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Commentary

AT the beginning of last month a one-day colloquium was held at the Institution of Electrical Engineers on the HS 303 (Early Bird) satellite and the Post Office Earth Station at Goonhilly, at which the introductory survey was given by the Assistant Engineer-in-Chief of the Post Office, Mr. H. Stanesby, followed by six lectures describing the Early Bird Communication Satellite System, the alterations that have been made at the Goonhilly station and the performance obtained to date with the system.

Early Bird was launched earlier this year in April and it attracted considerable attention from the general public due mainly to the number of live television broadcasts made across the Atlantic. But Early Bird was designed primarily as a communications satellite and is still in the experimental stage.

Goonhilly was, of course, erected nearly four years ago for the original Telstar and Relay satellites, and considerable alterations were necessary to the aerial system and station equipment in order to cope with Early Bird due to its greater height and different operating frequencies as well as its relatively stationary position some 22 300 miles above the Atlantic.

Goonhilly is one of the three main European ground stations associated with Early Bird, the others being located at Pleumeur Bodou (France) and Raisting (Germany). Under the existing operating conditions each station in rotation is fully operational for a week, followed by a week in the standby condition and a week for maintenance, the week-end being taken over by the lower capacity ground station at Fucino (Italy) thus avoiding the need for standby equipment at the individual stations but necessitating a special broadband network of terrestrial links interconnecting the stations with switching centres at London, Paris and Frankfurt.

In June this year the first public telephone service via Early Bird was set up using 70 of the 240 telephone circuits with which the satellite is equipped, and the stage has now been reached when the results obtained over the last four months are being closely studied.

It seems that the performance of Early Bird and the ground stations has been of a high standard and the next stage, if it is accepted that global communications are to be carried out by a stationary satellite system, is that a minimum of three synchronous satellites, similar to Early Bird will be launched so as to provide a continuous twenty-four hour service.

But this is by no means certain.

On the one hand there are already in existence the several transatlantic telephone cables providing some five hundred telephone circuits for use between Europe and the North American continent together with the round-the-world Commonwealth telephone cable system for intercontinental communication and while the growth rate is expanding it is unlikely that the full potential of a synchronous satellite system will be reached for some time to come.

Cost too is becoming an important consideration and already the number of television relays has been reduced on this score.

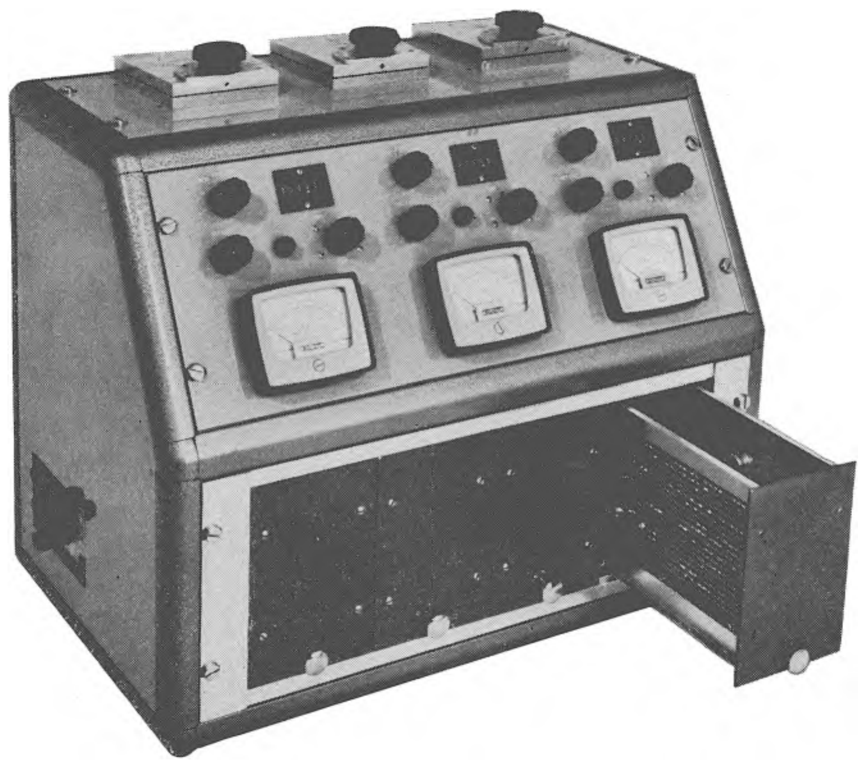
Another feature of the synchronous satellite system is the inherent time delay of about a quarter of a second due to the path length and this seems to be not acceptable as yet for telephone communications.

It is therefore becoming apparent that the type of system and orbit for global communications has not yet been decided and must wait until the Interim Communication-Satellite Committee which is the deciding body, has examined more fully not only the technical but economic aspects.

The Post Office has certainly not made up its mind at this stage to support the adoption of a synchronous satellite system and it is therefore difficult to see what in fact the next step will be.

Three Channel Low Speed Counter for measuring alpha particles

By P. B. Roberts*



A three-channel instrument is described for the measurement of alpha emitting radio-active materials of activities up to a few hundred pico-curies. It is fully transistorized and may be operated from a battery or mains supply. The scaling circuits may have applications where events to be recorded are produced at low rates.

(Voir page 846 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 853)

THE instrument described is used for estimating simultaneously the activities of three samples of alpha emitting radio-active materials. It is fully transistorized, enabling the unit to operate from a battery or from the mains supply and it is portable. It was specifically designed for the measurement of pico-curie activities of alpha-emitting radio-active materials which are encountered when estimating the body radio-activity of persons occupationally exposed to alpha-emitting radio-nuclides by analyses of their excreta and the alpha activities of naturally occurring radio-active materials, for example, in food, water, and samples of soils.

Almost all of the conventionally available 'counting' equipment is capable of receiving a very high rate of incoming pulses. It is therefore unnecessarily complicated when used for the measurement of small quantities of radio-active material. In the instrument described, the rate of recording pulses is limited by a mechanical register to about 10 pulses per second which is ample for activities up to a few hundred pico-curies. Apart from their use in the measurement of alpha-emitting radio-active materials the scaling circuits may have applications where the events to be recorded are produced at low rates.

Description of the Instrument

GENERAL

A diagram showing one of the three detectors is given in Fig. 1. The radio-active material to be measured is plated on to a metal planchette and placed face down on the zinc sulphide coated 'Perspex' disk at A. Light scintillations which are produced when alpha particles are absorbed in the zinc sulphide layer, pass through the Perspex disk and the Perspex window and impinge on the photo-cathode of the photomultiplier tube, subsequently producing voltage pulses suitable for the circuits of the scaling equipment which follows. In operation the photomultiplier is in the light tight enclosure shown in Fig. 1. The hinged cover is secured by a locking mechanism operated by a control knob which also engages with the spindle of a rotary toggle switch in the e.h.t. circuit. The position of the switch is arranged so that the high voltage is applied to the photomultiplier tube only when the cover is fully closed, thus protecting the photomultiplier from exposure to light originating outside the box. Two 'O' rings, located at each end of the assembly, are used to obtain light tight seals. The case and lid may be constructed of any convenient metal but care should be taken to avoid alpha-emitting contamination of parts within alpha-ray range of the zinc sulphide screen. For this reason a disk of stainless steel, which has a low alpha activity, is inserted in the cover and the Perspex window is recessed producing a lip which prevents any alpha particles originating in the walls of the enclosure from reaching the zinc sulphide layer. Perspex also has a low alpha activity. The complete instrument, a photograph of which is shown above comprises three detectors and four modules. Each of the latter contains two circuit

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boards which are inserted into edge connectors fitted at the rear of the instrument case. One module is shown exposed in the photograph. The events are recorded on three mechanical registers which give a measure of the activity of each source. The circuit boards of three of the four modules perform the following functions associated with each of the three detectors:

(a) Amplification.
(b) Pulse-shaping
(c) Driver for mechanical register.
(d) High-voltage generation.
(e) Stabilization for power supplies.

The fourth module contains the low-voltage power supply and the test-oscillator circuits. Photomultiplier tubes type 6097S (EMI) are used with their cathodes connected to ground potential and their dynode chain register networks connected to a high-voltage supply of between 500 and 1 000V.

AMPLIFIER

The circuit for the amplifier is shown in Fig. 2. Pulses from the anode of the photomultiplier are transmitted, by means of capacitor C_2, to the input of the emitter follower, VT_1, which has a high input and low output impedance. The output pulse at the emitter is connected to the base of transistor VT_2 by means of capacitor C_3. This single-stage amplifier provides a gain of 10, resistors R_{19}, R_{20}, R_{21} and R_{22} serving to stabilize the gain and operating point. The positive output pulse is applied through capacitor C_6 to the input of the pulse-shaping circuit which consists of transistors VT_3 and VT_4.

PULSE SHAPER

The threshold level of this circuit (also shown in Fig. 2) is set to a value of one volt which provides sufficient discrimination against noise and unwanted signals. The positive going signal applied to the base of the normally conducting transistor VT_4 by means of capacitor C_8, causes a reduction in the value of the current flowing through resistor R_{31}. The negative pulse produced at the collector of transistor VT_4 is connected to the base of transistor VT_3 by means of resistor R_{28} and capacitor C_9, causing an increase in the value of the current flowing through resistor R_{25} and produces a positive going pulse which will enhance the effect caused by the original input pulse. As the input pulse decays to its initial voltage level, capacitor C_8 charges through resistor R_{29} (in parallel with the input impedance of transistor VT_4) and enables the voltage of the base to recover to its initial value, allowing the collector to return to near-earth potential and causing transistor VT_3 to return to its non-conducting condition. An emitter-follower, transistor VT_5, transfers the negative going output pulse at the collector of transistor VT_4 to the input of the register drive circuit. If the count switch, S_2, is in the off position the output from the emitter-follower stage is connected to earth potential.

