracy is maintained indefinitely.

For facilitating precise phase measurements, each slotted section is equipped with a vernier centimetre scale which indicates the distance between the probe and the outer end. The probe travel is approximately one guide wavelength at the lowest frequency in each range, except in the two lower frequency lines where it is three-quarters wavelength. Waveguide inside dimensions are maintained accurately to provide a standard of impedance. Slot ends are tapered to reduce the residual v.s.w.r. below 1·01, except in the two lower frequency lines where the discontinuity is negligible.

EE 15 753 *for further details*

CONDUCTIVITY BRIDGE

(Illustrated below)

A. M. Lock & Co. Ltd, Prudential Buildings, 79 Union Street, Oldham, Lancashire

The Lock portable conductivity bridge, type BC1, uses the latest transistor circuits and manufacturing techniques to ensure long trouble-free service under laboratory or field conditions. It is housed in a wooden carrying case measuring approximately 12in × 7½in × 5½in overall, and is supplied complete with a dip type conductivity cell.

Ease of operation and reading is ensured by carefully positioned controls, clearly marked on the sloping front panel. A quick release wooden lid gives protection for transit purposes; in use it is completely removed.

The a.c. measuring bridge is energized at a low potential by a 1·5kc/s transistor oscillator. This high frequency is used to overcome inaccuracies which would otherwise occur at high solution conductivities.

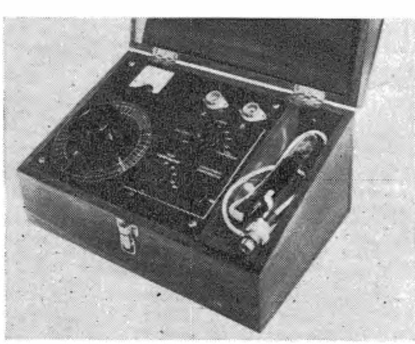
In use, bridge balance is shown by maximum deflexion of the indicator meter. The conductivity of the solution is then read from the large calibrated dial. By using this null-point method, variations in battery voltage or transistor characteristics do not affect the accuracy of measurement.

The instrument covers a measuring range, 0·5 micromhos to 1×10^6 micromhos in six ranges.

The bridge is suitable for use with most types of conductivity cell. If the cell has a constant other than unity, the actual solution conductivity is obtained by multiplying the bridge reading by the cell constant.

Manual temperature compensation over the range 5 to 35°C is provided on Ranges 2 and 3 at the rate of 2 per cent/°C. This correction factor covers certain condensates and unprocessed waters, and most aqueous solutions with conductivity values in the range 5 to 1 000 micromhos.

On other ranges the temperature compensator is not connected because the correction necessary varies according to the composition and concentration of the electrolyte.



A built-in variable capacitor is provided to balance out cell and cable capacitance which would otherwise give a flat balance point.

EE 15 754 *for further details*

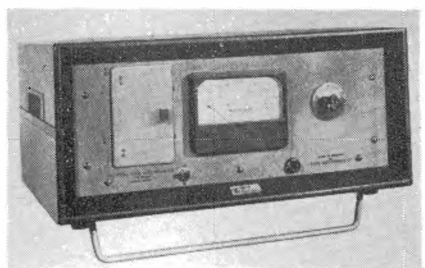
VIBRATING REED ELECTROMETER

(Illustrated below)

Ekco Electronics Ltd, Southend-on-Sea, Essex

The vibrating reed electrometer type N616 has been designed to measure very small direct currents and voltages with high zero stability. The d.c. input is converted to a.c. by means of a vibrating reed capacitor and amplified by a stable a.c. amplifier.

Seven ranges are provided—3, 10, 30, 100 and 300mV and 1 and 3V. The output is rectified and measured on a 5in panel meter calibrated to indicate the d.c. voltage at the input.



For voltage measurements, accuracy is about ±5 per cent while current measurements may be made to the same accuracy provided that the total value of the input resistance is known and the calibration control is adjusted as necessary.

EE 15 755 *for further details*

NON-LINEAR FUNCTION GENERATOR

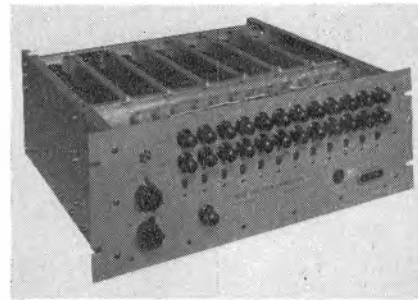
(Illustrated above right)

The Solartron Electronic Group Ltd, 45 Thames Street, Kingston, Surrey.

Recently announced by Solartron is the diode function generator type TR 829. In standard 19in rack-unit form, the TR 829 will generate precise functions of a variable to a high degree of accuracy and stability. It is complementary to the existing range of Solartron non-linear analogue computing units.

The function required is generated in 12 segments, simply set-up by adjusting break points for each segment against a digital voltmeter monitor. These adjustments are made on high-resolution helical potentiometers, over the voltage range zero to plus or minus 100V. Two TR 829 units may be used together if desired, for more complex functions.

The diode function generator incorporates a total of 14 high-gain d.c. amplifiers and does not require additional operational amplifiers to provide full output. Drift and gain errors for each segment are held inside 100mV and 0·2 per cent respectively, and, since these errors are non-cumulative over the whole function generated, the long-term stability overall is always better than 0·5 per cent average.



EE 15 756 *for further details*

DIFFERENTIAL TRANSFORMER TRANSDUCER

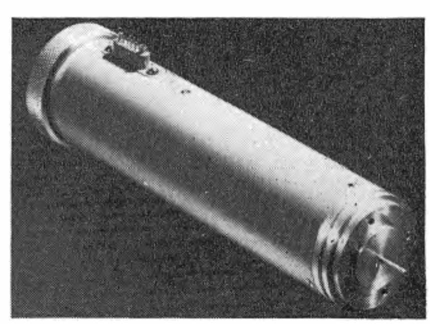
(Illustrated below)

James Scott & Co. (Electrical Engineers) Ltd, 68 Brockville Street, Carntyne Industrial Estate, Glasgow, E.2.

In the basic LINTRAN instrument the input is the movement of a shaft and the electrical output is accurately proportional to this movement. However, it is not limited to this function only, since by the addition of calibrated bellows it becomes a pressure measuring gauge and so on. In fact, any quantity in which a change can be translated into movement can be accurately measured by this device.

The LINTRAN uses the well-known principle of a 'differential transformer'. This has two windings accurately spaced relative to a central winding so that the signals from the windings are equal. If an iron armature is now placed inside the windings, movement of this armature results in changes of the internal magnetic fields and the outputs are no longer equal. By careful design the difference in the output can be made accurately proportional to the movement of the armature from its central position. The difference in output is normally shown on a meter which can be scaled directly in terms of movement of the armature. In this design the size of the armature which is included can be varied, and depending on this size full scale movement on the meter can be obtained for as small a movement of the input shaft as one thousandth of an inch or as large as one inch.

Previous transducers using the differential transformer arrangement have tended to be large and cumbersome when designed for use on the normal mains supply. Alternatively, they have required the addition of expensive electronic cir-



cuits to obtain the accurate output required. The LINTRAN, however, has been designed for operation on the normal supplies, but the advantages of a small size, plus accurate output have been retained.

The basic unit is 7¼in long by 1¾in diameter and weighs 2lb (907grms).

EE 15 757 *for further details*

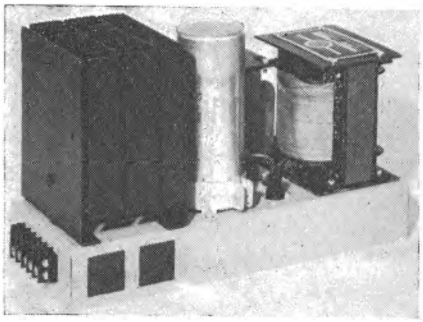
LOW VOLTAGE POWER SUPPLY

(Illustrated below)

Allied Electronics Ltd, 28 Upper Richmond Road, Putney, S.W.15.

The LS353-H supply unit is intended for use in situations requiring a compact, stabilized d.c. source such as heater supplies in d.c. amplifiers, exciter lamps in photo-electric measuring instruments and other applications where batteries are normally used.

The basic unit is a germanium junction rectifier circuit with a transistor voltage stabilizer and a reference potential device.



The system described above is a development of the 'Altron' constant potential chargers supplied for use with 'DEAC' mercury-cadmium, hermetically sealed accumulators. 'DEAC' chargers are available for a range of voltages from 1·5V to 24V and currents up to 100A or more.

The circuit provides a high degree of stabilization against changes in input voltage, and a low ripple content: a pre-set control permits variation of the output voltage over a small range.

Supply units are available either in chassis form or in cases.

EE 15 758 *for further details*

CAPACITANCE SENSING TACHOMETER

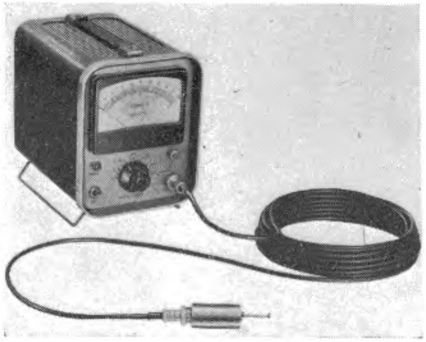
(Illustrated above right)

Grunther Instruments Ltd, 14 Oriental Street, London, E.14.

This capacitance sensing tachometer was developed for a project which required the measurement of gas turbine speeds of up to 120 000rev/min with local temperatures up to 700°C.

This equipment makes use of a probe mounted near to, but not in contact with, any irregularity on a rotating part moving at the speed to be measured or at some multiple or sub-multiple of it. The irregularity may be a keyway, spline, cam, gear tooth, large hexagon nut, or a hole or pip not smaller than 1/16in in diameter and depth.

Each time the irregularity (prominance or depression) passes the probe, a capaci-



tance change is produced. This develops a voltage pulse which is fed via a small diameter flexible cable to the counting instrument, which may be up to several hundred yards distant from the probe.

The tachometer itself is an electronic counting instrument using established techniques. The indicator is a moving-coil meter calibrated directly in revolutions per minute. There are six reading ranges which, with the use of correction factors in certain cases, permit measurements from below 50rev/min to 200 000rev/min with an accuracy better than 2 per cent of full-scale deflexion throughout.

The probe, although sensitive to capacitance changes, is unaffected by extremes of temperatures, vibration or the presence of oils etc.

EE 15 759 *for further details*

STABILIZED DUAL LOW VOLTAGE SUPPLY

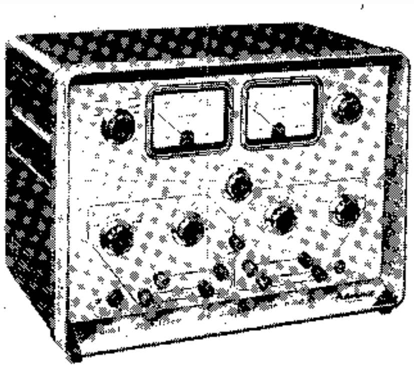
(Illustrated below)

Advance Components Ltd, Roebuck Road, Hainault, Ilford, Essex.

The P.P.3 comprises two transistorized 0 to 30V 1A stabilized d.c. power supplies in one case operating off a common mains transformer. The two sections are floating and can be operated separately, connected in series to give a maximum output of 60V at 1A or to provide positive and negative voltage supplies simultaneously (max. 30V each).

Each section covers the range from 0 to 30V in three overlapping 10V ranges with the variable control directly calibrated in volts. In addition to this a voltmeter and ammeter are provided which can be independently switched to either section, the ammeter having two ranges.

Four terminal output networks are provided to enable the resistance of the connecting leads to the load to be virtually eliminated.



Overload protection is effected by an electronic circuit which gives complete protection against both progressive and sudden (short-circuit) overloads. This circuit being reset from the front panel by operation of the on-off switch. Protection against internal failure or excessive mains surges is handled by a panel mounted fuse in the mains circuit.

EE 15 760 *for further details*

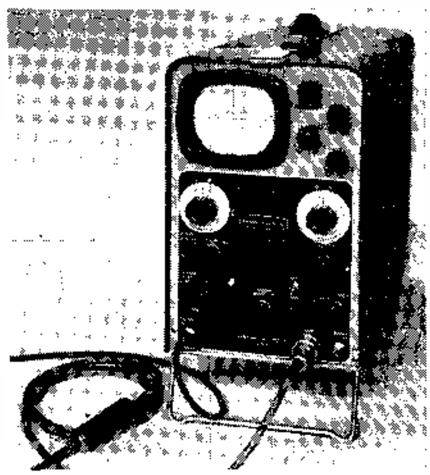
PULSE MONITORING OSCILLOSCOPE

(Illustrated below)

Mullard Equipment Ltd, Mullard House, Torrington Place, London W.C.1

Mullard Equipment Limited have introduced a new and inexpensive type of pulse monitoring oscilloscope with the fast rise-time of 2mμsec, equivalent to a bandwidth of d.c. to 220Mc/s.

The instrument, type L.362, has been



designed for the display or measurement of pulses of between 3mμsec and 3μsec duration, and will therefore have particular value in nucleonic and computer research work, and for the examination of radar pulses, especially in short-range, high-resolution equipment.