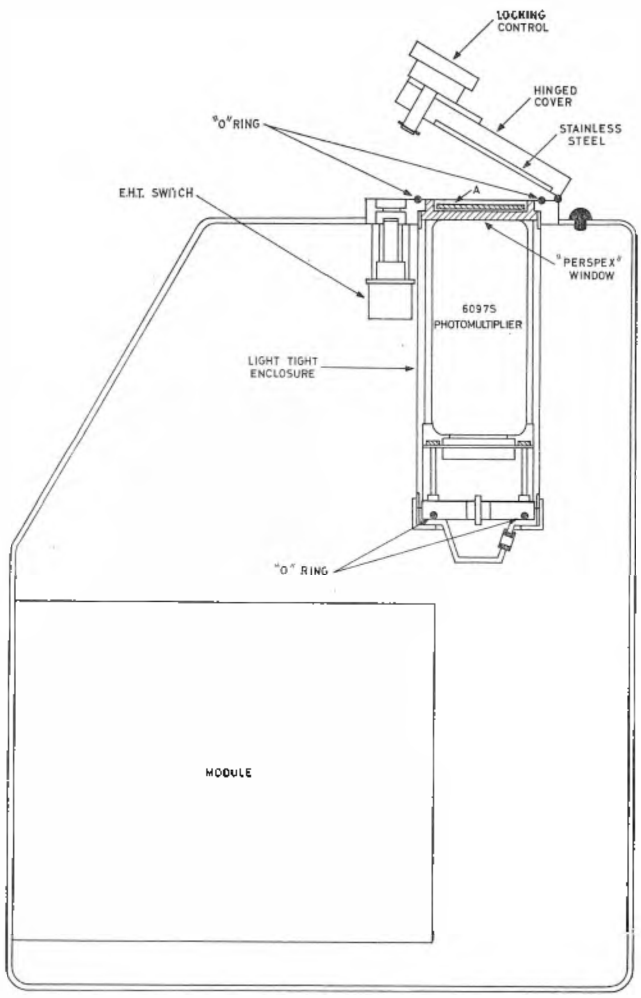
REGISTER DRIVE

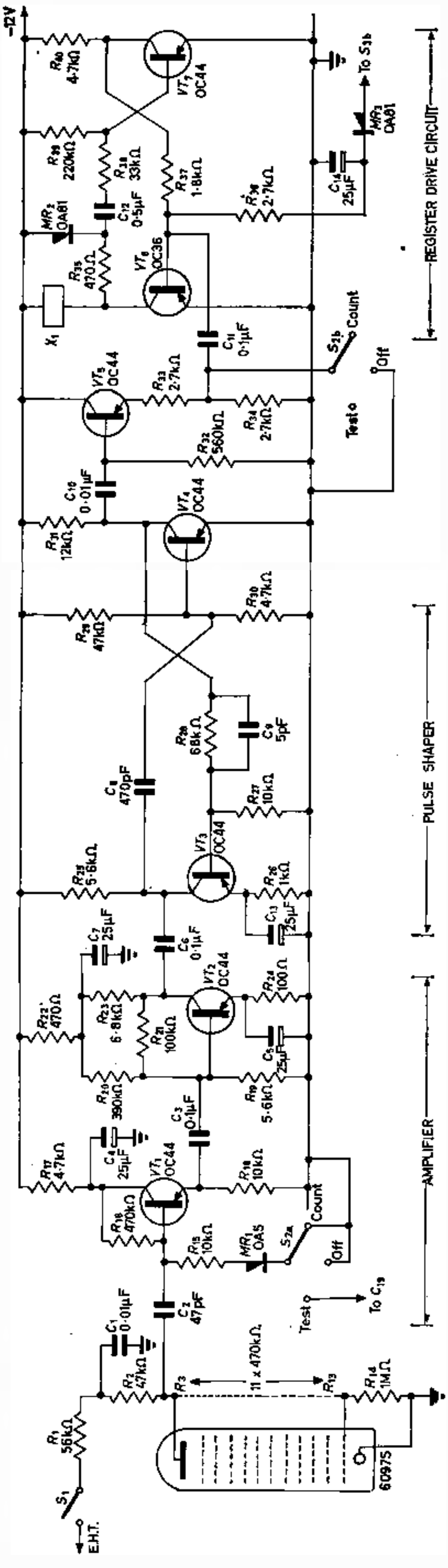
The register drive circuit is shown in Fig. 2. Initially, transistor VT_6 is non-conducting and transistor VT_7 is conducting: therefore, the mechanical register, X_1, is not energized. The negative waveform from the pulse shaper is transmitted, by means of capacitor C_{11}, to the base of transistor VT_6 which is positive with respect to its emitter due to the potential at the junction of resistor R_{36} and rectifier diode MR_3. Transistor VT_6 conducts, causing the register to record a count, and the resultant positive going pulse at the collector is passed to the base of transistor VT_7 by means of resistors R_{35} and R_{38} and capacitor C_{12}. The value of the current flowing through resistor R_{40} is reduced and the negative waveform at the collector of transistor VT_7 is passed by means of resistor R_{37} to the base of transistor VT_6, where it will assist the input pulse to bring about complete conduction. When the input pulse disappears, capacitor C_{12} will charge through resistors R_{38} and R_{39}, allowing the base of transistor VT_7 to return to its initial value and causing transistor VT_6 to become non-conducting. Since the register X_1 presents an inductive load to the collector of transistor VT_6 a large negative overshoot occurs when the current through it is rapidly reduced. To prevent this large overshoot from exceeding the maximum value of base-to-collector voltage, diode MR_2 is used in conjunction with resistor R_{35}, which is selected to give minimum overshoot. The diode will conduct when the collector voltage equals the supply voltage.

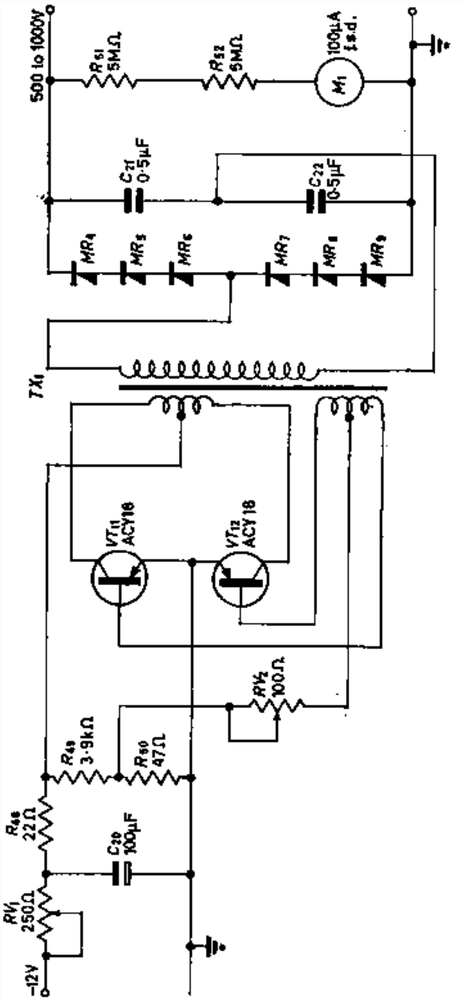
HIGH-VOLTAGE GENERATOR

The high voltage required by the photomultiplier tube is generated by a push-pull d.c. to d.c. convertor (see Fig. 3) in which the primary winding of the transformer T_1 is connected in series with the collectors of the two transistors VT_{11} and VT_{12}. The transistors are used to interrupt the 12V negative supply to give a square wave signal which is stepped up by means of the transformer, the output of which is rectified and doubled to give the desired direct voltage. A feedback winding is connected to the bases of the transistors by means of the preset series

Fig. 1. One of the three detectors







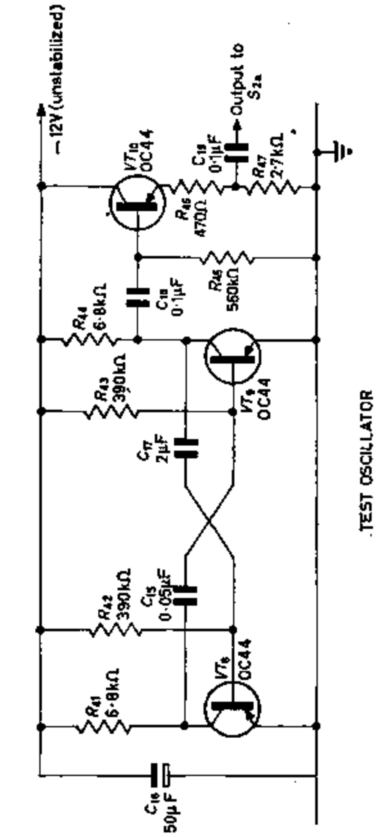


Fig. 2 (above). The amplifier, pulse shaper, register drive and test oscillator circuits

Fig. 3 (right). The d.c. to d.c. convertor

resistor RV_3, which is used to obtain optimum working current for different transistors of the same type. In this type of circuit an efficiency of 70 to 80 per cent is obtained, the losses occurring in the transformer, transistors and the feedback winding resistor RV_2. The variable resistor RV_1 varies the input voltage which controls the output voltage in the way shown in the graph of Fig. 4. The output voltage is monitored by a 100μA meter M_1, calibrated in kilovolts, connected in series with two fixed 5MΩ resistors, R_{51} and R_{52}, across the output of the convertor. Each convertor is capable of delivering a current of 200μA, at a voltage of 1kV, this being the amount required by the dynode network.

Stabilized Power Supplies

Each channel has its own stabilizing circuits (see Fig. 5) the inputs of which are taken from the common power unit described below. The series regulation principle is used with a differential voltage amplifier, transistors VT_{13} and VT_{14} and series regulator transistors VT_{15} and VT_{16}. The voltage reference, formed by the Zener diode MRZ_1 is applied to the base of transistor VT_{13} and is matched by a similar voltage applied to the base of transistor VT_{14} by the variable resistor RV_3. The error signal output from the collector of transistor VT_{14} is applied to the two emitter-followers, transistors VT_{15} and VT_{16}, thus producing a change in the output voltage, the sense of which will be opposite to the change in the voltage on the base of transistor VT_{14}. The regulators are capable of delivering a current of 50mA at a fixed output voltage of 12V with a stability of 2 per cent against $\pm$10 per cent variation in the mains supply.

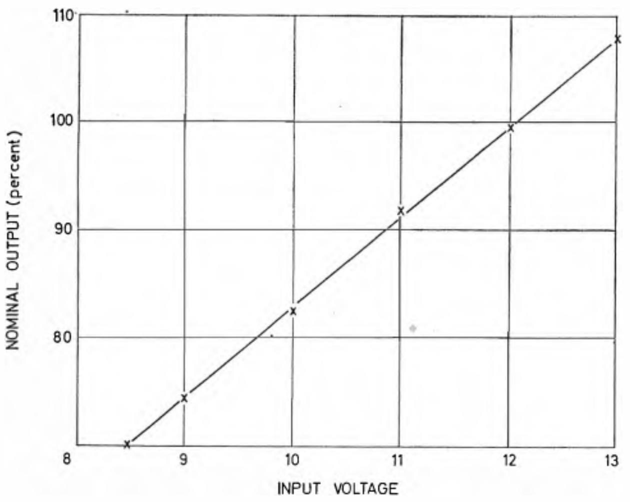


Fig. 4. Variation of output voltage with input voltage

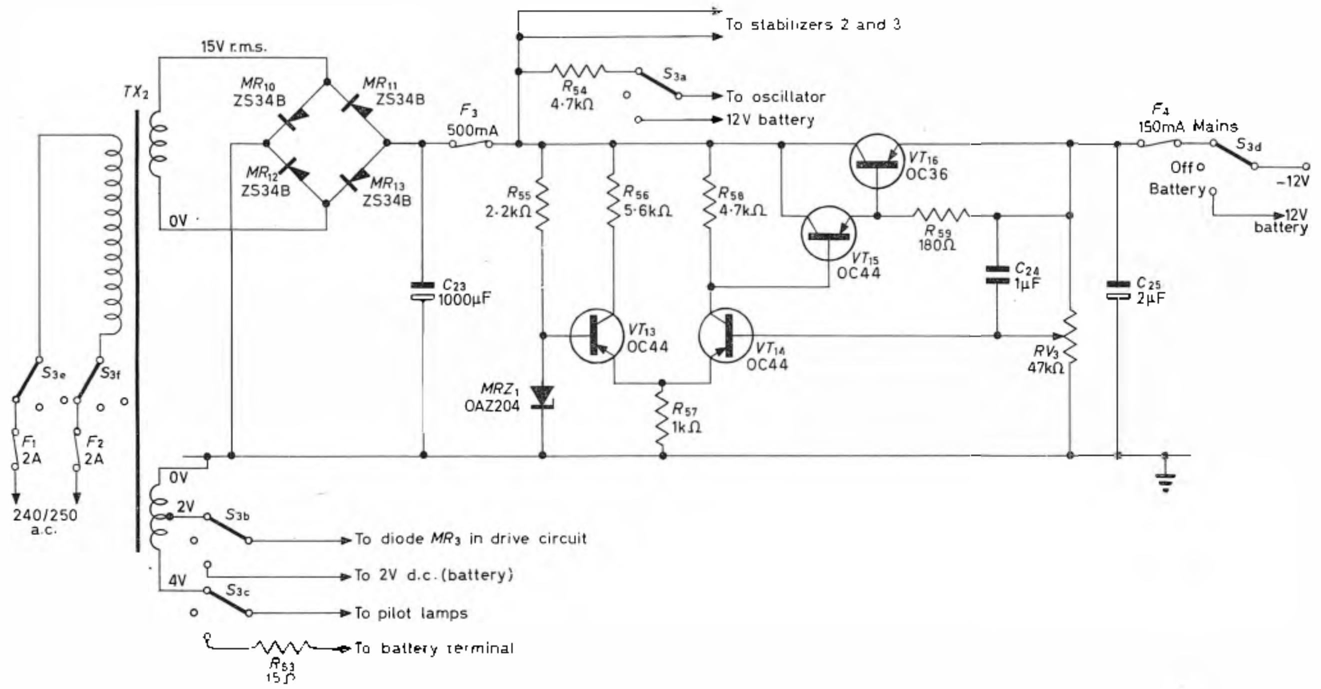


Fig. 5. Power supplies

Power Supply

This is a conventional bridge rectifier (see Fig. 5) the output of which is connected through a fuse link F_3 to the series regulator. The 4V a.c. winding on the secondary of the mains transformer provides a supply for the pilot lamps and a supply for the 2V positive required by the mechanical register drive circuit. Switch S_3 provides the means of selecting mains or battery supply, the total power consumption being 20W.

Test Oscillator

In the test oscillator circuit shown in Fig. 2 the frequency of the multivibrator circuit providing the internal test signal is about 3c/s and is set by the C_{15} R_{43} combination. The output waveform is passed to an emitter-follower, transistor VT_{10}, which provides three outputs, one for each channel. The internal test condition is selected by the count switch S_2.

Performance

The instrument is used for measuring very small activities of alpha-emitting materials and it is important, therefore, that the background response is also small. As mentioned earlier, the Perspex disk and window and the stainless steel insert in the cover lid have very low alpha activities. The zinc sulphide screen is prepared by moistening the adhesive on Sellotape with amyl acetate and transferring the adhesive to the Perspex disk. After the solvent has evaporated, zinc sulphide phosphor is sprinkled on the adhesive and any surplus material removed by tapping, producing a very thin layer of the phosphor.

The Perspex disk and zinc sulphide layer are therefore simple to make at small cost and may be renewed whenever it is suspected that they have become contaminated with any alpha-emitting material. For a freshly prepared screen the background response of the instrument is about three pulses per hour. This corresponds to an alpha activity of a few hundredths of a pico-curie. However, the emission of alpha particles is a random process and when measuring activities as low as a few hundredths of a pico-curie it is necessary to record pulses over periods of about one day in order to obtain a reasonable statistical accuracy. It is important, therefore, that the electronic circuits should function reliably over very long periods of time. The instrument described has proved very stable over observation periods of a few days and has so far been in continuous use for more than six months without need of maintenance.

Acknowledgments

The author wishes to thank D. Stubberfield for constructing and testing the electronic circuits, and L. Smith for his work on the mechanical construction of the photomultiplier tube assemblies.

Voltage Stabilization Using Cold Cathode Trigger Tubes

By E. P. T. Atkinson*, B.Sc., and C. F. Hill†, B.Sc.

Details are given of the use of cold cathode trigger tubes in the stabilization of small rectified power supplies with variable output. Performance and limitations of the circuit, together with factors governing cold cathode circuit design are described. Reference is made to the Z806W with its stable close tolerance ignition characteristics. Finally the graphical determination of circuit values is given for output voltages in the range 200V to 360V.

(Voir page 846 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 853)

COLD cathode trigger tubes with their closely controlled trigger-cathode ignition voltage, lend themselves favourably to the stabilization of small rectified power supplies with variable outputs.

The basic circuit is shown in Fig. 1(a), it is substantially a conventional shunt stabilizer with the trigger tube as the regulating element. A rectified a.c. supply V_{in} is applied to the input, see Fig. 1(b); at t_0 with the tube extinguished the anode voltage V_a follows the positive going input voltage. At t_1 the potential V_c of the reservoir capacitor C_L is reached; the diode MR_3 is then opened and C_L is charged via the resistor R_1. When V_c reaches a pre-determined value $V_{c(max)}$ at t_2, the tube is made to ignite causing the anode voltage V_a to be immediately lowered to the maintaining voltage V_m of the discharge. MR_3 is now blocked off and the output circuit is isolated from the trigger tube circuit until the instant t_1 in the following positive half cycle. From the time t_2, C_L commences to discharge via the load resistor R_L; the tube is passing current via R_1 until t_3. At this point V_{in} drops below the tube maintaining voltage, causing the tube to be extinguished. The action is repeated in the next positive half cycle.

Turning to the trigger, it is this which pre-determines the level of V_c at which the tube ignites. During the interval to $t_0 - t_2$, the trigger voltage V_{tr} will follow V_a by means of the attenuator R_2, R_3 viz. $V_{tr} = R_3/(R_2 + R_3) \cdot V_a$. It is when $V_{tr} = V_{tr(ign)}$ that the tube will ignite, when this occurs V_{tr} will be lowered to its maintaining voltage being approximately V_m. With the tube extinguished at t_3, the trigger will again follow the anode in the manner $V_{tr} = R_3/(R_2 + R_3) \cdot V_a$.

Points arising from this mode of operation include the fact that the repetitive stabilizing action relies on the cyclic extinction of the tube, hence an unsmoothed supply must be used; Fig. 1(b) depicts full wave rectification, but half-wave could equally be used. The output voltage can easily be pre-set by inserting a potentiometer in the attenuator chain; the circuit stabilizes not the mean but the peak output voltage and this is dependent only on the ratio of R_2, R_3 and $V_{tr(ign)}$. The ripple on the output voltage is dependent on C_L and R_L and the regulation will depend on these and only to a lesser extent on the input voltage.

Performance and Limitations

The performance of the circuit is determined by the permitted ranges of load, input supply voltage, and output voltage. The factors governing the limits to the performance are threefold:

(1) The point at which the tube no longer provides stabilizing action.

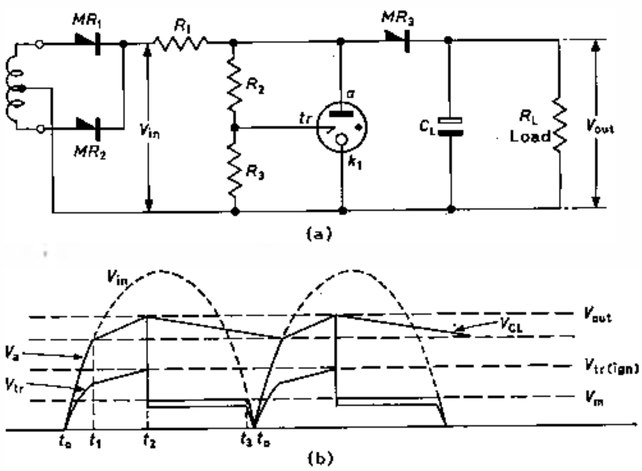


Fig. 1 (a). Basic stabilizing circuit
(b). Waveforms of basic circuit

(2) The point at which the tube parameters are taken outside their published operating limits.

(3) The point at which the regulation of the circuit is no longer acceptable.

Maximum Load

For a given set of circuit conditions, if the load current is increased then C_L will be charged at a slower rate during the period $t_1 \rightarrow t_2$, and discharged more rapidly during period $t_2 \rightarrow t_3$. The charging time is therefore increased, this is accompanied by a corresponding decrease in the conduction time $(t_2 - t_3)$ of the tube. The limit is given by one of two situations, the first being the mean current through the tube during the conducting period falling below the published minimum; if this situation does not arise then the limit will occur when the recharging time has increased to such an extent that the required $V_{c(max)}$ is not reached, that is the tube is not ignited and there is no stabilizing action. The adverse input voltage condition is the minimum value. Circuit components influencing the recharging time and hence the maximum load are the charging resistor R_1 and the reservoir capacitor C_L.

As the load current is increased, so the ripple will increase, and mean output voltage will decrease; the output voltage will still be stabilized at its peak value.

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† *Mullard Ltd.*

Minimum Load

As the load current is decreased, C_L will be charged up more rapidly causing the tube to fire earlier, in this way stabilization is brought about by an increase in the mean current shunted through the tube. In the limit the minimum load current will be given by the mean current through the tube reaching its maximum value; if half-wave rectification is used there is a possibility that the maximum peak current rating will be reached before the maximum mean current. In determining the minimum load current the adverse input voltage condition will be the maximum value.

Input Voltage

Variations in the supply voltage will not affect the maximum peak voltage at which the circuit stabilizes, this being determined by the $V_{tr(ign)}$ of the tube and the trigger attenuator circuit R_2, R_3 of Fig. 1(a). When the trigger reaches its ignition voltage, the anode voltage must lie within a range, the maximum value of which must not exceed the anode ignition voltage of the tube; the minimum value is given by the minimum anode voltage at which the tube will successfully ignite when a trigger discharge is present. The regulation of the supply is dependent on the ripple in the output; this in turn depends only to a small extent on the input voltage, as the latter increases so does the period during which the reservoir capacitor is discharged.

Output Voltage

As mentioned before this is determined by the $V_{tr(ign)}$ of the tube, and the trigger attentuator circuit R_2, R_3; a potentiometer forming part of the latter can be used to pre-set the output voltage. The maximum output voltage will be limited by the requirement that the anode ignition voltage (anode hold-off) must not be attained when the trigger reaches its ignition voltage; the adverse input voltage for this limit will be the maximum value.

The minimum output voltage will be limited by the requirement that the anode voltage will have attained a sufficiently high value to ensure transfer, that is to ensure anode ignition when the trigger has reached its ignition voltage; the adverse input voltage for this limit will be the minimum value.

Circuit Design

As a means to describing the circuit design approach, reference will be made to the Z806W—a tube most suited to this application. The Z806W is a newly developed trigger tube, its anode working range is 190V to 390V d.c. In addition the stability of its trigger ignition voltage is ± 1 per cent over life, depression of the trigger voltage due to thermal effect of the cathode (hysteresis) has been kept to a minimum.