Apart from its very wide bandwidth, the L.362 has several other notable features. The sensitivity is 150mV/cm, which, combined with the d.c. coupled amplifiers, makes the instrument particularly suitable for work on transistor circuits. The input probe is compact, easily handled, and imposes a load of only 2pF on the circuit under investigation. There are facilities for displaying identically shaped pulses of random separation, as occur, for example, with nuclear detectors; and short pulses of low p.r.f. can be displayed with high brightness. Finally, due to the extensive use of transistors, the instrument is portable and compact.

The L.362 has been developed from a design of the Atomic Energy Research Establishment, Harwell, and makes use of the sampling principle: the complete c.r.t. trace is built-up, as a low-frequency replica of the input, of a series of dots, each of which corresponds to a sample taken from successive repetitions of the input. A charge storage circuit maintains the position of a dot until the succeeding sample is taken. Hence, under most conditions, the c.r.t. beam is 'on' for about 80 per cent of the cycle time, and a display of high brilliance is obtained without post deflexion acceleration.

The functional centre of the instrument is the comparator and the X and Y charge-storage circuits. The output of the X charge-storage circuit is compared with a time-base waveform from the fast sweep generators, and when these are equal the comparator produces a pulse which triggers the sampling circuits. Simultaneously, a pulse is passed to the X charge-storage circuit, thus reducing the stored-charge. The sampling trigger pulse therefore occurs relatively earlier, and the cathode-ray tube spot moves to the left on succeeding repetitions of the input waveform. When the stored charge falls to a predetermined level the circuit is reset and the flyback occurs.

The input to be displayed is applied to the sampling circuit, which is contained in a small, separate probe. The input voltage charges the input capacitance of the instrument, and when the sampling trigger pulse is received this charge is transferred via the charge amplifier to the Y charge-store. The stored charge is thus proportional to the input voltage at the instant of sampling.

A second pulse from the X circuit is passed to the Y charge-store, and the voltage corresponding to the stored charge is then applied to the Y plates of the c.r.t. via the Y deflexion amplifier. The X charge-storage circuit also produces pulses which black out the tube while the spot is moving.

A second input of triggering pulses, which must precede the signal input by between 0·05 and 2μsec, is applied to the trigger delay, and thence to the fast sweep generator, which produces the time-base waveform applied to the compartor.

EE 15 761 *for further details*

IMAGE CONVERTOR POWER UNIT

(Illustrated below)

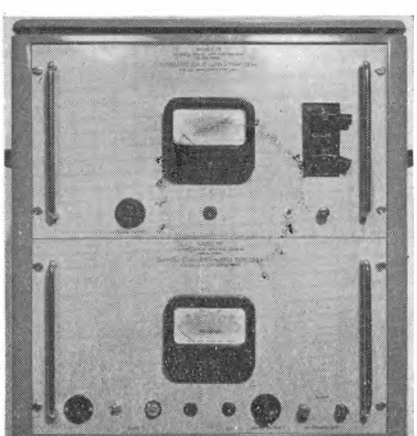
Sionic Ltd, Cromwell House, 93 Wood End Green Road, Hayes, Middlesex.

The stabilized power supply unit type 359A has been designed specifically for the operation of Mullard image convertor tubes.

The current stabilized focus coil unit supplies 200 to 500mA for a nominal load of 700Ω. A change of load by ±100Ω gives a current change of 1 per cent. The ripple is less than 50μA and the mains input and temperature stability better than 1 per cent.

The e.h.t. supply unit provides 5 to 14·5kV at 1μA with a grid supply tapping to give 1μA. Without the grid supply the current available is 100μA. The ripple is less than 1V peak-to-peak and the overall stabilization better than 2 per cent.

EE 15 762 *for further details*



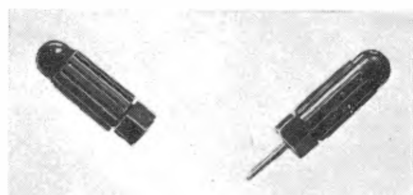
MINIATURE PLUGS AND SOCKETS

(Illustrated below)

Harwin Engineers Ltd, Rodney Road, Portsmouth, Hampshire.

An extremely small plug and socket in Nylon, the Harwin W.3000/3001 can be mounted at only 3/16in centres. It has gold plated contacts which will carry up to 7·5A at 750V, contact resistance averages 2mΩ, insulation is better than 10^{6}Ω. Working conditions up to 100°C can be tolerated and vibration and impact resistances are stated to be excellent.

EE 15 763 *for further details*



IN-LINE DIGITAL DIALS

(Illustrated above right)

Amphenol (Great Britain) Ltd, Ormond House, 26/27 Boswell Street, London, W.C.1.

It is known that forced fast reading and setting of dials is a cause of errors in reading perception and it has been determined that for precise numerical values an inline digital presentation provides the greatest accuracy in conditions where fast reading and setting are required.

These facts have led to the development of the Borg 1300 series Microdials. To obtain the maximum advantages from inline digital presentation, numbers are made in direct contrast to their background, e.g. white on black or black on white. A large, one-piece curved and sealed window provides wide-angle viewing. For maximum sensitivity and accuracy, the control knob is mounted directly on the shaft to be controlled. Backlash, which could result from gearing, is thus eliminated.

Types are available with finger tip brakes to retain settings under conditions of shock and vibration. Available types are:

- Model 1307 3 digit dial counting from 0 to 999 in ten turns of the control knob and fitted finger tip brake.
- Model 1308—as above, without finger tip brake.
- Model 1304—4 digit dial, counting from 0 to 9999 in 100 turns.

- Model 1305—5 digit dial, counting from 0 to 99999 in 1 000 turns.

EE 15 764 *for further details*

V.H.F. SIGNAL GENERATOR

(Illustrated below)

Airmec Ltd, High Wycombe, Buckinghamshire.

A recently announced addition to the range of electronic instruments manufactured by Airmec Limited is the v.h.f. signal generator type 204.

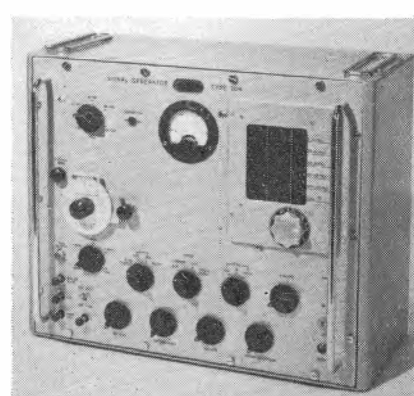
This instrument has a wide frequency coverage from 1Mc/s to 320Mc/s in five ranges. The operating frequency is shown on an illuminated film scale with a length of 4ft on each range, and a crystal check oscillator is incorporated to ensure absolute accuracy of calibration.

A notable feature is the stabilization of the output level at any value between 2μV and 220mV, which avoids the tedious checking and resetting of the output when a number of different frequencies are in use.

An internal 1kc/s oscillator is provided for amplitude or frequency modulation. Depth of modulation up to 50 per cent, or deviation up to 80kc/s, is monitored by a built-in meter to an accuracy of ±5 per cent. External modulation at any frequency from 30c/s to 15kc/s may also be applied either separately or simultaneously with the internal modulation. External pulse modulation with pulse repetition frequencies from 30c/s to 50kc/s may also be applied.

A constant impedance 52Ω step attenuator is fitted, and a matching cable is supplied to transform the source impedance to 75Ω when required.

EE 15 765 *for further details*



Meetings this Month

THE BRITISH COMPUTER SOCIETY LIMITED

Date: 25 November. Time: 6.15 p.m.
Held at: Northampton College of Advanced Technology, St. John Street, London E.C.1.
Lecture: The Metrovick 1010 Data Processing System.
By: J. C. Gladman.

Birmingham Branch
Date: 9 December. Time: 6.30 p.m.
Held at: Large Theatre, Electrical Engineering Department, The University, Edgbaston.
Lecture: Bank Automation.

Hull Section
Date: 2 December. Time: 7.30 p.m.
Held at: Hull Chamber of Commerce, Samman House, Bowlally Lane, Hull.
Lecture: Bankers' Approach to Electronc Applications.
By: R. Hindle.

Liverpool Branch
Date: 24 November. Time: 6.30 p.m.
Held at: Mathematical Institute, University of Liverpool.
Lecture: Choosing and Installing a Computer—A Case Study.
By: A. J. Platt.

Manchester Branch
Date: 1 December. Time: 7 p.m.
Held at: Manchester College of Science and Technology, Sackville Street, Manchester.
Lecture: A Large Data Processing System.
By: W. E. Scott.

Newcastle Branch
Date: 1 December. Time: 7 p.m.
Held at: University Computing Laboratory, 1 Kensington Terrace, Newcastle.
Lecture: Linear Programming and its Applications.
By: A. Muir.

THE BRITISH INSTITUTION OF RADIO ENGINEERS

All London meetings will be held at the London School of Hygiene and Tropical Medicine, Keppel Street, Gower Street, London W.C.1.

Computer Group Meeting
Date: 10 December. Time: 6.30 p.m.
Lecture: The Simulation of Nuclear Reactors and Power Plants.
By: W. J. G. Cox and J. Dowsing.

North Western Section
Date: 10 December. Time: 6.30 p.m.
Held at: Reynolds Hall, College of Technology, Sackville Street, Manchester.
Lecture: Learning Machines.
By: P. Huggins.

South Midlands Section
Date: 27 November. Time: 7 p.m.
Held at: North Gloucestershire Technical College, Cheltenham.
Lecture: A Vidicon Television Camera Channel
By: B. J. Pover.

THE INSTITUTE OF PHYSICS

Date: 11 December. Time: 6 p.m.
Held at: The Institute of Physics, 47 Belgrave Square, London S.W.1.
Lecture: The Ultrasonic Camera—An Alternative Approach to Ultrasonic Testing.
By: J. F. Sayers.
Date: 11 December. Time: 6.15 p.m.
Held at: University College, Cardiff.
Annual General Meeting followed by:
Lecture: Nuclear Magnetic Resonance in Liquids.
By: Dr. J. A. Pople.

THE INSTITUTION OF ELECTRICAL ENGINEERS

All London meetings will be held at Savoy Place, commencing at 5.30 p.m.

Electronics and Communications Section
Date: 25 November.
Lecture: Radio Aspects of the International Geophysical Year.
By: R. L. Smith-Rose.
Date: 7 December.
Lecture: Frequency Patterns for Multiple-Radio-Channel Routes.
By: B. B. Jacobsen.

Measurement and Electronics Sections
Date: 1 December.
Lecture: Frequency Variations of Quartz Oscillators and the Earth's Rotation in Terms of the N.P.L. Caesium Standard.
By: L. Essen, J. V. L. Parry and J. McA. Steele.

Ordinary Meeting
Date: 3 December.
Lecture: The Transmission of News Film over the Transatlantic Cable.
By: C. B. B. Wood and J. J. Shelley.

Cambridge Electronics and Communications Group
Date: 8 December. Time: 8 p.m.
Held at: Cavendish Laboratory, Free School Lane, Cambridge.
Lecture: The Reliability of Components.
By: G. W. A. Dummer.

North-Eastern Measurement and Electronics Group
Date: 7 December. Time: 6.15 p.m.
Held at: Rutherford College of Technology, Newcastle-on-Tyne.
Lecture: Dielectric Materials—Trends and Prospects.
By: C. G. Garton.

North Lancashire Sub-Centre
Date: 9 December. Time: 7.15 p.m.
Held at: North-Western Electricity Board Lecture Theatre, Jubilee Street, Blackburn.
Lecture: Subscriber Trunk Dialling.
By: D. A. Barron.

North Midland Centre
Date: 1 December. Time: 6.30 p.m.
Held at: Bradford Institute of Technology.
Lecture: A System for the Automatic Recognition of Patterns.
By: R. L. Grimsdale, E. W. Sumner, C. J. Tunis and T. Kilburn.

North Scotland Sub-Centre
Date: 10 December. Time: 7 p.m.
Held at: Electrical Engineering Department, Queen's College, Dundee.
Lecture: Automatic Digital Recording of Experimental Values.
By: W. H. P. Leslie.
Date: 11 December. Time: 7.30 p.m.
Held at: Robert Gordon's Technical College, Aberdeen.
Repeat of above lecture.

Rugby Sub-Centre
Date: 2 December. Time: 6.30 p.m.
Held at: Rugby College of Technology of Arts.
Lecture: A System for the Automatic Recognition of Patterns.
By: R. L. Grimsdale.

South Midland Centre
Date: 4 December. Time: 6 p.m.
Held at: Midland Institute, Birmingham.
Lecture: The Jodrell Bank Radio Telescope.
By: H. C. Husband.
Date: 7 December. Time: 6 p.m.
Held at: College of Technology, Birmingham.
Lecture: Some Industrial Applications of Electro Heat.
By: E. May.

Tees-Side Sub-Centre
Date: 2 December. Time: 6.30 p.m.
Held at: Cleveland Scientific and Technical Institution, Middlesbrough.
Lecture: Subscriber Trunk Dialling.
By: D. A. Barron.

THE INSTITUTION OF MECHANICAL ENGINEERS

Date: 25 November. Time: 6 p.m.
Held at: The Institution, 1 Birdcage Walk, Westminster, London S.W.1.
Thomas Hawksley Lecture: The Effect of Nuclear Radiation on Engineering Materials.
By: Professor A. H. Cottrell.

THE JUNIOR INSTITUTION OF ENGINEERS

Date: 11 December. Time: 7.30 p.m.
Held at: Pepys House, 14 Rochester Row, Westminster, London S.W.1.
Inaugural Meeting:
Induction of Professor J. M. Kay as President 1959-60 and presentation of Awards won during the 1958-59 Session followed by an address: The Engineer's Role in Nuclear Power.