To achieve the large voltage operating range, a screen anode is employed between main anode and trigger. This screen anode is usually connected to the main anode supply through a potential divider network. In normal operation the triggering action initiates first a screen anode to cathode discharge which immediately initiates the main discharge. Limitations on the working range of the tube are mainly due to the operational limits of the screen anode. Fig. 2(a) will be referred to in discussing detailed circuit design.

Anode Circuit

The full range of anode working d.c. voltage is from 190V to 390V, with an added limitation that the anode to screen anode voltage must not exceed 160V; the latter point is provided for by correct design of the screen anode potential divider network.

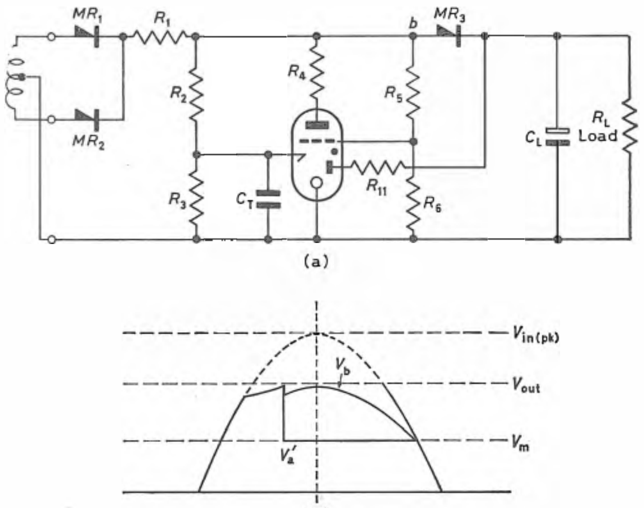


Fig. 2 (a). Detailed circuit using Z806W
(b). Waveforms of Z806W circuit

In the basic circuit of Fig. 1(a), by keeping R_1 to a minimum the efficiency of the circuit is increased due to the reduction of power and voltage loss in this dropping resistor; in addition a fast charging of the reservoir capacitor is attained. However, a lower limit is set to R_1 by the maximum permissible anode current that the trigger tube may pass.

Control of the tube current need not, however, be achieved solely with R_1; in the circuit of Fig. 2(a) it is shared between R_1 and the resistor R_4, the latter being connected between the anode of diode MR_3 and the anode of the trigger tube; R_4 does not limit the charging current of C_2.

In turn, there is a limit to R_4; if it is too large the voltage at its upper end (point b) may exceed the required output voltage for some time during the conducting period of the tube. This might raise the output voltage above its stabilized value; the waveform at point b is given in Fig. 2(b). The maximum voltage at point b is given by $V_m + R_4/(R_4 + R_1)(V_{in(pk)} - V_m)$. Assuming that there is little ripple on the output voltage, then the value of R_4 is given by the requirement that MR_3 must be non-conducting during the period in which the trigger tube conducts, viz.

$$V_m + \frac{R_4}{R_1 + R_4}(V_{in(pk)} - V_m) \leqslant V_{out}$$

In the limit for a given V_{in}, V_{out}, V_m and R_1

$$R_4 = \frac{V_{out} - V_m}{V_{in} - V_{out}} \cdot R_1$$

R_1 is a more fundamental variable in the circuit, and in conjunction with the input and output voltages and the trigger tube maintaining voltage it determines the operating range of current in the circuit. The relationship between them is given later on by means of reference graphs. Once the best values of R_1 and V_{in} for a given V_{out} have been chosen by means of these graphs, then R_4 is determined as above.

Screen Anode

In normal use the screen is connected to the anode supply through a potential divider network, R_5, R_6. The maximum working voltage of the tube is given by the voltage at the anode when the screen anode-cathode breakdown occurs; the minimum working voltage of the tube is given by the voltage at the anode when the

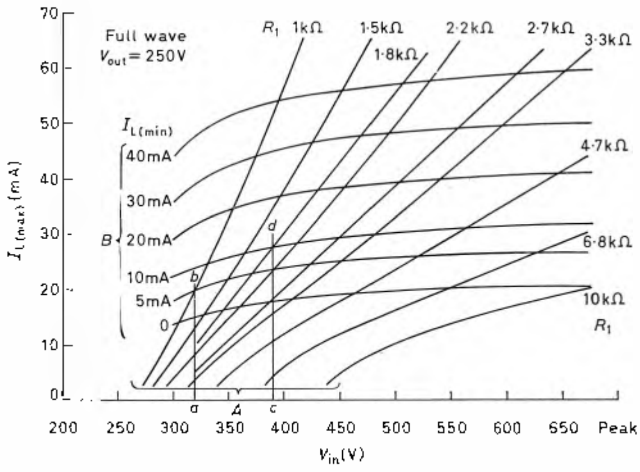


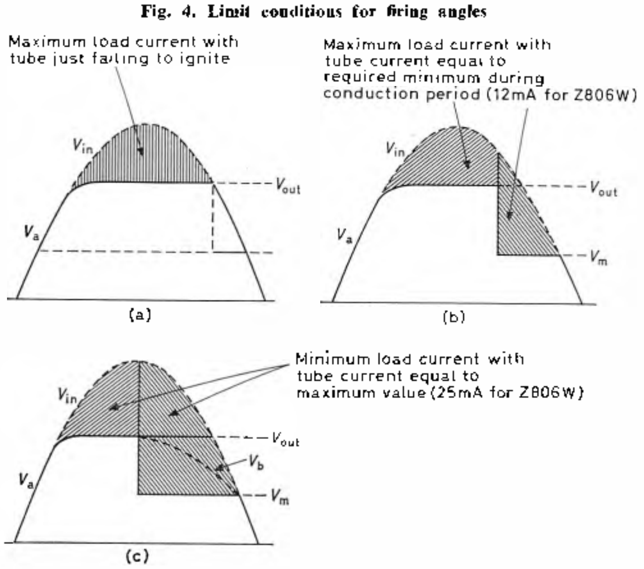
Fig. 3. Design curves for $V_{out} = 250V$

trigger-cathode breakdown fails to cause transfer to the screen anode-cathode gap. The operating range of the tube, in fact, becomes equal to the screen anode operating range multiplied by the inverse of the attenuation ratio, $(R_5 + R_6)/R_6$ and the tube operating range can be varied simply by changing the value of the attenuator feeding the screen anode.

The extreme upper limit to which the working point may be extended in this way occurs when the voltage between screen anode and anode becomes sufficiently large for breakdown between them. This gives a maximum approximately equal to the sum of the screen anode-cathode hold off and the anode-screen hold off.

The extreme lower limit to which the working point may be extended is given by the minimum anode voltage which is required for transfer from the screen anode cathode discharge to the anode-cathode discharge. This is lower than the screen anode transfer voltage, so in order to utilize it the screen anode would have to be connected to a voltage higher than the anode—no attempt has been made to utilize this situation in this article. In the circuit of Fig. 2(a), the attentuator R_5, R_6 must provide an effective impedance of approximately 120kΩ to the screen anode. The calculation of the attenuation ratio R_5/R_6 is simplified by the fact that the

Fig. 4. Limit conditions for firing angles



peak voltage at point *b* just before the tube is required to ignite is stabilized, and is equal to the output voltage. The screen anode voltage should lie between 190 and 250V when the tube is required to be ignited. For output voltages between 250 and 390V suggested values are $R_5 = 180\text{k}\Omega \pm 5$ per cent and $R_6 = 390\text{k}\Omega \pm 5$ per cent.

Trigger Circuit

In order to ensure reliable transfer, triggering by a capacitive discharge is recommended. However, the resultant time delay, due to the charging of the trigger capacitor, should be made as small as possible. For the Z806W it is suitable to have a trigger capacitor of 500pF together with an effective impedance of 100kΩ for the attenuator R_3, R_8. In using 500pF it is necessary that the screen anode is not allowed to go within 20V approximately of its minimum value; if this condition cannot be met, then a 1 000pF should be used with a trigger source impedance of 50kΩ.

Primer Circuit

Best results are obtained with the primer burning con-

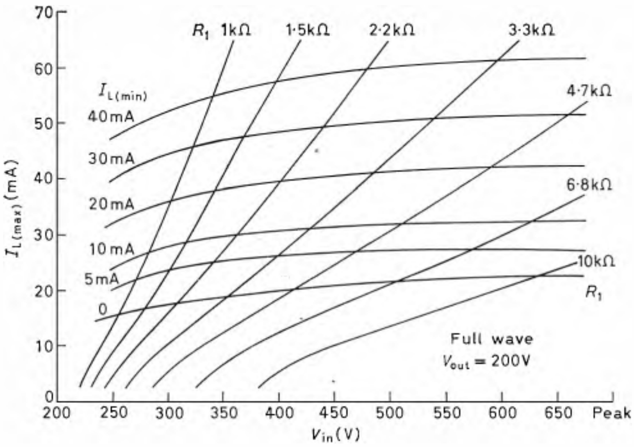


Fig. 5. Design curves for $V_{out} = 200V$

tinuously, hence the primer supply is taken from the output terminal via a 10MΩ resistor.

Graphical Presentation

In the discussion on design of the anode circuit, it was mentioned that the operating range of the load current $(I_{L(max)} - I_{L(min)})$ is dependent on R_1, the peak input voltage V_{in} and the output voltage V_o. The interdependency of these factors does not allow a single graphical presentation to be feasible. The prime requirement of a stabilized power supply is usually its output voltage; therefore the approach taken is to present a series of graphs, each for a particular output voltage. The relationship between the remaining parameters may be taken from the relevant graph. An example of such a graph is given in Fig. 3, this has been drawn up for an output voltage of 250V using full wave rectification in the circuit of Fig. 2(a). The curves *A* give the maximum values of load current that will be stabilized, against the peak value of the input voltage, values of R_1 give the parameter. The curves *B* have been obtained by plotting the minimum load current values along the R_1 curves against the same abscissa, viz. V_{in}. The minimum load current involves R_4 as well as R_1, this is because the maximum tube current dictates the value of the minimum load current. To avoid the complication of a further parameter in the reference graphs, it is assumed

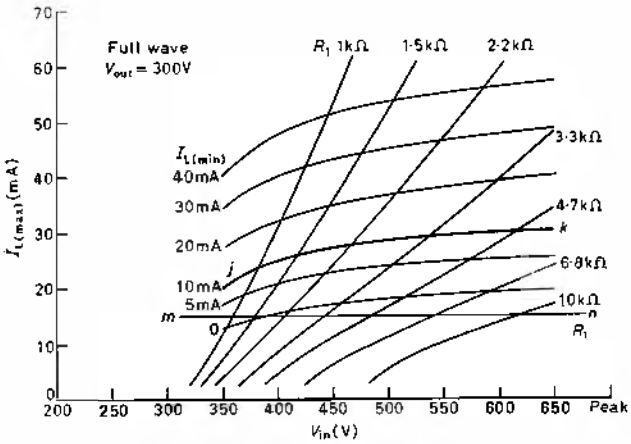


Fig. 6. Design curves for V_{out} = 300V

that the optimum R_4 value is always used with the chosen R_1.