THE RADAR AND ELECTRONICS ASSOCIATION

Date: 24 November. Time: 7.30 p.m.
Held at: The Royal Society of Arts, John Adam Street, Adelphi, London W.C.2.
Lecture: Waveguides for Long Distance Communications.
By: Professor H. M. Barlow.

THE TELEVISION SOCIETY

Date: 3 December. Time: 7 p.m.
Held at: Cinematograph Exhibitors' Association, 164 Shaftesbury Avenue, London W.C.2.
Lecture: Television in Germany.
By: Dr. R. Möller.

WOMEN'S ENGINEERING SOCIETY

Date: 9 December. Time: 7 p.m.
Held at: ' Hope House ', 45 Great Peter Street, Westminster, London S.W.1.
Lecture: Radar and Telecommunications Research and Development.
By: Dr. E. Laverick.

THE SOCIETY OF INSTRUMENT TECHNOLOGY

Data Processing Section
Date: 7 December. Time: 7 p.m.
Held at: Manson House, 26 Portland Place, London W.1.
Lecture: Discontinuous Skill Teaching Machines.
By: G. Pask.

Manchester Section
Date: 8 December. Time: 6.45 p.m.
Held at: ' Manchester Room ', Central Library, St. Peter's Square, Manchester 1.
Lecture: Instrumentation of the Dounreay Reactor.
By: K. R. Sandiford.

Midland Section
Date: 11 December. Time: 7 p.m.
Held at: Lecture Theatre, Byng Kendrick Suite, The Gosta Green College of Technology, Aston Street, Birmingham.
Lecture: Radioactive Isotopes in Industry.
By: W. G. Busbridge.

South Wales Section
Date: 25 November. Time: 6.45 p.m.
Held at: Newport & Monmouthshire College of Technology, Newport.
Lecture: Transistors.
By: S. S. Goldberg.

Tees-Side Section
Date: 10 December. Time: 7.30 p.m.
Held at: Cleveland Scientific & Technical Institute, Corporation Road, Middlesbrough.
Lecture: Practical Applications of Strain Gauge Weighing.
By: E. I. Lowe.

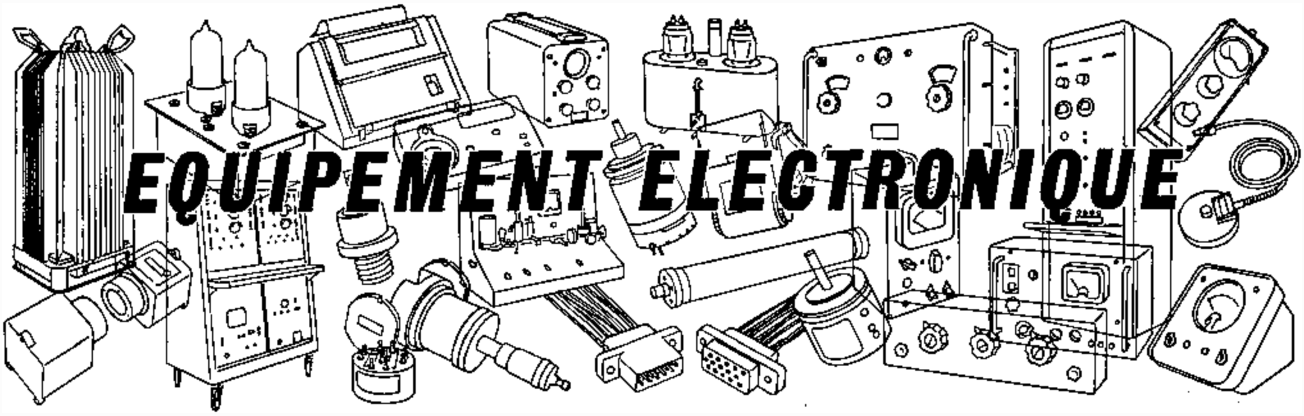
PUBLICATIONS RECEIVED

INSULATION FOR SMALL TRANSFORMERS —A GUIDE TO DESIGN AND TESTING by J. H. Mason and C. G. Garton is a handbook providing guidance and data for designers of all forms of electrical insulation, and will be useful for assessing the applications of new materials, and the insulation requirements for onerous or unusual service conditions. Although largely based on experience at the Electrical Research Association, this book presents data from many other sources and 136 references are given. The Electrical Research Association Laboratory, Cleeve Road, Leatherhead, Surrey. Price 37s. 6d. plus postage.

TRAINING SERVICES IN INDUSTRY is a booklet just published by the Industrial Training Council describing what some employers' organizations and joint bodies are doing in the way of providing training services for the benefit of the firms in their industries. The Council was set up a year ago to keep under review the recruitment and training of workpeople and to encourage, assist and inform industries about these matters. This story of the development of training schemes and advisory services describes the main functions of these services under the headings: Provision of information; Practical training and training aids; Supervision of training and further education; Recruitment; Selection and induction; Indenture and Registration; Tests and certificates and research into training methods. Industrial Training Council, 36 Smith Square, London S.W.1. Price 1s. 6d.

SIEMENS HALL GENERATORS FOR ALL PROBLEMS OF MODERN ELECTRONICS and PHOTOELECTRIC COMPONENTS OF HIGH LIGHT-SENSITIVITY are two catalogues on new Siemens & Halske devices. Copies can be obtained from their authorized agents and distributors R. H. Cole (Overseas) Ltd, 2 Caxton Street, Westminster, London S.W.1.

THE CALCULATION OF THE MEDIAN SKY WAVE FIELD STRENGTH IN TROPICAL REGIONS is a special report by the Department of Scientific and Industrial Research, published by Her Majesty's Stationery Office, price 2s. 6d. The analysis given in this report is presented in a form in which it is reasonably simple to identify the type of ionospheric reflection which is likely to be effective in particular circumstances.



EQUIPEMENT ELECTRONIQUE

Une description basée sur des renseignements fournis par les fabricants de nouveaux organes, accessoires et instruments d'essai *Traduction des pages 698 à 701*

BLOC D'ALIMENTATION STABILISÉE

(Illustration à la page 698)

Advance Components Ltd, Roebuck Road, Hainault, Ilford, Essex

Le bloc d'alimentation stabilisée Advance Type L.101 est un appareil à usage spécial d'une souplesse considérable, qui fournit une source constante de tension positive de 600 volts c.c. et deux alimentations négatives stabilisées de 150 volts, dont l'une est une ligne variable ayant une haute source d'impédance. Deux alimentations de 6,3 volts, 4A, c.a. peuvent également être fournies.

Une ligne négative stabilisée variable de 0 à 150 volts, d'une haute impédance de source est prévue comme alimentation de polarisation de grille ou pour l'utilisation dans des circuits portant une faible charge de courant, ne dépassant pas 5 mA. Une ligne non-stabilisée de 100 à 800 volts est également prévue pour l'usage général.

Les deux lignes positives ont une charge nominale maximum de 300 mA au total. La sortie positive de 600 volts c.c., ainsi que la sortie négative de 150 volts c.c., peuvent être employées en série pour fournir une sortie totale de courant continu stabilisé de 750 volts, avec un régime maximum de 30 mA.

La variation de tension sur la ligne positive de 600 volts et sur la ligne négative de 150 volts est supérieure à 0,03% de la sortie, pour un changement de tension d'entrée de 10% et la variation de sortie des deux lignes de 6,3 volts c.a. est inférieure à 0,02 volts dans les mêmes conditions.

L'impédance de sortie est inférieure à 0,1 ohm sur courant continu et inférieure à 0,5 ohm à 50 kHz sur toutes les lignes de courant continu. La ligne négative variable de 0 à 150 volts c.c. a une impédance de source de 12 kiloohms. La tension d'ondulation est inférieure à 3 mV efficaces sur toutes les sorties stabilisées, à 150 volts et au-dessous. Au dessus de 150 volts, l'ondulation est inférieure à 0,002% de valeur efficace, et sur le courant continu non stabilisé de 100 à 600 volts elle est d'environ 3 volts efficaces à 800 volts, avec une charge de 300 mA.

Deux excellents enregistreurs indiquent les mesures effectuées, le voltmètre étant étalonné de 0 à 250 volts et de 0 à 1 kV, et l'ampèremètre de 0 à 50 mA et de 0 à 500 mA.

Les avantages multiples qu'offre l'appareil comprennent la possibilité de contrôler la ligne de courant continu stabilisé haute tension, de façon à régler la sortie à n'importe quel niveau entre 0 et 600 volts, ainsi qu'une commande de la résistance de source dans la gamme de 0 à 40 ohms. On peut également superposer la tension de ronflement à la ligne de courant stabilisé h.t. et la faire varier entre 0 et 6 volts efficaces afin d'obtenir une onde sinusoïdale satisfaisante à la fréquence principale d'entrée.

Toutes les sorties sont flottantes par rapport au châssis, les lignes négatives et positives étant reliées à travers la borne neutre.

EE 15 751 *pour plus amples renseignements*

OSCILLOSCOPE D'IMPULSIONS

(Illustration à la page 698)

Furzehill Laboratories Ltd, 57 Clarendon Road, Watford, Hertfordshire.

L'Oscilloscope Furzehill 0,140, actuellement en pleine production, permet l'examen d'impulsions et de hautes fréquences à un prix compétitif. Il comprend une base de temps qui peut être employée soit par déclenchement soit en marche libre dans une gamme de 0,1 µsec/cm à 0,5 sec/cm. Une caractéristique d'importance de l'appareil est constituée par le réglage précis du déclenchement qui permet de commencer le balayage horizontal à n'importe quel point voulu de la forme d'onde sous examen. Pour le réglage périodique, le synchronisme positif et immédiat est effectué par un amplificateur à part pouvant être alimenté soit par l'amplificateur intérieur soit par une source extérieure.

L'amplificateur de déviation est à couplage direct et il comporte 8 sensibilités étalonnées, soit 50, 15, 5, 1.5, 0,5 et 0.15V/cm de niveau à 10MHz, plus deux autres sensibilités de 50 mV/cm à 3,3MHz et 15mV/cm à 1MHz. Etant donné que l'instrument est destiné à l'examen d'impulsions, les amplificateurs sont à amortissement critique afin d'éviter toute sonnerie.

La précision de l'étalonnage du temps et de la tension est supérieure à 5% mais une onde intérieure d'étalonnage d'une précision de 1% permet la vérification à tout moment.

Un tube à extrémité plate de 4 pouces de diamètre est utilisé avec commande astigmatique, cependant que le micromètre est muni d'éclairage variable, ce qui facilite la mesure rapide et exacte.

EE 15 752 *pour plus amples renseignements*

SECTIONS À FENTES DE GUIDE D'ONDES

(Illustration à la page 698)

Distributeurs: Leland Instruments Ltd, Abbey House, Victoria Street, London, S.W.1.

Ces instruments, qui sont fabriqués par la Société Polytechnic Research Development Co. de New-York, permettent des mesures précises d'ondes stationnaires et d'impédance dans une gamme d'hyperfréquences allant de 2,6 à 40 kHz par seconde. On peut les utiliser tant pour les essais de construction que pour ceux de service, afin d'obtenir le taux d'ondes stationnaires et les données de phase sur les composantes de guide d'ondes. Ils peuvent être employés aussi, conjointement avec des dispositifs d'accord, pour éliminer les désaccords de lignes de transmission. En outre, ils peuvent servir de pièces permanentes de radar pour vérifier les caractéristiques de réflexion de l'antenne. La gamme de fréquences est 'couverte' par huit formats de guides d'ondes standard.

La section à fente consiste en une longueur de guide d'ondes, usinée avec précision à partir d'un bloc de métal solide, avec une étroite fente longitudinale au centre de la paroi supérieure. Un chariot est prévu pour le déplacement de la sonde de captage le long de la fente. Ce chariot est porté par trois roulements à billes, roulant sur pistes cannelées, durcies et minutieusement meulées, fixées sur le corps même de la section. Le mouvement du chariot peut être réglé par rapport au corps principal de façon à ce que le bout de la sonde suive parallèlement la surface intérieure du guide d'ondes.

Les chariots de sonde ont été calculés pour un mouvement régulier et sûr, sans saccades ou jeu. L'absence de contacts glissants réduit l'usure au minimum et l'instrument garde sa précision indéfiniment.

Afin d'assurer l'exactitude des mesures de phase, chaque section à fentes est munie d'une échelle centimétrique vernier qui indique la distance entre la sonde et l'extrémité extérieure. La course de la sonde couvre, à peu près, une longueur d'onde du guide, à la plus basse fréquence dans chaque gamme, sauf dans les deux lignes de fréquence inférieures où elle couvre les trois-quarts d'une longueur d'onde. Les dimensions intérieures du guide d'ondes sont observées avec précision afin de fournir un étalon d'impédance. Les extrémités des fentes sont diminuées en vue de réduire le taux d'ondes stationnaires résiduel à moins de 1,01, sauf dans les deux lignes de fréquence inférieures où le discontinuité est insignifiante.

EE I5 753 *pour plus amples renseignements*

PONT DE CONDUCTIVITÉ

(Illustration à la page 699)

A. M. Lock & Co. Ltd, Prudential Buildings, 79 Union Street, Oldham, Lancashire

Le pont de conductivité portatif LOCK Type BC1 met en oeuvre les tout derniers circuits à transistors et les techniques de construction les plus récentes afin de garantir un fontionnement sans panne et de longue durée en toutes circonstances. Il est logé dans un coffret en bois dont les dimensions extérieures sont d'environ 305 mm × 190 mm × 140 mm et il est fourni complet avec cellule de conductivité du type à immersion.

La simplicité de fonctionnement et de lecture est assurée par des commandes judicieusement disposées et clairement indiquées sur le panneau incliné avant. Un couvercle de bois facilement rabattable protège l'appareil pour les besoins momentanés. En cours d'emploi, on l'enlève complètement.

Le pont de mesure à courant alternatif BC1 est alimenté à un faible potentiel par un oscillateur à transistors de 1,5 kHz. Cette haute fréquence est utilisée pour éviter les erreurs qui pourraient se produire autrement avec des solutions à haute conductivité.

Durant l'emploi, l'équilibre de pont est indiqué par une déviation maximum du compteur indicateur. La conductivité de la solution est ensuite lue sur un grand cadran étalonné. Grâce à l'utilisation de la méthode du point d'équilibre, les variations de tension de la batterie ou des caractéristiques des transistors n'affectent pas la précision de la mesure.

L'appareil a une portée de mesure s'étendant de 0,5 micromhos à 1 × 166 micromhos en six gammes.

Il convient à l'emploi avec presque tous les modèles de cellules de conductivité. Si la constante de la cellule n'est pas l'unité, on obtiendra la conductivité effective d'une solution en multipliant la lecture du pont par la constante de la cellule.

La compensation de température manuelle dans la gamme de 5 à 35°C est obtenue dans les gammes 2 et 3 à raison de 2%/O°C. Le facteur de correction s'applique à certains condensés et à certaines eaux non traitées, ainsi qu'à presque toutes les solutions aqueuses dont les valeurs de conductivité vont de 5 à 1000 micromhos.

Dans d'autres gammes, le compensateur de température n'est pas branché car la correction nécessaire varie selon la composition et la concentration de l'électrolyte.

Un condensateur variable incorporé équilibre la cellule et le câble de capacité qui donneraient, autrement, un point de compensation plat.

EE I5 754 *pour plus amples renseignements*

ÉLECTROMÈTRE À LAME VIBRANTE

(Illustration à la page 699)

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

L'électromètre à lame vibrante type N616 a été conçu pour la mesure de courants continus très faibles et de tensions à haute stabilité O. L'entrée de courant continu est transformée en courant alternatif au moyen d'un condensateur à lame vibrante et elle est amplifiée par un amplificateur de courant alternatif stable.

Sept gammes ont été prévues: 3, 10, 30, 100 et 300 mV, ainsi que IV et 3V. La sortie est redressée et mesurée sur un enregistreur à panneau de 13 cm, étalonné pour l'indication de la tension c.c. à l'entrée.

Pour les mesures de tension, la précision est à peu près de ± 5%, les mesures de courant pouvant être effectuées avec le même degré de précision, à condition de connaître la valeur totale de la résistance d'entrée et de régler la commande d'étalonnage lorsque cela est nécessaire.

EE I5 755 *pour plus amples renseignements*

GÉNÉRATEUR DE FONCTIONS NON-LINÉAIRES

(Illustration à la page 699)

The Solartron Electronic Group Ltd, 45 Thames Street, Kingston, Surrey.

La Société Solartron vient d'annoncer la réalisation du Générateur de fonctions à diodes Type TR829. De format standard, cet appareil a été étudié pour produire des fonctions précises allant d'un degré variable à un haut degré d'exactitude et de stabilité. Il s'ajoute à la série existante de calculateurs analogiques Solartron.

La fonction requise est produite en 12 segments, obtenus simplement en ajustant des points d'interruption pour chaque segment en fonction d'un contrôleur de voltmètre numérique. Ces ajustages s'effectuent avec des potentiomètres hélicoïdaux à haute résolution, dans la gamme de tensions 0 à plus ou moins de 100V. On peut employer conjointement, si l'on veut, deux appareils TR829 pour des fonctions plus compliquées.

Le générateur de fonctions à diodes englobe au total 14 amplificateurs c.c. à grand gain et ne nécessite pas d'amplificateurs de fonctionnement supplémentaires pour fournir un plein rendement. Les erreurs de dérive et de gain pour chaque segment sont retenues à l'intérieur de 100 mV et de 0,2% respectivement, et, puisque ces erreurs ne sont pas cumulatives dans toute la fonction produite, la stabilité générale à long terme est toujours supérieure à 0,5% en moyenne.

EE I5 756 *pour plus amples renseignements*

TRANSDUCTEUR À TRANSFORMATEUR DIFFÉRENTIEL

(Illustration à la page 699)

James Scott & Co. (Electrical Engineers) Ltd, 68 Brockville Street, Carntyne Industrial Estate, Glasgow, E.2.

Dans l'appareil de base LINTRAN, l'entrée est constituée par le mouvement d'un axe, tandis que la sortie électrique est absolument proportionnelle à ce mouvement. Il n'est, cependant, pas limité à cette seule fonction, puisque, par l'adjonction d'un soufflet étalonné, il devient un manomètre. En fait, n'importe quelle quantité dans laquelle un changement peut être traduit en un mouvement, peut être mesurée avec précision au moyen de cet appareil.

Le LINTRAN utilise le principe bien connu du transformateur différentiel. Ce dernier a deux bobinages espacés avec précision par rapport à un bobinage central, de sorte que les signaux des bobinages sont égaux. Si l'on insère une armature métallique dans ces bobinages, le mouvement de cette armature provoquera des changements des champs magnétiques et les sorties ne seront plus égales. Par une construction minutieuse, on pourra rendre la différence de sortie étroitement proportionnelle au déplacement de l'armature de sa position centrale. La différence de sortie est normalement indiquée par un appareil de lecture qui pourra être gradué directement en fonction du mouvement de l'armature. On pourra faire varier, dans cette construction, la grandeur de l'armature insérée et, selon cette grandeur, on obtiendra sur l'appareil de lecture une indication du mouvement de grandeur naturelle, pour un mouvement de l'axe d'entrée pouvant se réduire à un millième de pouce ou s'élever à un pouce.

Les transducteurs antérieurs utilisant le dispositif à transformateur différentiel avaient tendance à être trop grands et encombrants lorsqu'ils étaient destinés à l'usage d'un secteur normal. Ou bien alors, il fallait leur ajouter des circuits électroniques coûteux pour obtenir la sortie précise voulue. Le LINTRAN, en revanche, tout en ayant été étudié pour le fonctionnement sur secteurs normaux, a gardé les avantages d'une grandeur réduite et d'une sortie précise.

L'appareil de base a 184,15 mm de long par 44,5 mm de diamètre et il ne pèse que 907 grammes.

EE I5 757 *pour plus amples renseignements*

BLOC D'ALIMENTATION À BASSE TENSION

(Illustration à la page 700)

Allied Electronics Ltd, 28 Upper Richmond Road, Putney, S.W.15.

Le bloc d'alimentation LS 353-H est destiné à l'emploi dans les cas nécessitant une source compacte de courant continu stabilisé, tels que pour l'alimentation de chauffage d'amplificateurs c.c., pour les lampes d'excitation d'instruments de mesure photoélectriques ainsi que pour les autres applications où l'on emploie généralement des batteries.

L'élément de base est un circuit redresseur à jonction au germanium avec stabilisateur de tension à transistors et dispositif potentiel de référence.

Le système précité dérive des chargeurs de potentiel constant 'ALTRON' fournis à l'usage des accumulateurs hermétiquement scellés au mercure-cadmium 'DEAC'. Les chargeurs 'DEAC' sont livrables pour une gamme de tensions comprise entre 1,5V et 24V et pour des courants de 100A ou davantage.

Le circuit assure un haut degré de stabilisation contre les changements de tension d'entrée ainsi qu'une faible composante ondulée. Une commande préréglée permet de varier la tension d'entrée dans une petite gamme.

Les blocs d'alimentation sont fournis soit sous forme de châssis soit en coffrets.

EE 15 758 *pour plus amples renseignements*

TACHOMÈTRE SENSIBLE AUX CAPACITÉS

(Illustration à la page 700)

Grunther Instruments Ltd, 14 Oriental Street, London, E.14.

Ce tachomètre sensible aux capacités a été conçu à partir d'un projet pour la mesure de vitesses de turbines à gaz allant jusqu'à 120.000 tours/minute dans des températures locales atteignant 700°C.

L'appareil utilise un palpeur placé près—mais non en contact—de toute irrégularité sur une pièce rotative, tournant à la vitesse qui doit être mesurée ou à un multiple ou sous-multiple de cette vitesse. Cette irrégularité peut être une mortaise, un ergot, une came, une dent d'engrenage, un grand écrou hexagonal, ou bien encore un trou ou une saillie n'ayant pas moins de 1,5 mm de diamètre ou de profondeur.

Chaque fois qu'une irrégularité (saillie ou dépression) passe près du palpeur, il se produit un changement de capacité. Ceci provoque une impulsion de tension qui est transmise, au moyen d'un câble flexible de faible diamètre, à l'instrument de comptage pouvant se trouver à quelques centaines de mètres de distance du palpeur.

Le tachomètre proprement dit est un instrument de comptage électronique utilisant une technique longuement éprouvée. L'indicateur est un appareil de mesure à cadre mobile étalonné directement en tours/minute. Il comporte six gammes de lecture permettant, par l'emploi de facteurs de correction en certains cas, d'effectuer des mesures allant de moins de 50 tours/minute à 200.000 tours/minute, avec une précision supérieure à 2% de déviation totale.

Le palpeur, quoique sensible aux changements de capacité, n'est pas affecté par des extrêmes de température, par les vibrations ou la présence d'huiles, etc.

EE 15 759 *pour plus amples renseignements*

ALIMENTATION À DOUBLE BASSE TENSION STABILISÉE

(Illustration à la page 700)

Advance Components Ltd, Roebuck Road, Hainault, Ilford, Essex.

Le bloc P.P.3 comporte deux sections transistorisées d'alimentation en courant continu stabilisé, de 0 à 30V 1A, contenues dans un coffret fonctionnant à partir d'un transformateur de secteur ordinaire. Les deux sections sont flottantes et peuvent être utilisées séparément, soit reliées en série pour donner une sortie maximum de 60V à 1A, soit pour donner simultanément des tensions positives et négatives (30V max. chacune).

Chaque section couvre une gamme de 0 à 30V en trois étapes se chevauchant de 10V, avec la commande variable étalonnée directement en volts. En outre, l'instrument comprend un voltmètre et un ampèremètre qui peuvent être branchés indépendamment sur chaque section, l'ampèremètre étant à deux gammes.

Quatre réseaux de sortie aux bornes permettent d'éliminer virtuellement la résistance des conducteurs de connexion à la charge.

Le disjoncteur à maxima est constitué par un circuit électronique assurant une protection totale contre les surcharges progressives ou soudaines (court-circuit). Ce circuit électronique peut être réenclanché à partir du panneau avant au moyen de l'interrupteur arrêt-marche. Enfin, grâce à un fusible monté sur panneau, dans le circuit de secteur, l'instrument est protégé contre toute panne intérieure ou surtension de secteur.

EE 15 760 *pour plus amples renseignements*

OSCILLOSCOPE DE CONTRÔLE D'IMPULSIONS

(Illustration à la page 700)

Mullard Equipment Ltd, Mullard House, Torrington Place, London W.C.1

La Société Mullard vient de réaliser un nouveau modèle bon marché d'oscilloscope de contrôle d'impulsions, avec temps de montée rapide de 2 mμsec, équivalent à une largeur de bande allant du courant continu à 220 MHz.

Cet appareil, qui porte la désignation L.362, a été étudié pour la présentation ou la mesure d'impulsions d'une durée de 3 mμsec à 3 μsec, et il sera donc des plus utiles pour les travaux de recherches nucléaires et de calcul électronique ainsi que pour l'examen d'impulsions de radar, particulièrement en ce qui concerne les appareils à portée réduite et à haute résolution.

En plus de sa largeur de bande très étendue, le Modèle L.362 comporte plusieurs autres avantages d'importance. Ainsi, sa sensibilité est de 150 mV/cm ce qui, ajouté à ses amplificateurs à couplage direct, le rend particulièrement indiqué pour les travaux sur les circuits à transistors. Le palpeur d'entrée est compact, aisément maniable, et la charge qu'il impose sur le circuit en cours de contrôle n'est que de 2 pF. L'instrument permet aussi la représentation d'impulsions de forme identique à séparation intermittente, telles que celles qui se produisent, par exemple, dans les détecteurs nucléaires. De plus, les images d'impulsions courtes à fréquence réduite, peuvent être présentées fort clairement. Enfin, grâce, à l'utilisation étendue de transistors, le L.362 est un instrument portatif des plus compacts.

Conçu d'après un dessin des Etablissements de Recherches Atomiques de Harwell, le L.362 applique le principe discriminateur ou "d'échantillonnage", l'oscillogramme entier étant formé, en tant que reproduction à basse fréquence de l'entrée, d'une série de points dont chacun correspond à un "échantillon" pris des répétitions successives de l'entrée. Un circuit d'accumulation de charge maintient la position du point jusqu'à ce que l'échantillon suivant soit pris. Le faisceau cathodique demeure, ainsi, en circuit, dans la plupart des cas, pendant 80% environ de la durée du cycle, et on obtient une image de forte brilliance sans accélération de post-déviation.