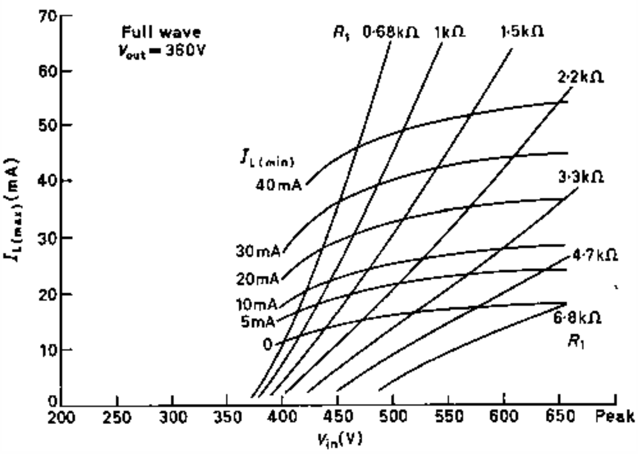
$$\text{i.e. } R_4 = R_1 \frac{V_{out} - V_m}{V_{in} - V_{out}}$$

The graphical presentation has been derived by solving the expression for load currents at striking angles at limit conditions for the tube current; the latter are shown in Fig. 4. In using these graphs it is assumed that C_L is so large that its voltage rise during charging may be neglected. Figs. 5, 6, 7 give working graphs for stabilized output voltages of 200V, 300V, 360V. Working graphs for a more comprehensive range of stabilized output voltage may be obtained from the Mullard Valve Co. Ltd.

To give an example, suppose a 250V stabilized supply is to be made with an input mains supply of 250V $\pm$10 per cent r.m.s. Referring to the 250V graph, Fig. 3, the vertical lines *ab*, *cd* represent the limits of the 250V mains input.

It was stated that the minimum load current was determined under conditions giving the maximum tube current, and the adverse input voltage condition for this was the maximum value. Therefore, in considering the minimum load current the intersection with *cd* is taken; in Fig. 3 it is seen that for an I_L of 5mA then $R_L > 2{\cdot}2\text{k}\Omega$. Likewise in considering the maximum load current, the intersection with *ab* is taken, so that for $I_{L(max)}$ = 8mA then $R_k < 2{\cdot}2\text{k}\Omega$. Hence, with $R_k = 2{\cdot}2\text{k}\Omega$

Fig. 7. Design curves for V_{out} = 360V



the circuit will stabilize between 5mA and 8mA load current.

Still considering the V_{in} limit lines *ab*, *cd*, if it were required to stabilize down to 0mA then a value of $R_1 > 2{\cdot}7\text{k}\Omega$ approximately is required, with $R_1 < 2{\cdot}7\text{k}\Omega$ a value of $I_{L(max)}$ = 5mA is permissible given by the intersection of *ab* and $R_1 = 2{\cdot}7\text{k}\Omega$; hence with $R_L = 2{\cdot}7\text{k}\Omega$ the circuit will stabilize between 0 and 5mA current load. Again with $I_{L(max)}$ = 10mA then $R_L < 1{\cdot}8\text{k}\Omega$ is required, however, for $I_{L(min)}$ = 10mA then $R_L > 1{\cdot}8\text{k}\Omega$, therefore the maximum current at which the tube will stabilize with the given V_{in} and V_{out} requirements is 10mA. The required R_1 in such a situation is 1·8kΩ.

If the limits *ab*, *cd* are brought in, i.e. a closer tolerance input supply, then for a given value of R_1 the value of $I_{L(min)}$ is lowered (given by intersection of R_1 and *cd*) also $I_{L(max)}$ is increased, that is the load current range is extended. It should be noted that nominal values of R_1 have been taken; in practice the tolerance limits should be used on the graphs, the maximum R_1 being

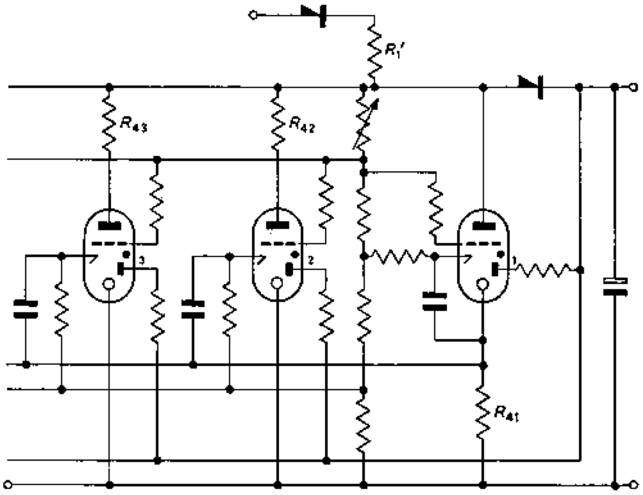


Fig. 8. Circuit for load currents greater than 25mA

used to determine $I_{L(min)}$ and minimum R_1 to determine $I_{L(max)}$.

It may happen that the power supply is specified by the load current range, say 10 to 15mA at 300V output. The current limits—*jk*, *mn* are drawn on the 300V curve, Fig. 6. The intersection between the selected R_1 and the limits of V_{in} must be such that they lie below $I_{L(min)}$ *jk* and above $I_{L(max)}$ *mn*. If a nominal $R_1 = 1\text{k}\Omega$ is taken, then a swing of 360 to 385V for V_{in} is obtained; that is 372V peak $\pm$3 per cent. However, if a nominal $R_1 = 4{\cdot}7\text{k}\Omega$ is taken then a swing of 485 to 620V is obtained, that is 552V peak $\pm$12 per cent; at this higher voltage a greater efficiency is obtained. Again the adverse tolerance condition of R_1 should be used with the graph. When an operating range of V_{in} is used, then R_4 must be calculated with the maximum value of V_{in}, in the last example quoted $R_4 = (300-120)/(620-300) \times 4{\cdot}7\text{k}\Omega = 2{\cdot}7\text{k}\Omega$. Full wave rectification has been considered throughout the article, half-wave rectification may equally well have been used. This would provide a cheaper circuit but care must be taken to ensure that the maximum peak current rating of the tube (200mA for Z806W) is not exceeded. In using half-wave rectification the mean current will be half that of the full wave rectification for the same values of R_1 and R_4; hence the curves of Figs. 3, 5, 6, 7 may be used simply by halving the values of R_1.

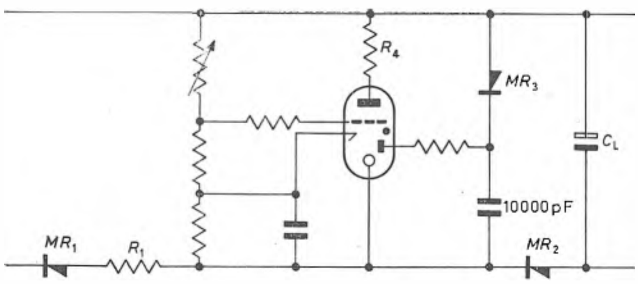


Fig. 9. Circuit for negative stabilized voltage

Modifications

The range over which the available current may vary is rather limited, the theoretical limit being 25mA at a single infinitely high input voltage. It is possible, however, to extend the principle to medium output supplies by using several trigger tubes in parallel. The output current of one trigger tube is simply multiplied by the number of trigger tubes used. Fig. 8 shows the type of circuit envisaged. The limiting resistor R_4 of the first tube V_1 is in its cathode lead instead of the anode and the pulse from this cathode is used to fire the other tubes. In this way one tube only determines the stabilized voltage and has to be set up to regulate the firing point. If n tubes are used then the R_1 value for each tube will be $1/n\ R_1$ as given by Figs. 3, 5, 6, 7, however, the R_4 value will be that derived from the graphs.

Another modification to which the circuit lends itself provides a negative stabilized voltage, as shown in Fig. 9. The diodes and dropping resistor R_1 are transferred to the negative side of the circuit; the primer supply is now given by MR_4, C_4.

Conclusion

The Z806W trigger tube may be used as a very convenient regulating element for small power supplies, as it is easy to set the output to any given level over a fairly wide range, and the output stability is dependent principally on its very stable trigger ignition characteristic. The output current range is determined by the current range of the trigger tube and so is limited to an equivalent amount. The output current range becomes narrower the larger the input voltage variations that have to be tolerated, but as a rough guide 0 to 10mA can be obtained from a supply liable to 10 per cent input variations when the nominal peak input voltage equals approximately twice the required output voltage. The largest ranges are obtained with low output currents and high input voltages.

If the supply voltage is required variable then the output current becomes even further restricted but would be very suitable for fixed or very low outputs such as for a reference voltage. However, the system can easily be adapted to handle higher outputs by the addition of more tubes.

The limits to which any circuit can be extended are plotted in the accompanying graphs which facilitate the otherwise rather complex estimation of performance or component values for required circuits.

Inertial Navigation for Civil Airlines

As part of their evaluation of inertial navigation for long-range navigation, Qantas Airways, Australia's overseas airline, has decided to carry out operational trials with the Elliott E.5 inertial navigation system in one of their Boeing 707's. In support of this programme, the British Ministry of Aviation has purchased two E.5 inertial systems for their own operational trials.

Qantas is the first of several major airlines which will be carrying out their own operational trials and, to meet this world-wide interest, Elliotts are making available a number of E.5's, which will allow airlines to build up a background of operational experience with this new type of navigation aid. Detailed arrangements with several airlines are already in hand.

The E.5 is developed from the E.3, which was flown for almost 3 000 hours in a BOAC Boeing 707 during recent months.

Inertial navigation, which is entirely independent of ground-based and celestial navigation aids, weather and surface conditions, has special attractions for airlines operating over very long stages and round the world. The equipment will operate anywhere in the world, it simplifies the work of the crew and helps the airlines to overcome any shortage of specialist crew members, and it offers a degree of accuracy which may materially reduce traffic separation problems over oceanic areas.

Although inertial navigation has long been a recognized military aid, in aircraft, missiles and ships, its use by civil operators introduces completely new problems and requirements. First, the cost has to be reduced far below the levels formerly accepted. Second, the reliability of the system has to be far higher. A military aircraft may fly only 300 or so hours a year. A missile flies for a matter of minutes. A civil airliner must fly 12 or more hours per day over a wide geographical area. There is no possibility of frequent inspection, and the system must operate for at least a year or for 3 000 or 4 000 flying hours without major maintenance.

The E.5 combines the well proven 'inside-out' gimbal layout of the E.3 with the Autonetics* G.10A gyro which, as a non-floated instrument, represents a new departure in reliability, simplicity and low cost in true inertial-quality gyros.

Instead of running on conventional ball bearings, the G.10A's gyro wheel spins about a spherical ball from which it is separated by a thin film of gas generated by shapes engraved in the wheel itself. There is therefore no wear while the wheel is running. The ball provides for both spin and two-axis displacement and the wheel is thus the only moving part. The overall number of parts is radically reduced and flexible electrical contacts and slip-rings eliminated altogether. In this way, both cost and reliability are very greatly improved.

The gyro platform of the E.5 can be installed on the same mounting attachments presently used for the Doppler aerial in the Boeing 707 and fits the same volume. The Boeing 707 has a built-in bay for two Doppler aerials side-by-side, and single or duplicated E.5 platforms can replace either one or both Doppler aerials without structural modifications.

The E.5 incorporates a special-purpose digital/analogue navigation computer which continuously calculates geographical position in latitude and longitude by digital means. Thereafter it produces steering information and autopilot coupling signals in the analogue form appropriate to indicators and associated equipment conforming to accepted ARINC standards. Alternatively the new Elliott MCS 920M microminiaturized digital general-purpose computer can be substituted where broader functions such as fuel management calculations are to be included.

A single initial switching operation by the pilot puts the E.5 into its alignment mode, during which the platform is warmed-up and aligned by gyrocompassing without further supervision and without the need for any external magnetic reference.

All indications and navigational operations are then carried out from a single small control panel next to the pilot, with nine different forms of output available. Latitude and longitude are continuously shown and four way-points along the route can be set in terms of latitude and longitude. A fifth can be set in terms of distance and bearing to a chosen point, a facility unique to the E.5. The panel then indicates across-track error, groundspeed, time to go to the way-point selected, selected track, instantaneous track, drift angle or track error, according to simple switch settings. The way-points can be inspected or reset without interrupting running computations. All necessary navigation operations are thus available simply and quickly to the pilot at his normal flight station.

A feature of the E.5 is that it continuously monitors its own performance in all important sections and incorporates first-line test facilities and failure-warning as an integral feature. Faults are detected and traced to the appropriate 'unit' for all significant failures. Elliotts are guaranteeing mean time between failures for each unit in the system and are also guaranteeing performance accuracy.

* *A division of North American Aviation Inc.*

A Versatile Transistor Stimulator

By W. J. Bannister* and R. H. Kay*

A versatile transistor stimulator for neurophysiological research is described. It consists of three separate units:—

(a) The Delay Unit, driven from the time-base waveform of the final oscilloscope display, providing a single or a pair of pulses to trigger (b) and (c). These triggers, locked to the time-base, can be independently 'delayed', i.e., set stably to any position along the c.r.t. x-axis. They remain fixed at the same spatial positions along the time-base, irrespective of time-base velocity.

(b) The 'Tetanus' Unit, providing gated trains of pulses to trigger unit (c), whenever energized by a trigger from unit (a). Pulse trains can also be initiated independently of unit (a) by a manual push-button.

(c) The Stimulator Output Unit, providing a rectangular pulse of controlled amplitude, duration and polarity, whenever triggered by units (a) or (b), and also when triggered manually by a push-button. The pulse outputs are normally isolated by a well designed doubly screened transformer, which retains the rectangular pulse shape for pulses up to 10msec duration. A d.c. output is also provided for these and longer pulses.

(Voir page 846 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 853)

THE convenience of triggering stimulators at chosen points along an oscilloscope *X*-sweep, rather than at selected intervals of time, e.g., by some *RC* device, has been described before[1]. If the time-base is triggered by time marking impulses or by submultiples thereof, and the stimulator by the time-base, then time, stimulus monitors and a repeatable biological response all remain steadily superimposed on the display from sweep to sweep, and the interstimulus time delay is as accurate as the time-base velocity, which must in any case be good for useful work to be done. The performance of the stimulator is summarized below. It will be seen to follow rather closely the specification of the valve versatile stimulator[1], but enjoys the advantages of:

(1) Transistor operation.
(2) Pulses in a tetanus train which superimpose from sweep to sweep.
(3) Transformer isolation at all pulse durations from 100μsec to 10msec.

Specifications

(a) The Delay Units

These each provide one trigger pulse per time-base sweep, the spatial position along the *X*-axis being controlled by a stepped 10-position switch, with a smooth control spanning each step. Two delay units are available for each stimulator channel. They can be used to trigger the stimulator to provide:

(1) By direct coupling, a single or a pair of output pulses at any position and separation along the X-axis with either leading, or
(2) a single pulse, directly, plus a train of pulses of selected p.r.f. and train duration, via unit (b) below, the single pulse and train each being independently controlled to any desired position along the *X*-axis, or
(3) a 'tetanus' train alone, of controlled p.r.f. and train duration at any position on the time-base.

(b) The Tetanus Unit

This unit generates a train of trigger pulses of controlled p.r.f. and train duration, whenever triggered by a single pulse from (a). The p.r.f. can be varied between 1/sec and 2 000/sec, and the overall duration of the train from 10msec to 10sec. The pulses superimpose on the display from sweep to sweep, whatever their p.r.f.

(c) The Stimulator Unit

In its commonest form, this unit generates a rectangular wave stimulus for each trigger pulse fed to it from either units (a) or (b). Output pulses are available in three current ranges: 0 to 100μA, 0 to 1mA and 0 to 10mA, each range being divided by a stepped control into ten equal divisions, each division being spanned by a smooth fine control. The pulse duration is controlled in three ranges: 100μsec to 1msec, 1msec to 10msec, 10msec to 100msec, each range being divided by a stepped control into ten equal divisions. A two-position switch determines pulse polarity at the output leads. The stimulating current can be monitored by the potential drop across a resistor in the primary (or, rarely, secondary) of the output transformer.

For some purposes, longer pulses are needed, and the d.c. stimulator unit extends the pulse duration range to 10sec. The output is positive or negative going. A relay closes for the duration of the stimulus to allow e.g., a backing-off potential to be applied at the same time as the stimulus.

Triggered Operation

The complete stimulator has two channels, each capable of the functions described in (a), (1), (2), (3), above. A different pattern of stimuli, pulse length, polarity, tetanus p.r.f. etc., can thus be set up in the different channels, and these, if need be, can be combined with each other on common, or different, stimulating electrodes, with the same, or opposite polarities.

Manual Operation

Push-buttons can release from the stimulator either

(a) A single pulse.
(b) A tetanus train of pulses of controlled p.r.f. and train duration.
(c) Pulses continuing at the selected p.r.f., but the train continuing for as long as the push-button is held down.

* *The University Laboratory of Physiology, Oxford.*

In all cases the output pulse remains controlled in amplitude, duration and polarity.

Circuits

As is so often the case, each functionally different unit of the circuit is conventional, and it is their appropriate interconnexion which combines versatility with handiness. Experience in the testing conditions of an elementary class with earlier simple stimulators, (Bannister & Kay[2], Bannister[3]) has helped in the selection of the most reliable transistor circuit configurations. The delay and pulse length generator is, however, a new arrangement which allows a much wider range in pulse durations to be selected for a given number of components, and enjoys very rapid pulse rise and fall times.

(a) The Delay Unit (Fig. 1)

The operation of this unit has been described by Bannister[4]. Spatial position along the time-base is selected by S_1 and RV_1, for each unit. The whole stimulator has four of these driven in parallel from a triple emitter-follower energized by the time-base waveform of a Tektronix 502 oscilloscope.

(b) The Tetanus Unit (Fig. 2)

The pulse repetition frequency is derived from the multivibrator VT_{13}, VT_{14}. This is rapidly thrown into oscillation by the 'gating' transistor VT_{12}, and the flip-flop VT_9, VT_{10}, which determines the gate length (tetanus train duration). The operation of this unit has also been described[4]. Manual operation is by push-button PB_1 (gated train) or PB_2 (continuous train). Gated trains of pulses superimpose from sweep to sweep for every repetition rate.

(c) The Stimulator Unit (Fig. 3)

Pulse duration is controlled by flip-flop VT_{16}, VT_{17}.

Fig. 1. Delay unit

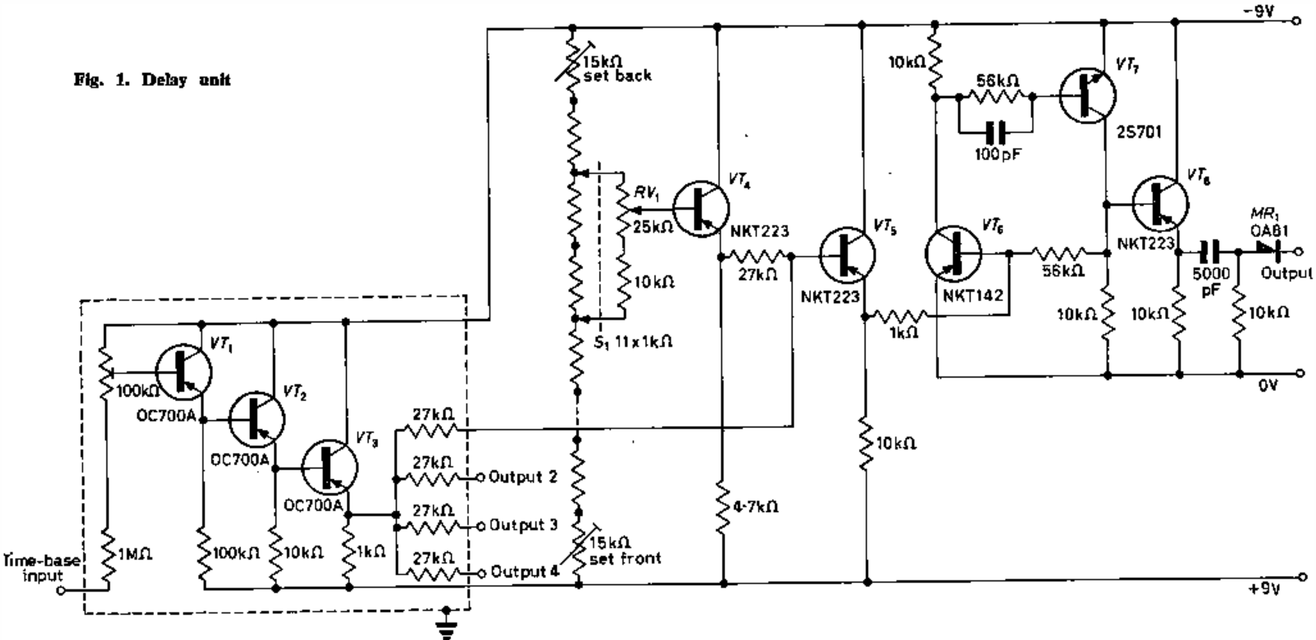
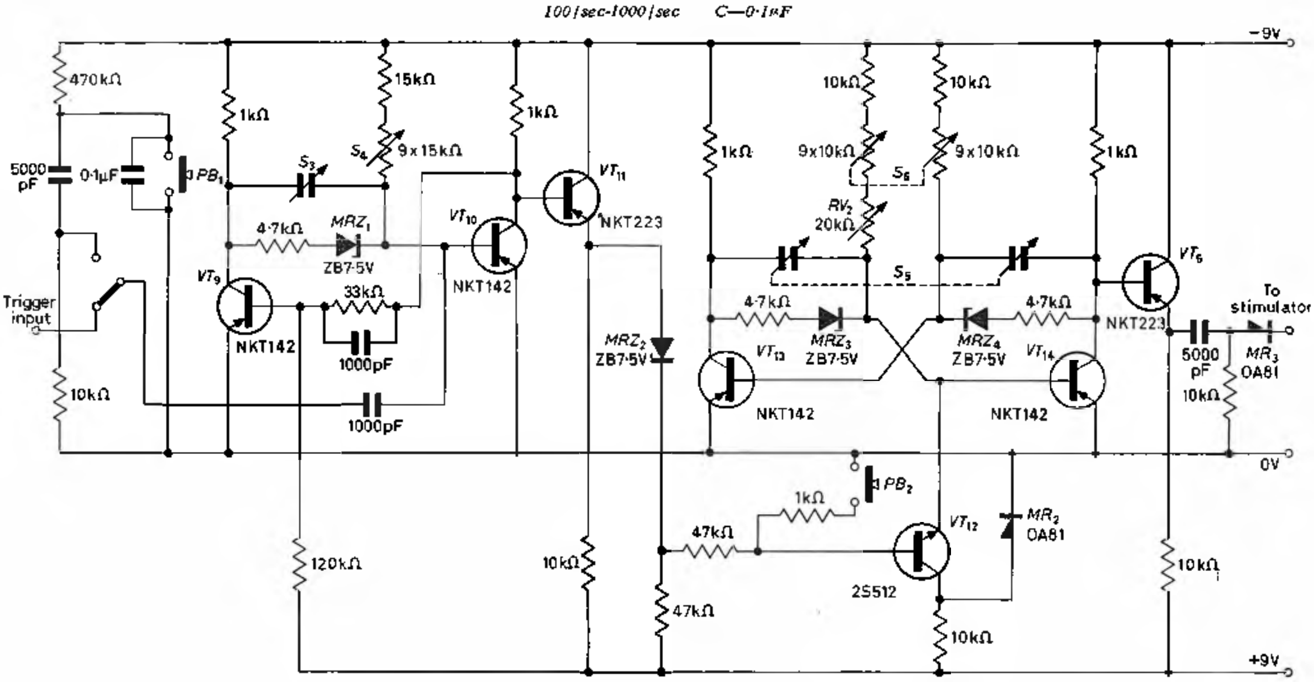