Le comparateur et les circuits d'accumulation de charge X et Y constituent le centre fonctionnel de l'appareil. La sortie du circuit d'accumulation X est comparée avec une forme d'onde de base de temps des générateurs à balayage rapide, et lorsque leurs valeurs sont égales, le comparateur produit une impulsion qui déclenche les circuits de prise d'échantillons. Une impulsion est transmise, au même moment, au circuit d'accumulation de charge X, réduisant ainsi la charge accumulée. L'impulsion de déclenchement d'échantillonnage se produit donc relativement plus tôt et les points cathodiques se déplacent à gauche au cours des répétitions successives de la forme d'onde d'entrée. Lorsque la charge accumulée tombe à un niveau prédéterminé, le circuit est réenclanché et le retour du spot se produit.

L'entrée devant être représentée est appliquée au circuit d'échantillonage qui se trouve dans un petit palpeur à part. La tension d'entrée charge la capacité d'entrée de l'appareil, et lorsque l'impulsion de déclenchement d'échantillonnage a été reçue, cette charge est renvoyée, à travers l'amplificateur de charges, à l'élément d'accumulation de charges Y. La charge accumulée est donc proportionnelle à la tension d'entrée au moment de l'échantillonnage.

Une seconde impulsion du circuit X est transmise à l'élément d'accumulation de charges Y et la tension correspondant à la charge accumulée est ensuite appliquée, par le truchement de l'amplificateur de déviation Y, aux plaques Y du tube cathodique. Le circuit d'accumulation de charges X produit également des impulsions qui suppriment le faisceau du tube pendant que le spot se déplace.

Une seconde entrée d'impulsions de déclenchement, qui doit précéder le signal d'entrée de 0,05 à 2 µsec, est appliquée au retardateur de déclenchement, puis au générateur à balayage rapide qui produit la forme d'onde de base de temps appliquée au comparateur.

EE 15 761 *pour plus amples renseignements*

BLOC D'ALIMENTATION CONVERTISSEUR D'IMAGE

(Illustration à la page 701)

Sionic Ltd, Cromwell House, 93 Wood End Green Road, Hayes, Middlesex.

Le bloc d'alimentation stabilisée type 359A a été étudié pour la commande des tubes convertisseurs d'images MULLARD.

L'élément à bobine de focalisation de courant stabilisé fournit 200 à 500 mA pour une charge nominale de 700 ohms. Un changement de charge de ± 100 ohms donne un changement de courant de 1%. L'ondulation est inférieure à 50 µA, tandis que l'entrée secteur et la stabilité de température sont supérieures à 1%.

L'élément d'alimentation très haute tension fournit de 5 à 14,5 kV, à 1 µA, avec prise d'alimentation de grille pouvant donner 1 µA. Sans l'alimentation de grille, le courant disponible est de 100 µA. L'ondulation est inférieure à 1V de crête à crête et la stabilisation globale est supérieure à 2%.

EE 15 762 *pour plus amples renseignements*

FICHES ET PRISES MINIATURE

(Illustration à la page 701)

Harwin Engineers Ltd, Rodney Road, Portsmouth, Hampshire.

La fiche avec prise Harwin W.300/3001, de format extrêmement réduit et en nylon, peut être montée entre pointes de 4,7625 mm seulement. Elle est munie de contacts dorés pouvant porter une tension atteignant 7,5A à 750V. La résistance de contact est en moyenne de 2 mΩ et l'isolement est supérieur à 10^6Ω. Des conditions de fonctionnement allant jusqu'à 100°C peuvent être tolérées et la résistance aux chocs et aux vibrations est, semble-t-il, excellente.

EE 15 763 *pour plus amples renseignements*

CADRANS NUMÉRIQUES "EN LIGNE"

(Illustration à la page 701)

Amphenol (Great Britain) Ltd, Ormond House, 26/27 Boswell Street, London, W.C.1.

On sait que la lecture et le réglage rapides des cadrans sont la cause d'erreurs de lecture et il a été établi que pour obtenir des valeurs numériques précises, la présentation numérique dite "en ligne" garantit le plus haut degré d'exactitude lorsqu'une lecture et un réglage rapides sont nécessaires.

Ces constatations ont conduit à la réalisation des microcadrans de la série Borg 1300. Afin de tirer le maximum d'avantages de la présentation numérique en ligne, les chiffres font contraste direct avec le fond, étant, par exemple, en blanc sur fond noir ou en noir sur fond blanc. Une grande fenêtre courbe et scellée offre un grand angle de vision. Afin de garantir le maximum de sensibilité et de précision, le bouton de réglage est monté directement sur l'axe. Tout jeu pouvant résulter de l'engrenage est donc éliminé.

Il existe des modèles avec frein de bout de dogt pour maintenir les réglages en cas de chocs ou vibrations. Les modèles livrables sont comme suit:

Modèle 1307: cadran à trois numéros, comptant de 0 à 999, en dix tours du bouton de réglage et muni de frein de bout de doigt.

Modèle 1308: comme le précédent, mais sans frein de bout de doigt.

Modèle 1304: cadran à quatre numéros, comptant de 0 à 9999 en 100 tours.

Modèle 1305: cadran à cinq numéros, comptant de 0 à 99999 en 1000 tours.

EE 15 764 *pour plus amples renseignements*

GÉNÉRATEUR DE SIGNAUX POUR HYPERFRÉQUENCES

(Illustration à la page 701)

Airmec Ltd, High Wycombe, Buckinghamshire.

Le générateur de signaux hyperfréquences type 204 est venu s'ajouter depuis peu à l'assortiment d'instruments électroniques fabriqués par la Société Airmec Limited.

Sa plage de fréquences s'étend de 1 MHz à 320 MHz sur 5 gammes. La fréquence employée est indiquée sur une échelle à film lumineux d'une longueur de 121,9 cm pour chaque gamme, cependant qu'un oscillateur à cristal de contrôle, incorporé à l'instrument, assure la précision absolue de l'étalonnage.

Une des caractéristiques les plus remarquables du générateur type 204 est constituée par la stabilisation du niveau de sortie à n'importe quelle valeur comprise entre 2µV et 220 mV, ce qui rend inutile le contrôle fastidieux et le recalage de la sortie lorsqu'on emploie un nombre de fréquences différentes.

Un oscillateur interne de 1 kHz a été prévu pour la modulation d'amplitude ou de fréquence. On peut contrôler, avec un compteur incorporé, une profondeur de modulation allant jusqu'à 5% ou une déviation atteignant 80 kHz, jusqu'à ± 5% de précision. On peut aussi appliquer une modulation extérieure, soit séparément, soit en même temps que la modulation intérieure, à n'importe quelle fréquence comprise entre 30 Hz et 15 kHz. Enfin, on peut également appliquer une modulation d'impulsions extérieures à des fréquences de répétition d'impulsions allant de 30 Hz à 50 kHz.

L'appareil est muni d'un atténuateur à plots de 52 ohms à impédance constante, tandis qu'un câble d'adaptation transforme l'impédance de source à 75 ohms lorsque cela est nécessaire.

EE 15 765 *pour plus amples renseignements*

Résumés des Principaux Articles

Un système de commutation statique utilisant des tubes à cathodes froides (Système AUSTIN de commutation électronique à cathodes froides) par R. W. Brierley

Résumé de l'article aux pages 646 à 654

L'automatisation très poussée des procédés de production qui s'effectue actuellement a donné lieu à des systèmes de commande fort compliqués. La plupart de ces systèmes sont du type marche/arrêt, utilisant des relais comme éléments logiques.

A partir d'un certain degré de complexité, il devient difficile de maintenir un niveau élevé d'efficacité de fonctionnement avec ces équipements. On a donc cherché des variantes plus sûres du relais électro-mécanique, ce qui a amené la réalisation, depuis peu, de systèmes à dispositifs statiques. Un de Ces systèmes, utilisant des tubes à cathodes froides, est décrit dans cet article.

Un enregistreur vectoriel par K. G. Freeman

Résumé de l'article aux pages 655 à 658

Le Vectorscope dont il est question dans cet article donne une image vectorielle des signaux de chrominance du type dit NTSC. C'est un appareil des plus sensibles, comportant un nombre d'éléments fort utiles permettant, entre autres, une vérification continue du gain, ainsi que des réglages de quadrature. Grâce au Vectorscope, on peut évaluer rapidement les caractéristiques de phase et d'amplitude de signaux de chrominance, qui sont d'une importance particulière pour le décodage direct des tubes à rayons cathodiques pour télévision en couleur à canon de signalisation.

Semi-conducteurs composés par D. A. Wright

Résumé de l'article aux pages 659 à 665

L'auteur examine. en premier lieu, les principaux paramètres régissant le comportement des semi-conducteurs. Ces considérations sont, ensuite, appliquées aux différents types de semi-conducteurs composés conçus à ce jour et elles sont complétées par des données de valeurs de paramètres moyens. L'utilité des matériaux pour différentes applications est analysée à la lumière de ces chiffres. Pour chaque type de semi-conducteur composé, certaines tendances bien définies ont été observées dans le rapport entre paramètres de semi-conducteurs et autres propriétés physiques. Ces tendances sont définies par l'auteur, qui en indique leur interprétation. Il mentionne brièvement les paramètres d'importance pour les applications thermoélectriques et rappelle, en conclusion, que le choix des matériaux pour ces usages dépend précisément des paramètres cités au début.

Indicateurs de rapport d'amplitude de tension à lecture automatique par M. Kollanyi et R. M. Verran

Résumé de l'article aux pages 666 à 671

Un indicateur automatique donnant une lecture directe, soit sur enregistreur, soit sur appareil à image numérique, peut remplacer avantageusement l'indicateur classique de rapport d'amplitude de tension dont les réglages prennent du temps. Les circuits peuvent être utilisés avec n'importe quelle ligne rainurée ou dispositif équivalent. Un simple mouvement du chariot à sonde est tout ce qu'il faut pour que l'instrument donne une lecture correcte du rapport d'amplitude de tension.

Le système décrit dans cet article est basé sur un détecteur maximum/minimum qui relève les valeurs extrêmes de tension rencontrées durant le cycle d'enregistrement. Le rapport de ces tensions peut être calculé par diverses méthodes électroniques qui sont également décrites.

Une valeur résiduelle de rapport d'amplitude de tension de $<1{,}5$ *peut être obtenue avec une portée maximum de* 1 : 3 *ou davantage, si nécessaire, quoique cette portée dépende de la méthode choisie.*

Les tubes à gaz en tant que générateurs de musique par Alan Douglas

Résumé de l'article aux pages 672 à 673

L'attrait évident que présente l'emploi de deux tubes à gaz à électrodes comme oscillateurs à relaxation semblerait indiquer que de nombreux problèmes de production de musique électrique pourraient être ainsi résolus économiquement. En réalité, toutefois, aucun instrument commercial n'utilise pareils dispositifs. Certains circuits récents tirent le meilleur effet possible des limites physiques inhérentes aux tubes à gaz non coûteux.

Un système de modulation en triangle à transistors à jonction par B. E. Williams

Résumé de l'article aux pages 674 à 680

Description d'un système de modulation en triangle fonctionnant à une fréquence numérique de 14 *kHz et fournissant un lien simple de parole d'une intelligibilité parfaite. L'appareil n'a qu'une seule commance manuelle, soit un bouton-poussoir pour parler, les autres réglages étant préréglés ou s'effectuant automatiquement. Le récepteur régénère le signal capté, par tension et "découpage" de temps, avant de le transmettre au décodeur. L'appareil fonctionne de façon pleinement satisfaisante dans une gamme étendue de températures ambiantes et il convient à l'emploi dans des lignes égalisées ou dans des chaînes de stations-radio. L'alimentation nécessaire à l'appareil est de* 6 *à* 12 *volts.*

Un petit bloc d'alimentation réglée haute tension avec sortie variable par J. D. O'Toole

Résumé de l'article aux pages 681 à 683

Les champ des applications pour l'alimentation h.t. à sortie stable, n'exigeant pas une puissance élevée, s'étend constamment. Il semble, cependant, que la demande dans ce domaine n'ait été qu'imparfaitement satisfaite, en raison, vraisemblablement, d'une certaine insuffisance en ce qui concerne les petites lampes actuelles. L'auteur décrit ici un circuit de base qui obvierait à cette apparente inaptitude, ainsi qu'un circuit-type de ce genre, dont il indique les chiffres de rendement.