Fig. 2. Tetanus unit

S_4 range 1/sec-10/sec C—10μF
10/sec-100/sec C—1μF
100/sec-1000/sec C—0·1μF



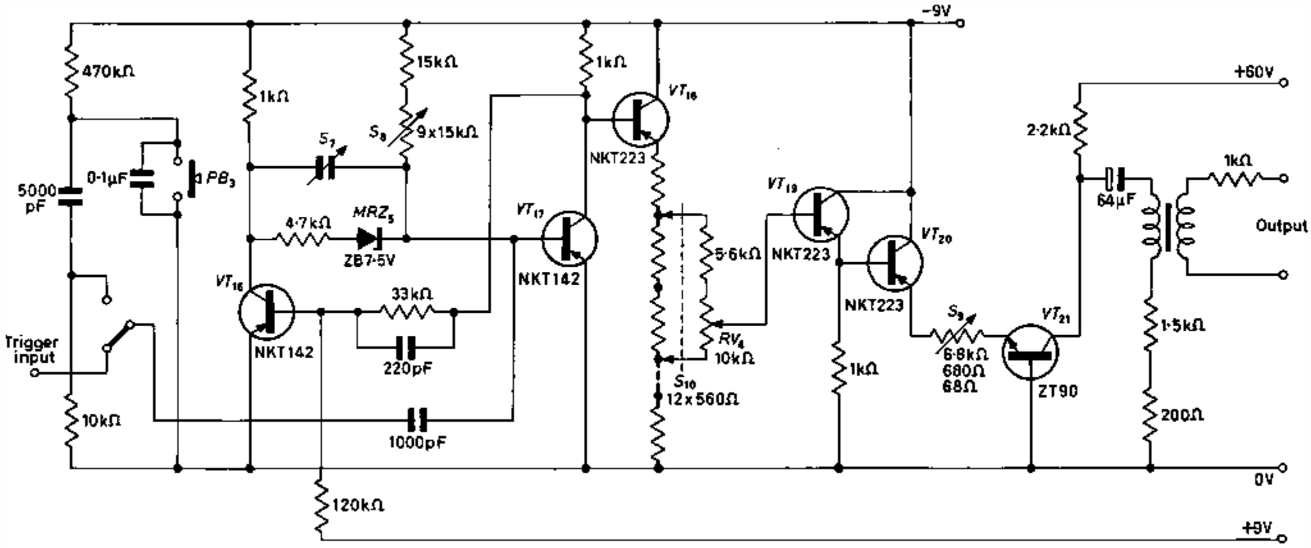


Fig. 3. Stimulator unit

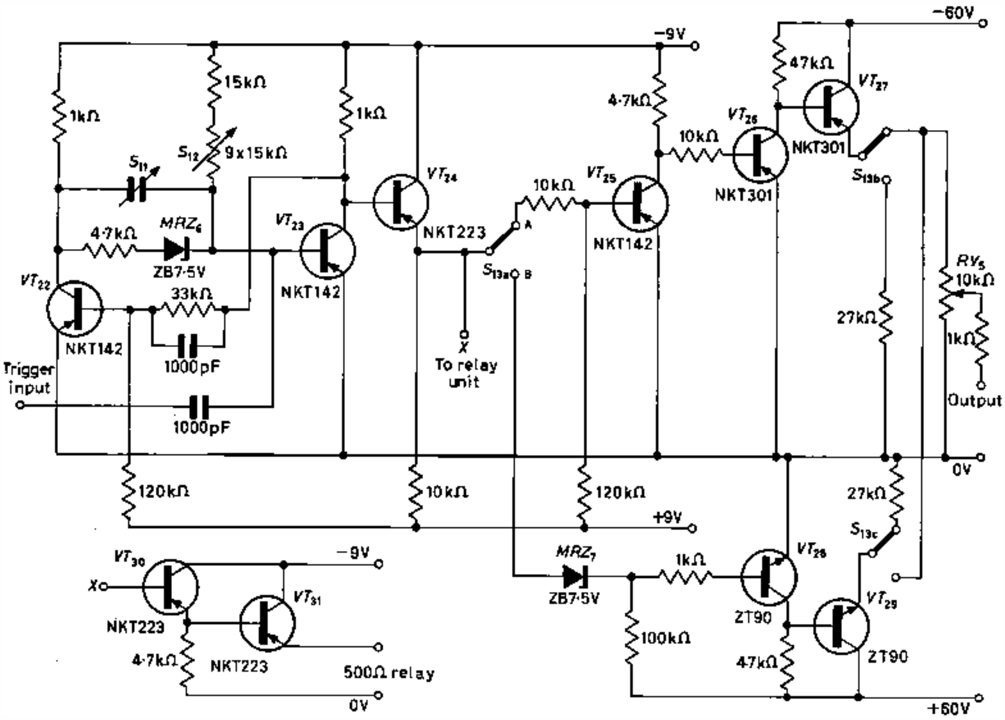


Fig. 4. D.C. output unit

All flip-flops. Switches S_3, S_7 and S_{11}

Range : 100μsec-1msec C—0·01μF

1msec-10msec C—0·1μF

10msec-100msec C—1μF etc.

Automatic triggering is from units (a) and/or (b), manual by push-button PB_3

The pulse duration circuit operates in the same way as the gate length flip-flop above. The output pulse is derived via the grounded base transistor VT_2. The switched coarse and intermediate, and the continuously variable fine current amplitude controls are S_9, S_{10}, and RV_4 respectively. The output transformer is type 5K. 9445/3 by Jorgen Schou of Copenhagen, and has been described by Guld[5]. This is by far the best transformer experienced for this kind of work. The output pulses remain tolerably rectangular, and of amplitude constant to about 10 per cent up to 10msec duration, when working at currents of 0 to 10mA (0 to 150V) into loads of from zero to 15kΩ. (A 'constant voltage' output can easily be arranged by driving the transformer from an emitter-follower instead, see also Fig. 4).

The d.c. input stage (Fig. 4), provides rectangular pulses, positive or negative going, of up to 60V. The pulse duration, controlled by flip-flop VT_{22}, VT_{23}, is variable from 10msec to 10sec in three ranges. Switch S_{13} applies the pulse to either output *A*, or to output *B*, at the same time switching the appropriate output to the final amplitude potentiometer. The relay is driven from the output of VT_{24}, through VT_{30}, VT_{31}.

Operation

The overall mode of operation has been described previously[1,6].

One apparatus has been used in neurophysiological research for 18 months, two others for more than a year. It is apparently reliably stable.

Power Supplies

The power requirements are as follows:

−9V at 500mA
+9V at 150mA
−60V at 100mA
+60 at 150mA

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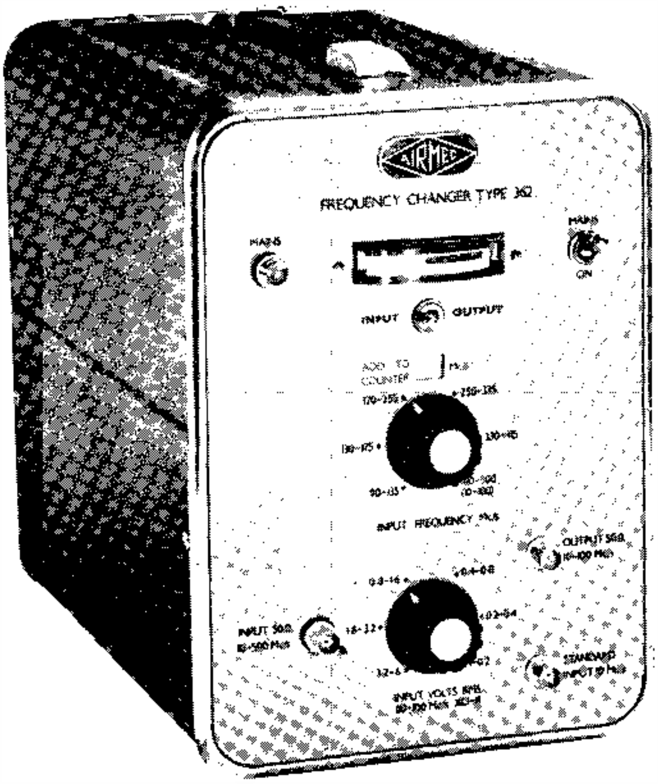

A Frequency Changer for High Speed Counters

By P. Adby*, B.Sc., and E. A. Edis†, B.Sc.(Hons.), A.M.I.E.E.

The frequency changer described extends the upper frequency limit of a typical 100Mc/s counter to 900Mc/s, while retaining the basic measurement accuracy of the counter. The design eliminates the continuous tuning control normally found in similar instruments, so that a series of frequencies can be measured rapidly by selecting only one of six switched ranges. The provision of input and output level monitoring permits an unknown signal to be measured without ambiguity. With the exception of an input buffer valve, the circuit uses semiconductors throughout and various points of interest in the design are described in detail.

(Voir page 847 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 854)

THE technique of measuring frequency by means of transistorized decimal counters is now well established. In such counters the input signal of unknown frequency is allowed to pass through a gate for a precisely controlled period of time to a decimal counter unit which produces a visual display on a read-out indicator. The gate period is controlled by means of frequency divider circuits from a reference crystal oscillator so that the display is an accurate measure of the input signal frequency. Further control circuits within the counter determine the display period and reset the decimal counter units to zero for the next cycle, when the measurement is repeated.

Providing direct coupling is used in the appropriate input signal circuits there is no lower limit to the frequency that can be measured but there is always an upper frequency limit defined by the following factors:

(1) The open and close transition times of the gate.

(2) The resolving time of the decimal counter unit, i.e. the smallest time interval between successive pulses or cycles of a sinusoid in which the numerical state of the counter can change.