Un système pour fournir une tension vectorielle précise par D. J. Collins et J. E. Smith

Résumé de l'article aux pages 684 à 685

Dans de nombreux domaines de l'électronique, on a souvent besoin de produire une tension de phase et d'amplitude connues, se rapportant à quelque niveau de référence. Abstraction faite des dispositifs électro-mécaniques, la précision avec laquelle cette tension est produite dépend dans une grande mesure de la précision des composantes du déphaseur. Cet article traite d'un système pouvant produire une tension vectorielle précise, dont l'amplitude, quelle qu'elle soit, est inférieure au niveau de référence et dont le rapport de phase est de 0° *à* 360°. *Un pareil système est de grande utilité pour l'étalonnage des déphaseurs et des appareils de mesure de phase.*

Une autre méthode pour comparer les propriétés électriques des résines de moulage en époxie par D. H. Thompson

Résumé de l'article aux pages 686 à 687

Cet article décrit une méthode pour évaluer la constante diélectrique, les pertes et la résistivité de volume des résines de moulage d'époxie dans une gamme de températures étendue et pour des fréquences allant jusqu'à 5 *MHz.*

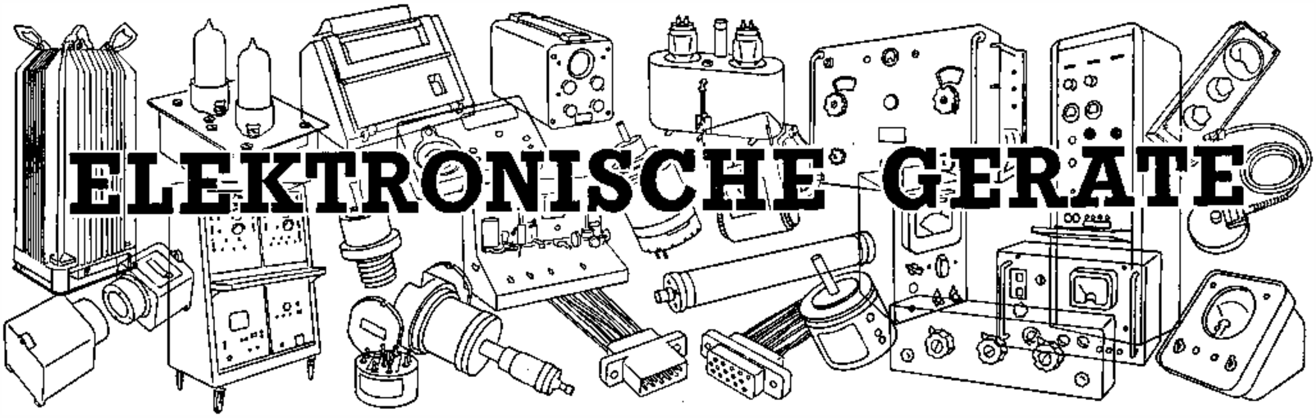
Cette méthode de mesure s'effectue aux dépens de la précision absolue réduite lorsqu'on la compare avec la méthode courante de Hartshorn et Ward. Si, par contre, on l'emploie essentiellement pour comparer des résines et pour évaluer la stabilité de température de leurs paramètres électriques, elle mérite alors un usage plus répandu.

En particulier, des valeurs absolues peuvent être citées avec cette méthode, à condition qu'on prenne soin de vérifier ou de normaliser les résultats aux températures de laboratoire.

Un régulateur basse tension transistorisé par J. H. Deichen

Résumé de l'article aux pages 688 à 689

Cet article traite des méthodes d'alimentation en courant continu stabilisé pour réchaffeurs à lampes et discute leurs limites. Il indique ensuite, un circuit simple utilisant un transistor et deux diodes Zéner dont il décrit la conception et les performances.



ELEKTRONISCHE GERÄTE

Beschreibung neuer Bauelemente, Zubehörteile und Prüfgeräte auf Grund der von Herstellern gemachten Angaben. *Ubersetzung der Seiten 698 bis 671*

STABILISIERTES NETZGERÄT

(Abbildung Seite 698)

Advance Components Ltd, Roebuck Road. Hainault, Ilford, Essex

Advance stellt für Sonderzwecke ein stabilisiertes Netzgerät Typ L 101 grosser Anpassungsfähigkeit her, das eine konstante Spannungsquelle für +600 V und zwei stabilisierte Leitungen für −150 V Gleichstrom hat. Eine der stabilisierten Leitungen ist variabel und hat eine Quelle hoher Impedanz. Zwei Anschlüsse für 6,3 V 4 A stabilisierten Wechselstrom sind auch vorhanden.

Eine stabilisierte, variable, negative Leitung für 0—150 V mit hoher Quellenimpedanz ist für Gittervorspannung und andere Kreise bestimmt, die nicht mehr als 5 mA entnehmen. Für allgemeinen Bedarf ist ein unstabilisierter Ausgang von 100—800 V vorgesehen.

Höchstbelastung für die beiden positiven Leitungen ist zusammen 300 mA. Der +600 V und der −150 V Ausgang können seriengeschaltet als stabilisierter Ausgang von 750 V Gleichstrom für 30 mA Höchstbelastung Verwendung finden.

Für 10% Änderung der Eingangsspannung sind die Spannungsabweichungen der +600 V und −150 V Leitungen geringer als 0,03%. Die Ausgangsschwankungen der beiden 6,3 V Wechselstromleitungen sind unter denselben Bedingungen geringer als 0.02 V.

Die Ausgangsimpedanz für alle Gleichstromleitungen ist bei Gleichstrom unter 0,1Ω und bei 50 kHz unter 0,5Ω. Die Quellenimpedanz der von 0 bis 150 V variablen negativen Leitung ist 12 kΩ. Für stabilisierte Ausgänge bis zu 150 V ist die Effektiv-Restwellenspannung unter 3 mV. Über 150 V ist die Restwellenspannung unter 0,002% eff und für den unstabilisierten Gleichstrom zwischen 100 und 800 V ungefähr 3 V eff, gemessen bei 800 V und 300 mA.

Es ist möglich, die stabilisierte Hochspannungs-Gleichspannung auf jeden gewünschten Pegel zwischen 0 und 600 V einzustellen, wobei gleichzeitig der Quellenwiderstand über den Bereich 0—40Ω regelbar ist. Eine von 0—6 V eff variable Brummspannung kann dem stabilisierten Hochspannungsgleichstrom überlagert werden, um eine hochwertige Sinuswellenform entsprechend der Eingangsnetzfrequenz zu erhalten.

Alle Ausgänge sind in Bezug auf das Chassis erdfrei. Die positiven und negativen Leitungen sind durch eine neutrale Klemme verbunden.

EE IS 751 *für weitere Einzelheiten*

IMPULS-OSZILLOGRAF

(Abbildung Seite 698)

Furzehill Laboratories Ltd. 57 Clarendon Road. Watford, Hertfordshire

Der Furzehill Oszillograf Modell O-140 ist nunmehr in laufender Produktion und ermöglicht Impuls- und HF-Untersuchungen mit geringem Geldaufwand. Der Ablenkteil kann entweder für Triggerung oder Dauerbetrieb im Bereich 0,1 μs/cm ... 0,5 s/cm benutzt werden. Nützlich ist die Feineinstellung der Triggerung, die es ermöglicht, die horizontale Bestreichung an jedem gewünschten Punkt der zu untersuchenden Wellenform zu beginnen. Bei wiederholter Einstellung findet positive und sofortige Synchronisierung durch einen getrennten Synchronisierungsverstärker statt, der entweder von einem eingebauten Verstärker oder einer Fremdquelle gespeist wird.

Der Ablenkverstärker ist direkt gekuppelt und hat acht Empfindlichkeitsstufen: 50, 15, 5, 1,5, 0,5 und 0,15 V/cm bis 10 MHz und zusätzlich 50 mV/cm bis 3,3 MHz und 15 mV/cm bis 1 MHz. Da das Gerät zur Untersuchung von Impulsen entwickelt wurde, sind die Verstärker zur Vermeidung von Eigenerregung kritisch gedämpft.

Die Eichgenauigkeit für Zeit und Spannung ist besser als 5%, kann aber jederzeit gegen eine eingebaute Eichwellenform mit 1% iger Genauigkeit geprüft werden.

Eine 10cm Elektronenstrahlröhre mit flachem Schirm und Anastigmatkorrektur findet Verwendung. Zur Erzielung schneller und genauer Messungen ist die Rasterskalenbeleuchtung variabel.

EE IS 752 *für weitere Einzelheiten*

HOHLLEITERSTÜCKE MIT SCHLITZ

(Abbildung Seite 698)

Vertrieb durch: Leland Instruments Ltd, Abbey House, Victoria Street, London, S.W.1

Diese von Polytechnic Research Development Co. in New York hergestellten Vorrichtungen ermöglichen genaue Messungen stehender Wellen und Impedanzen im Mikrowellenbereich von 2,6 ... 40 GHz. Sie können sowohl für Entwicklungsarbeiten, als auch für laufende Überprüfung der Stehwellenverhältnisse und Phasendaten von Hohlleiter-Bauelementen oder mit Abstimmvorrichtungen zur Vermeidung von Falschanpassungen in Leitungen Verwendung finden. Ausserdem können sie als festeingebaute Elemente zur Überprüfung von Antennenreflexionskennlinien in Radar- und Nachrichtensystemen benutzt werden. Der Frequenzbereich wird durch acht Standard-Leitungsstücke überstrichen.

Das Leitungsstück mit Schlitz besteht aus einem aus Vollmetall sorgfältig hergestellten Hohlleiter mit einem engen Längsschlitz in der Mitte der oberen Wand. Ein Schlitten bewegt den Messkopf über die ganze Länge des Schlitzes. Dieser Schlitten läuft auf drei Kugellagern, die auf gehärteten und mit grosser Genauigkeit geschliffenen Bahnen laufen, die ihrerseits an dem Gerätekörper angebracht sind. Die Schlittenbewegung kann in Bezug auf den Gerätekörper so geregelt werden, dass der Messkopf sich genau parallel zur Innenoberfläche des Hohlleiters bewegt.

Die Schlittenkonstruktion führt zu glattem und genauem Abtasten ohne toten Gang oder Spiel. Die Abwesenheit von Schleifkontakten reduziert die Abnutzung zu einem Minimum. Dadurch wird die Gerätegenauigkeit auf die Dauer erhalten.

Zur Ausführung genauer Phasenmessungen ist jedes Leitungsstück mit Schlitz auch mit einer Zentimeterskala für Feineinstellung versehen, auf der der Abstand zwischen Messkopf und dem Aussenende abgelesen werden kann. Die Länge des Messkopfdurchganges ist für die niedrigste Frequenz jedes Bereiches ungefähr eine Hohlleiterwellenlänge, mit Ausnahme der beiden Leitungsstücke für die niedrigsten Frequenzen, für die er einer dreiviertel Wellenlänge entspricht. Die Innenabmessungen werden genau eingehalten, um eine genormte Impedanz zu erzielen. Mit Ausnahme der beiden Leitungen für die niedrigsten Frequenzen, bei denen die Diskontinuität vernachlässigt werden kann, sind die Schlitzenden abgeschrägt, um das Reststehwellenverhältnis unter 1,01 herunterzubringen.

EE 15 753 *für weitere Einzelheiten*

LEITFÄHIGKEITSMESSBRÜCKE

(Abbildung Seite 699)

A. M. Lock & Co. Ltd, Prudential Buildings, 79 Union Street, Oldham, Lancashire

Die für die tragbare „Lock" Leitfähigkeitsmessbrücke Modell CB 1 eingesetzten neuesten Transistorschaltungen und Herstellungsmethoden garantieren einen langen, störungsfreien Gebrauch in Labor und Aussendienst. Der aus Holz bestehende Tragkoffer hat Aussenabmessungen von 305 × 190 × 140 mm und enthält das mitgelieferte Eintauch-Leitfähigkeitsgefäss.

Sorgfältige Anordnung und klare Beschriftung der Bedienungsknöpfe machen Einstellung und Ablesen leicht. Der schnell abnehmbare Deckel gibt Schutz beim Transport und wird für den Gebrauch entfernt.

Die Wechselstrommessbrücke wird von einem 1,5 KHz Transistor-Oszillator mit geringem Potential gespeist. Die Wahl der hohen Frequenz vermeidet die sonst durch hohe Leitfähigkeit der Lösungen auftretenden Ungenauigkeiten.

Im Gebrauch tritt der Brückenausgleich beim Höchstausschlag des Anzeigegerätes auf. Die Leitfähigkeit der Lösung wird dann auf einer grossen, geeichten Skala abgelesen. Durch Anwendung dieses Nullverfahrens haben Schwankungen der Batteriespannung oder Transistor-Kennlinien keinen Einfluss auf die Messgenauigkeit.

Das Gerät hat sechs Messbereiche für Messungen von 0,5 bis 1×10^6 micromhos.

Fast alle Leitfähigkeitsgefässe können für diese Brücke Verwendung finden. Falls die Gefässkonstante nicht 1 ist, wird die tatsächliche Lösungsleitfähigkeit durch Multiplikation des abgelesenen Wertes mit der Konstanten erreicht.

Für Temperaturen zwischen 5 und 35°C kann in Messbereichen 2 und 3 ein Ausgleich von 2%/°C manuell eingestellt werden. Dieser Ausgleich genügt für gewisse Kondensate und unbehandelte Gewässer, sowie die meisten wässrigen Lösungen mit Leitfähigkeitswerten von 5 bis 1000 micromhos.

Für andere Bereiche ist der Temperaturausgleich nicht eingeschaltet, da sich die notwendige Korrektur mit der Zusammensetzung und Konzentration der Elektrolyte ändert.

Ein eingebauter Drehkodensator dient dem Ausgleich der Gefäss-und Kabelkapazität zur Vermeidung einer flachen Nullabstimmung.

EE 15 754 *für weitere Einzelheiten*

ZUNGENELEKTROMETER

(Abbildung Seite 699)

Ekco Electronics Ltd, Southend-on-Sea, Essex

Das Zungenelektrometer Type N616 wurde zur Messung sehr kleiner Gleichströme und -spannungen mit grosser Nullstabilität entwickelt. Der Gleichstromeingang wird mittels eines Kondensators mit schwingender Zunge in Wechselstrom umgewandelt und in einem stabilen Wechselstromverstärker verstärkt.

Sieben Messbereiche sind vorgesehen: 3, 10, 30, 100 und 300 mV, 1 und 3 V. Der Ausgang wird gleichgerichtet und auf einem 13 cm Schalttafelinstrument angezeigt, das in Gleichstrom-Eingangsspannung geeicht ist.

Die Genauigkeit für Spannungsmessungen beträgt ±5%. Dieselbe Genauigkeit kann für Strommessungen erreicht werden, vorausgesetzt, dass die Gesamtgrösse des Eingangswiderstandes bekannt ist und die Eichung dementsprechend nachgestellt wird.

EE 15 755 *für weitere Einzelheiten*

UNGERADLINIGER FUNKTIONSGEBER

(Abbildung Seite 699)

The Solartron Electronic Group Ltd, 45 Thames Street, Kingston, Surrey

Solartron hat vor kurzem den Dioden-Funktionsgeber TR 829 herausgebracht. Der TR 829 erzeugt mit hoher Genauigkeit und Stabilität genaue Funktionen einer Variablen und ist für Einbau in ein 19 Zoll (482,6 mm) Normalgestell konstruiert. Er erweitert das Solartron Fertigungsprogamm für ungeradlinige Analogrechner-Bausteine.

Die gewünschte Funktion wird in 12 Sektoren erzeugt. Die Einstellung für jeden Sektor erfolgt sehr einfach durch Einregelung des Knickpunktes mit Hilfe des Ziffernvoltmeters eines Monitors. Die Einregelung über den Spannungsbereich 0 ... ±100 V erfolgt mittels eines Heli-Potentiometers hoher Auflösung. Zwei TR 829 Bausteine können zur Erzeugung komplexer Funktionen zusammengeschaltet werden.

Der Dioden-Funktionsgeber enthält insgesamt 14 Hochleistungs-Gleichstromverstärker und benötigt keine zusätzlichen Operationsverstärker für volle Ausgangsleistung. Drift- und Verstärkungsfehler für jeden Sektor werden unter 100 mV bzw. 0,2% gehalten, und da sie nicht kumulativ über die ganze erzeugte Funktion sind, wird die durchschnittliche Gesamtdauerstabilität immer unter 0,5% liegen.

EE 15 756 *für weitere Einzelheiten*

UMWANDLER MIT DIFFERENTIALÜBERTRAGER

(Abbildung Seite 699)

James Scott & Co. (Electrical Engineers) Ltd, 68 Brookville Street, Carntyne Industrial Estate, Glasgow, E.2

Für die Grundform des LINTRAN-Gerätes besteht die Eingabe aus der Bewegung eines Schaftes, der der elektrische Ausgang genau proportional ist. Jedoch ist es nicht auf diese Arbeitsweise beschränkt, da es durch Zusatz von geeichten Balgen als Druckmesser usw. Verwendung finden kann. Tatsächlich eignet sich das Gerät zur Messung jeder Quantität, deren Änderung in Bewegung umgesetzt werden kann.

LINTRAN bringt das bekannte Prinzip des Differentialübertragers zur Anwendung, in dem zwei Wicklungen gegenüber einer Mittelwicklung einen genauen Abstand haben, sodass Signale von beiden Wicklungen gleich sind. Die Bewegung eines Eisenankers innerhalb der Wicklungen ruft Änderungen im inneren magnetischen Feld hervor, und die Ausgänge sind nicht mehr gleich. Bei vorsichtiger Konstruktion können die Ausgangsdifferenzen der Bewegung des Ankers von seiner Mittellage genau proportional gemacht werden. Die Ausgangsdifferenz wird gewöhnlich auf einem Messgerät angezeigt, das direkt in Einheiten der Ankerbewegung geeicht werden kann. In dieser Konstruktion kann die Ankergrösse geändert werden. Es hängt von der Ankergrösse ab, ob das Instrument für eine Schaftbewegung von nur 0,025 mm oder 25,4 mm Vollausschlag anzeigt.

Bisherige, für direkten Netzanschluss konstruierte Umwandler mit Differentialübertrager-Anordnung neigten dazu, gross und umständlich zu sein oder teuere elektronische Kreise zu bedingen, um die geforderte Ausgangsgenauigkeit zu erreichen. LINTRAN arbeitet jedoch mit normalen Netzspannungen, ist klein und hat doch die Ausgangsgenauigkeit beibehalten.

Das Grundgerät ist 184 mm lang und hat einen Durchmesser von 44,5 mm; es wiegt 907 g.

EE 15 757 *für weitere Einzelheiten*

NIEDERSPANNUNGS-NETZGERÄT

(Abbildung Seite 700)

Allied Electronics Ltd, 28 Upper Richmond Road, Putney, S.W.15

Das Netzgerät LS353-H ist eine gedrängt gebaute, stabilisierte Gleichstromquelle für Heizfäden, Belichtungslampen in fotoelektrischen Messgeräten und andere Verwendungszwecke, für die normalerweise Batterien benutzt werden.

Das Grundgerät besteht aus einer Ge-Flächengleichrichterschaltung, einem Transistor-Spannungsstabilisator und einer Bezugsspannungsanordnung.

Das oben beschriebene System ist eine Weiterentwicklung des „Altron" Ladegerätes, das eine konstante Spannung für „DEAC" luftdicht verschlossene Quecksilber-Kadmium-Akkumulatoren liefert.

„DEAC"-Ladegeräte sind für eine Reihe von Spannungen zwischen 1,5 ... 24 V und Stromstärken bis zu 100 A und mehr lieferbar.

Die Schaltung führt zu hochgradiger Stabilisation gegen Schwankungen der Eingangsspannung und geringer Restwelligkeit. Die Ausgangsspannung kann mittels eines Trimmers über einen kleinen Bereich verändert werden.

Die Netzgeräte sind als Chassis oder in Gehäusen lieferbar.

EE 15 758 *für weitere Einzelheiten*

KAPAZITÄTSEMPFINDLICHES TACHOMETER

(Abbildung Seite 700)

Grouther Instruments Ltd, 14 Oriental Street, London, E.14

Dieses kapazitätsempfindliche Tachometer wurde für ein Projekt entwickelt, das Messung von Gasturbinengeschwindigkeiten bis zu 120,000 U/min bei örtlichen Temperaturen bis zu 700° erforderte.

Der benutzte Messkopf wird nahe einer Unebenheit des rotierenden Teiles angeordnet, ohne sie zu berühren. Der Teil kann mit der zu messenden Geschwindigkeit einer mehrfachen oder ganzzahligen Teilgeschwindigkeit laufen. Die Unebenheit kann aus einer Keilnute, Nocke, dem Zahn eines Zahnrades, einer grossen Sechskantenmutter, einem Loch oder Höcker über 1,6 mmϕ und Tiefe bestehen.

Die Kapazität ändert sich jedesmal, wenn die hervorstehende oder eingelassene Unebenheit am Messkopf vorbeigeht. Dadurch wird ein Spannungsimpuls hervorgerufen, der über ein Kabel kleinen Durchmessers einen Zähler speist, der mehrere hundert Meter vom Messkopf entfernt aufgestellt sein kann.

Das Tachometer besteht aus einem elektronischen Zähler bekannter Arbeitsweise. Ein direkt in U/min geeichtes Drehspulinstrument dient als Anzeiger. Sechs Messbereiche ermöglichen Messungen unter 50 U/min ... 200,000 U/min, wobei in gewissen Fällen Korrekturfaktoren benutzt werden. Die Messgenauigkeit ist immer besser als 2% des Vollausschlages.

Der kapazitätsempfindliche Messkopf wird nicht durch übermässige Temperaturen, Schwingungen oder Öl beeinflusst.

EE 15 759 *für weitere Einzelheiten*

STABILISIERTES ZWILLINGS-NIEDERSPANNUNGSNETZGERÄT

(Abbildung Seite 700)

Advance Components Ltd, Roebuck Road, Hainault, Ilford, Essex

Modell P.P.3 besteht aus zwei transistorisierten Netzteilen mit gemeinsamem Netztransformator in einem Gehäuse. Sie geben je 0 ... 30 V 1 A stabilisierten Gleichstrom ab, und da die beiden Teile nicht geerdet sind, können sie entweder getrennt arbeiten oder seriengeschaltet werden, um entweder eine Höchstleistung von 60 V 1 A oder gleichzeitig positive und negative Spannungen (max. je 30 V) abzugeben.

Der Spannungsbereich jedes Teiles von 0 ... 30 V ist in drei überschneidende 10 V-Bereiche unterteilt, deren variable Einstellung direkt in Volt geeicht ist. Eingebaute Volt- und Amperemeter können unabhängig zu jedem Netzteil geschaltet werden. Das Amperemeter hat zwei Bereiche.

Vier Abschlussnetzwerke für den Ausgang ermöglichen den praktischen Ausgleich des Widerstandes der Verbindungsleitungen zur Belastung.

Überlastungsschutz durch eine elektronische Schaltung gibt vollständigen Schutz sowohl gegen anwachsende, als auch Kurzschlussbelastung. Der Kreis kann durch Bedienung des „Ein-Aus"-Schalters auf der Frontplatte wieder hergestellt werden. Schutz gegen Störungen im Gerät und übermässige Netzstromstösse erfolgt durch eine auf der Fronttafel angebrachte Sicherung im Netzstromkreis.

EE 15 760 *für weitere Einzelheiten*

IMPULS-MONITOROSZILLOGRAF

(Abbildung Seite 700)

Mullard Equipment Ltd, Mullard House, Torrington Place, London W.C.1

Der von Mullard Equipment Ltd. eingeführte neue und preisgünstige Impuls - Monitoroszillograf hat eine schnelle Anstiegszeit von 2 ns, die einer Bandbreite von 0-220 MHz entspricht.

Modell L 362 wurde für Darstellung und Messen von Impulsen mit 3 ns bis 3 µs Dauer entwickelt und ist daher für Arbeiten auf dem Kerngebiet und an Rechnern, sowie für Untersuchung von Radarimpulsen für Nahverkehrsanlagen hoher Auflösung besonders wertvoll.

Modell L 362 hat ausser der grossen Bandbreite noch verschiedene, äusserst bemerkenswerte Eigenschaften. Eine Empfindlichkeit von 150 mV/cm eignet sich in Kombination mit den gleichstromgekoppelten Verstärkern besonders für die Untersuchung von Transistor-Schaltungen. Der Eingangs-Messkopf ist kompakt, leicht zu handhaben und belastet die zu untersuchende Schaltung nur mit 2 pF. Die z.B. in der Kernphysik auftretenden identischen Impulse mit willkürlichem Abstand können genau so gut dargestellt werden wie kurze Impulse geringer Frequenz, die besonders hell erscheinen. Weitgehende Transistor-Bestückung macht das Gerät kompakt und tragbar.

Modell L 362 entstand als Weiterentwicklung einer Konstruktion des Kernenergie-Forschungsinstitutes in Harwell und arbeitet nach dem Probenprinzip: die Elektronenstrahlspur entsteht als NF Gegenbild der Eingabe aus einer Reihe von Punkten, wobei jeder Punkt aufeinanderfolgenden Impulsen als Probe entnommen wird. Eine Speicherschaltung hält den Punkt fest, bis die folgende Probe entnommen ist. Dadurch ist der Elektronenstrahl in den meisten Fällen für über 80% der Periodenzeit „an", und es entsteht eine Spur grosser Helligkeit ohne Nachablenkungsbeschleunigung.

Betrieblich sind die Vergleichsschaltung, sowie die X und Y Speicherkreise das Herz des Gerätes. Der Ausgang der X Speicherschaltung wird mit einer Ablenkwellenform des schnellen Kippgenerators verglichen. Wenn die beiden übereinstimmen, gibt die Vergleichsschaltung einen Impuls ab, der das Nehmen einer Probe triggert. Gleichzeitig wird dem Speicherkreis ein Impuls zugeführt, der die gespeicherte Ladung reduziert. Der Triggerimpuls für die Probe tritt daher verhältnismässig früher auf, und der Kathodenstrahlpunkt bewegt sich für aufeinanderfolgende Wiederholungen nach links. Sowie die gespeicherte Ladung unter einen vorgewählten Pegel fällt, wird die Schaltung gelöscht und der Rücklauf beginnt.

Die zu untersuchende Eingabe wird einer Probenschaltung zugeführt, die in einen kleinen, getrennten Messkopf eingebaut ist. Die Eingangsspannung ladet den Eingangskondensator des Gerätes auf, und der Triggerimpuls für die Probe löst die Zuführung der Ladung über einen Verstärker zum Y Ladungsspeicher aus. Die gespeicherte Ladung ist daher proportional zur Eingangsspannung zur Zeit der Probennahme.

Ein zweiter Impuls wird vom X Kreis dem Y Ladungsspeicher zugeführt und dadurch eine der gespeicherten Ladung entsprechende Spannung über den Y Ablenkverstärker an die Y Platten der Bildröhre angelegt. Die X Ladungsspeicherschaltung gibt auch Impulse ab, die den Bildschirm verdunkeln, während sich der Punkt bewegt.

Eine zweite Triggerimpulseingabe kann der Triggerverzögerung und dann dem schnellen Kippgenerator zugeführt werden, der die Ablenkwellenform an die Vergleichsschaltung abgibt, muss aber der Signaleingabe um 0,05 bis 2 µs vorausgehen.

EE 15 761 *für weitere Einzelheiten*

NETZGERÄT FÜR BILDWANDLER

(Abbildung Seite 701)

Sionic Ltd, Cromwell House, 93 Wood End Green Road, Hayes, Middlesex

Das stabilisierte Netzgerät Type 359A wurde besonders für den Betrieb der Mullard Bildwandlerröhre entwickelt.

Die stromstabilisierte Versorgung der Fokussierungsspule liefert 200 ... 500 mA für eine Nennbelastung von 700Ω. Eine Belastungsänderung von ±100Ω ruft eine Stromabweichung von 1% hervor.

Der Hochspannungsteil liefert 5 ... 14,5 kV bei 1 µA mit einer 1 µA Anzapfung für das Gitter. Ohne Gitteranzapfung stehen 100 µA zur Verfügung. Der doppelte Spitzenwert der Restwelligkeit ist geringer als 1 V und die Gesamtstabilisierung besser als 2%.

EE 15 762 *für weitere Einzelheiten*

MINIATURSTECKER UND -SOCKEL

(Abbildung Seite 701)

Harwin Engineers Ltd, Rodney Road, Portsmouth, Hampshire

Der als Harwin W3000/3001 bezeichnete Kleinststecker und -sockel aus Nylon kann mit 4,76 mm Mittellinienabstand montiert werden. Die vergoldeten Kontakte sind für 7,5A bei 750 V berechnet, der Kontaktwiderstand ist durchschnittlich 2 mΩ und die Isolation besser als $10^{6}\Omega$. Arbeitstemperaturen bis zu 100° sind zulässig, und der Widerstand gegen Stoss und Schwingungen soll ausgezeichnet sein.

EE 15 763 *für weitere Einzelheiten*

EINREIHEN-ZIFFERNSKALEN

(Abbildung Seite 701)

Amphenol (Great Britain) Ltd, Ormond House, 26/27 Boswell Street, London, W.C.1

Es ist bekannt, dass das zwangsläufig schnelle Ablesen und Einstellen von Skalen der Grund für Ablesefehler ist, und es wurde festgestellt, dass die einreihige Zifferndarstellung die grösste Genauigkeit ergibt unter Bedingungen, in denen schnelles Ablesen und Einstellen erforderlich sind.

Diese Tatsachen führten zur Entwicklung der Kleinskalen der Borg 1300-Reihe. Um durch einreihige Zifferndarstellung den grössten Vorteil zu erreichen, werden die Ziffern in direktem Kontrast zu ihrem Hintergrund angefertigt, d.h. weiss auf schwarz oder schwarz auf weiss. Ein grosses, aus einem Stück bestehendes, gewölbtes und abgedichtetes Schutzglas ermöglicht Weitwinkelbeobachtung. Um Höchstempfindlichkeit und -genauigkeit zu erzielen, ist der Bedienungsgriff direkt auf die einzustellende Spindel montiert, sodass toter Gang, der durch eine Getriebeübersetzung auftreten könnte, vermieden wird.

Modelle können mit „Fingerspitzen"-Bremsen geliefert werden, die die Einstellung bei Stop und Vibration festhalten. Folgende Typen sind lieferbar:

Modell 1307. Dreistellige Skalen, die in zehn Umdrehungen des Bedienungsknopfes von 0-999 zählen. Mit „Fingerspitzen"-Bremse.

Modell 1308 wie oben, aber ohne Bremse.

Modell 1304. Vierstellige Skalen, die bei 100 Umdrehungen von 0-9999 zählen.

Modell 1305. Fünfstellige Skalen, die bis 1000 Umdrehungen von 0-99999 zählen.

EE 15 764 *für weitere Einzelheiten*

UKW-MESSENDER

(Abbildung Seite 701)

Airmec Ltd, High Wycombe, Buckinghamshire

Airmec Ltd hat vor kurzem die Erweiterung des Fertigungsprogramms durch den neuen UKW-Messender Type 204 angekündigt.

Das Gerät bestreicht das Frequenzband von 1...320 MHz in fünf Bereichen. Die Betriebsfrequenz wird auf einer beleuchteten Filmskala angezeigt, die für jeden Bereich 1,22 m lang ist. Die Eichgenauigkeit wird durch einen eingebauten Quarz-Kontrolloszillator gewährleistet.

Die Stabilisierung des Ausgangspegels auf jeden gewünschten Wert von 2 µV... 220 mV ist eine bemerkenswerte Eigenschaft, die das ermüdende Prüfen und Wiedereinstellen des Ausgangs bei Benutzung verschiedener Frequenzen vermeidet.

Ein eingebauter 1 kHz Oszillator kann für Amplituden- und Frequenzmodulation Verwendung finden. Ein eingebautes Messgerät überwacht den Modulationsgrad bis 50% und Frequenzhub bis 80 kHz mit einer Genauigkeit von 5%. Fremdmodulation von 30 Hz ... 15 kHz kann entweder getrennt oder gleichzeitig mit der eingebauten Modulation angelegt werden. Ebenso kann Impulsfremdmodulation mit Taktfrequenzen von 30 Hz ... 50 kHz benutzt werden.

Ein 52Ω Stufendämpfungsglied konstanter Impedanz ist eingebaut, und ein Anpassungskabel für die Umwandlung der Quellenimpedanz zu 75Ω wird mitgeliefert.

EE 15 765 *für weitere Einzelheiten*

Zusammenfassung der wichtigsten Beiträge

Access—ein ruhendes Schaltsystem mit Kaltkathodenröhren von R. W. Brierley

Zusammenfassung des Beitrages auf Seite 646-654

Die neuen, grossen Fortschritte in der Automatisierung der Fertigung haben zu sehr komplexen Steuersystemen geführt. Der grösste Teil der Anlagen benutzt die „Ein-Aus" Methode und verwendet Relais als logische Elemente.

Es wird schwierig, den hohen Betriebsleistungsstandard einer solchen Anlage aufrecht zu erhalten, wenn sie zu komplex wird. Ein zuverlässiger Ersatz für die elektromechanischen Relais wurde daher gesucht, und einige Systeme mit ruhenden Elementen sind bekannt geworden. Ein System, das Kaltkathodenröhren benutzt (ACCESS-Austin Cold-Cathode Electronic switching System), wird beschrieben.

Ein „Vektorgraph" von K. G. Freeman

Zusammenfassung des Beitrages auf Seite 655-658

Der beschriebene Vektorgraph gibt eine vektorielle Darstellung eines Chrominanzsignals des NTSC-Types. In dem hochempfindlichen Gerät sind auch eine Anzahl anderer Eigenschaften, einschliesslich laufende Prüfung der Verstärkung und Quadraturregelung, vorgesehen. Das Gerät ermöglicht schnelle Einschäzung der Phasen- und Amplitudencharakteristiken des Chrominanzsignals, die für den direkten Demodulationsvorgang des Signal-Strahlerzeugungssystems der Farbbildröhre von besonderer Bedeutung sind.

Halbleitende Verbindungen von D. A. Wright

Zusammenfassung des Beitrages auf Seite 659-665

Zuerst werden in diesem Beitrag die wichtigeren Parameter besprochen, die das Verhalten von Halbleitern bestimmen. Diese Überlegungen werden dann zur Beurteilung der verschiedenen chemischen Verbindungen herangezogen, die bisher untersucht wurden; mittlere Parameter werden angegeben. Unter Berücksichtigung dieser Werte wird die Nutzbarkeit der Materialien für verschiedene Anwendungen diskutiert. Für jeden Verbindungstyp werden bestimmte Tendenzen für die Beziehungen zwischen den Halbleiterparametern und den physikalischen Eigenschaften beobachtet. Diese Tendenzen werden beschrieben und ihre Ausdeutung aufgezeigt. Die für thermoelektrische Anwendungen wichtigen Parameter werden zusammen mit dem Einfluss der ersten Diskussion auf die Materialauswahl kurz besprochen.

Stehwellenverhältnis mit automatischer Anzeige von M. Kollanyi und R. M. Verran

Zusammenfassung des Beitrages auf Seite 666-671

Der übliche Stehwellenverhältnis-Anzeiger mit seinen zeitraubenden Einstellungen kann durch einen automatischen Anzeiger ersetzt werden, der mittels eines Mess– oder Ziffernsichtgerätes direktes Ablesen ermöglicht. Die Schaltung kann für jeden Hohlleiter mit Schlitz oder äquivalente Einrichtungen Verwendung finden. Nur eine einzige Bewegung des Messkopfschlittens ist notwendig, um das korrekte Stehwellenverhältnis abzulesen.

Das beschriebene System stützt sich auf einen Maximum-Minimum-Detektor, der die während der Abtastperiode auftretenden Grenzwerte der Spannungen entdeckt und speichert. Das Verhältnis dieser Spannungen kann mittels verschiedener elektronischer Methoden, die auch beschrieben sind, berechnet werden.

Ein Reststehwellenverhältnis unter 1,5 kann mit einem Höchstbereich von 1 : 3 für Stehwellenverhältnis-Messungen erreicht werden. Falls es gewünscht wird, kann dieser Bereich noch erweitert werden; diese Erweiterung hängt jedoch von den angewandten Methoden ab.

Gasgefüllte Röhren als Musikgeneratoren von Alan Douglas

Zusammenfassung des Beitrages auf Seite 672-673

Die offensichtlichen Vorteile gasgefüllter Zweielektrodenröhren für Kippgeneratoren sollten andeuten, dass viele Aufgaben der Erzeugung elektrischer Musik dadurch gelöst werden könnten; keines der käuflichen Geräte benutzt jedoch solche Bauelemente. Einige neue Schaltungen machen trotz der ihnen eigenen physikalischen Begrenzungen von diesen preisgünstigen gasgefüllten Röhren den bestmöglichsten Gebrauch.

Flächentransistoren in einem Dreieckmodulationssystem von B. E. Williams

Zusammenfassung des Beitrages auf Seite 674-680

Ein mit 14kHz Taktfrequenz arbeitendes Dreieckmodulationssystem wird beschrieben, das Einfachsprechverbindungen mit guter Verständlichkeit ermöglicht. Die einzige Handbedienung des Gerätes erfolgt durch den Sprechknopf; alle anderen Einstellungen werden entweder vorher gemacht oder automatisch betätigt. Der Empfänger regeneriert die ankommenden Signale durch Spannungs– und Zeitunterteilung, bevor sie zum Entschlüssler gespeist werden. Das Gerät arbeitet zufriedenstellend über einen sehr weiten Bereich von Umgebungstemperaturen und ist für Einsatz mit ausgeglichenen Fernsprechleitungen oder Funkbrücken geeignet. Eine Stromquelle für 6 . . . . 12V ist erforderlich.

Ein kleines, stabilisiertes Hochspannungs-Netzgerät mit variablem Ausgang von J. D. O'Toole

Zusammenfassung des Beitrages auf Seite 681-683

Ein wachsendes Anwendungsgebiet besteht für stabilisierte Hochspannungs-Netzgeräte, für die keine hohe Leistung erforderlich ist. Durch die scheinbaren Beschränkungen lieferbarer kleiner Röhren wurde der Bedarf für dieses Gebiet nicht genügend befriedigt. Eine Grundschaltung wird beschrieben, die diese scheinbaren Beschränkungen überwindet. Eine typische Schaltung dieser Art wird behandelt und Leistungswerte werden angegeben.

Anordnung für die Abgabe einer genauen Vektorspannung von D. J. Collins und J. E. Smith

Zusammenfassung des Beitrages auf Seite 684-685

Für viele elektronische Anwendungsgebiete werden Spannungen benötigt, deren Phase und Amplitude gegenüber einem Bezugspegel bekannt sind. Ihre Genauigkeit hängt hauptsächlich von der Genauigkeit der Bauelemente und des Phasenschifters ab, wenn man von elektro-mechanischen Vorrichtungen absieht. Der Beitrag beschäftigt sich mit einer möglichen Anordnung, die eine genaue Vektorspannung abgeben kann, solange ihre Amplitude geringer als der Bezugspegel ist und mit ihr in einem Phasenverhältnis von 0° bis 360° steht. Eine derartige Anordnung ist für die Eichung von Phasenschiftern und Phasenmessgeräten sehr wertvoll.

Eine alternative Vergleichsmethode für elektrische Eigenschaften von Epoxy-Giessharzen von D. H. Thompson

Zusammenfassung des Beitrages auf Seite 686-687

Die Beitrag beschreibt ein Verfahren zur Feststellung der Dielektrizitätskonstante, Verluste und des spezifischen Widerstandes von Epoxy-Giessharzen über einen grossen Temperaturbereich und für Frequenzen bis zu 5 MHz.

Gegenüber der üblichen Methode von Hartshorn and Ward muss eine geringere absolute Genauigkeit in Kauf genommen werden; jedoch verdient diese Methode ausgedehntere Anwendung, wo es sich in der Hauptsache um den Vergleich von Giessharzen und der Temperaturbeständigkeit ihrer elektrischen Parameter handelt.

Besonders können absolute Werte nach dieser Methode angegeben werden, vorausgesetzt, dass die Resultate bei Labortemperatur nachgeprüft und normalisiert werden.

Ein transistorisierter Niederspannungsregler von J. H. Deichen

Zusammenfassung des Beitrages auf Seite 688-689

Die zur Speisung von Röhrenheizfäden mit stabilisiertem Gleichstrom zur Verfügung stehenden Schaltungen werden untersucht und ihre Begrenzungen besprochen. Eine einfache Schaltung, die einen Transistor und zwei Zenerdioden verwendet, wird mit Entwurfsunterlagen und Leistung beschrieben.