The gate transition from close to open and vice versa must take place in time intervals considerably less than one period of the input signal, otherwise the unavoidable ± one count ambiguity present in all counters of this type[1] will be increased to a larger, indeterminate, and unacceptable error. The resolving time of one decade of a decimal counter unit, which is commonly made up of four binary stages, is longer than that of a single binary circuit because in the interval between the ninth and tenth input pulses the decade has not only to add one to its total, but also reset to the '0' condition. Papers have been published[2,3] describing digital decade counters accepting input frequencies up to 200Mc/s, but the normal course of development has resulted in commercially available direct reading counters up to 100Mc/s.

In these counters it is common practice to use an artifice to raise the frequency limit imposed by the above restrictions (1) and (2) and this is to divide the input frequency by cascaded 'divide by 2' binary stages so that the actual frequency applied to the signal gate and to the decimal counter units is well within their capabilities. Thus a typical 100Mc/s counter may have three 'divide by 2' stages preceding the gate so that when the input is 100Mc/s the frequency counted is 12·5Mc/s, but the read-out indicator is naturally scaled to read 100Mc/s. With this artifice the upper frequency limit is determined by the binary dividers and by using non-saturating circuits a figure of 100Mc/s has been obtainable in an economic design for some time.

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† *Airmec Ltd.*

The need to measure frequencies above 100Mc/s with the accuracy of the digital counter has assumed increased importance since communications are well established, for example, in the 100 to 500Mc/s band. A counter may be used to measure a frequency above its own upper limit by means of auxiliary equipment that transforms the unknown frequency into one within the counter pass-band. The ancillary instrument performing this function is known as a frequency changer or frequency convertor.

In this instrument the input signal is mixed with a standard frequency signal to produce a difference frequency which is measured by the counter. The standard frequency is derived by multiplication from the reference oscillator in the counter and therefore has the same accuracy as the counter itself. The unknown frequency is then simply the sum of the frequency displayed on the counter and the standard frequency, both of which are known to a high order of accuracy.

If the maximum frequency accepted by the counter is f_c and the reference frequency output from the counter is f_r the maximum input accepted by the frequency changer will be:

$f_{in} = f_c + nf_r$ where n is the ratio $\dfrac{\text{standard frequency}}{\text{counter reference frequency}}$

i.e. $f_{in}/f_c = 1 + (nf_r/f_c)$

So that for the n^{th} harmonic of the reference frequency the input frequency range of the counter is extended nf_r/f_c times by the frequency changer. There is thus a great advantage to be gained by using a counter with as high a maximum frequency limit (f_c) as possible since the number of harmonics (n) of the reference frequency (f_r) required to cover a given input frequency range is thereby reduced.

In a simple example if f_c is 100Mc/s and f_r 10Mc/s then for $n = 10$, 20, 30, 40, the frequency changer will cover up to 500Mc/s by selecting only four switched values of nf_r.

If the counter was not of a type capable of making direct measurements to 100Mc/s but only to, say, 10Mc/s then the number of harmonics of standard frequency required to extend its range to 500Mc/s is so numerous that a different technique is necessary. For example, the reference frequency (also 10Mc/s) is fed into a harmonic generator which produces a wide range of harmonics at 10Mc/s intervals up to 500Mc/s. A continuously tuned stage follows the harmonic generator and this must have sufficient selectivity to choose only one of the 10Mc/s harmonics which is passed on to the mixer. The tuning control is calibrated in terms of input signal frequency, at 10Mc/s intervals, and in operation is tuned to produce a difference frequency output from the mixer within the 0 to 10Mc/s range of the counter.

In all frequency changers it is important that spurious outputs from the mixer do not cause false readings on the counter, and to achieve this condition it is necessary that the input signal is fed to the mixer at the proper amplitude level. Since the input signal level may not conveniently be adjustable at source nor indeed may the level be known, a frequency changer should include an input attenuator and a level monitor to show the input to the mixer. A high degree of accuracy is unnecessary since the mixer will tolerate a reasonable range of signal amplitudes but the attenuator and monitor system must be free from gross amplitude errors over the whole range of input frequency covered by the frequency changer.

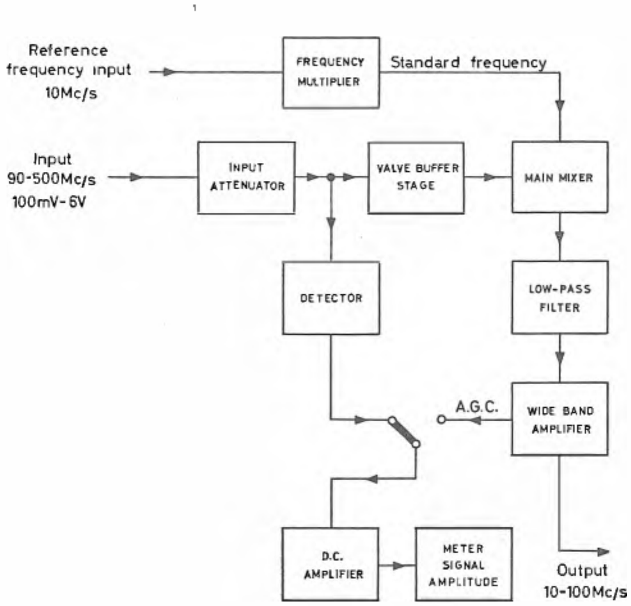


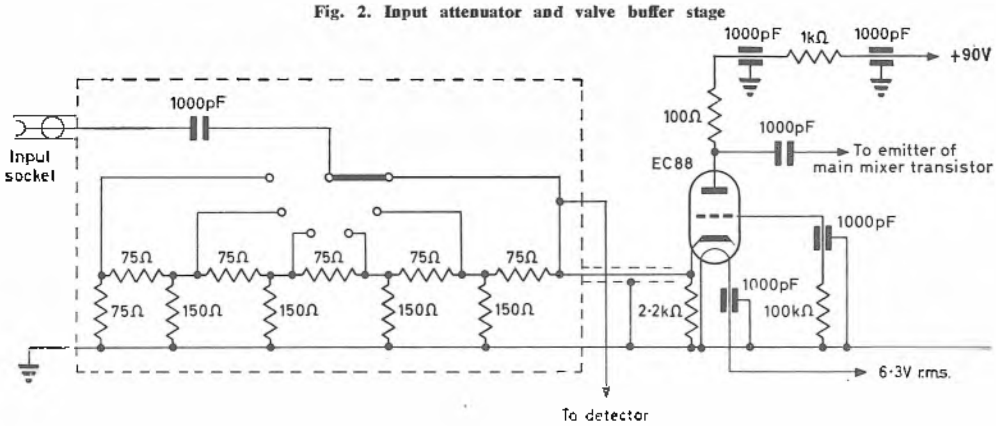
Fig. 1. Arrangement of a 90 to 500Mc/s frequency changer

The arrangement of a frequency changer designed specifically for simple operation is shown in Fig. 1. As an accessory for a 100Mc/s counter it makes accurate frequency measurements possible in the range from 90 to 500Mc/s and with the exception of an input valve buffer stage, semiconductors are used throughout.

The unknown input frequency is applied to an attenuator which is adjusted to provide a suitable signal level to pass via the grounded grid buffer stage to the main mixer. The signal level is monitored by a panel meter. In the mixer stage the unknown frequency mixes with a second signal of much larger amplitude to produce an output whose amplitude is essentially proportional to that of the unknown input. The second, larger signal fed to the mixer is the standard frequency mentioned above.

In this frequency changer only six standard frequencies are necessary, namely, 80, 120, 160, 240, 320 and 400Mc/s and this makes possible the simplicity of operation which is a feature of the design. The input frequency is always the counter reading plus one of the six standard frequencies. The mixer output is filtered so that only the required difference frequency passes to a 10 to 100Mc/s

Fig. 2. Input attenuator and valve buffer stage



aperiodic amplifier which raises the amplitude to a level sufficient to operate the high speed counter.

A number of interesting points of design arise in the application of this simple system to a commercial instrument and these are discussed as each section of the circuit is described.

Input Attenuator

The input attenuator shown in Fig. 2 is a standard ladder type with a constant input impedance of 50Ω. The resistor values are calculated for 6dB steps of attenuation assuming a buffer stage input impedance of 75Ω. This is a grounded grid stage and the input impedance is an approximation to 75Ω so that the actual voltage steps obtained are not precisely 2:1. The output resistance of the attenuator is 150Ω and is used as the cathode load of the valve. The resistors and input blocking capacitor are mounted in an aluminium casting to provide efficient screening from range to range.

Valve Buffer Stage

A valve buffer stage is used to isolate the input socket from the standard frequencies applied to the mixer. It is a grounded grid stage (shown in Fig. 2) which gives a low input impedance and provides good screening between input and output circuits. Efficient decoupling from the grid to earth is obtained by a lead-through capacitor soldered into a screen across the valve base and all grid pins are connected to this capacitor by very short leads. The effective anode load of the valve is the input impedance of the mixer transistor and is of the order of 50Ω. This low value ensures a sufficiently level frequency response up to 500Mc/s and the 100Ω anode feed resistor limits the output voltage swing to a value which will not damage the transistor. The minimum signal required at the input socket for satisfactory operation of the mixer is nominally 100mV r.m.s.

Main Mixer

The main mixer employs a high frequency transistor, and of the several circuit configurations which could be used the most suitable is that in which the input signal

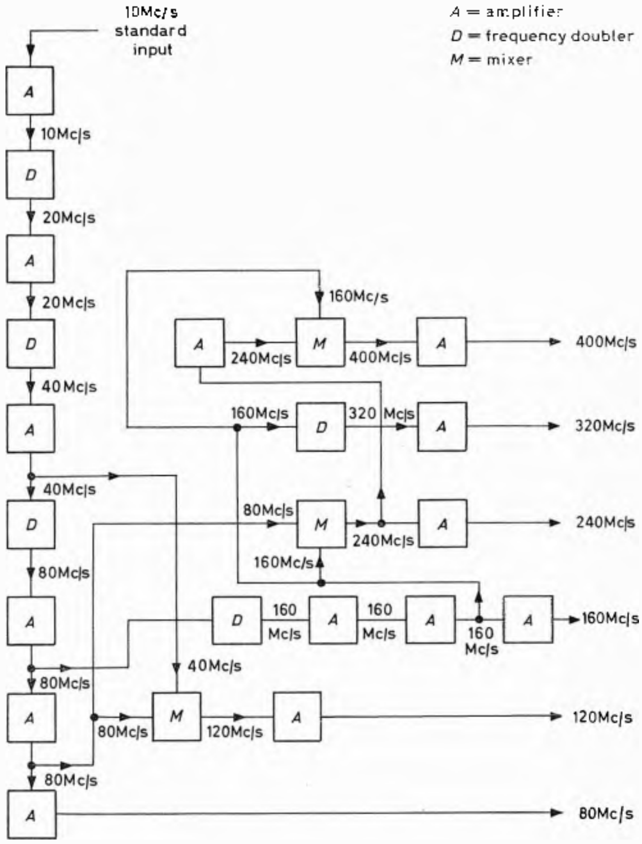
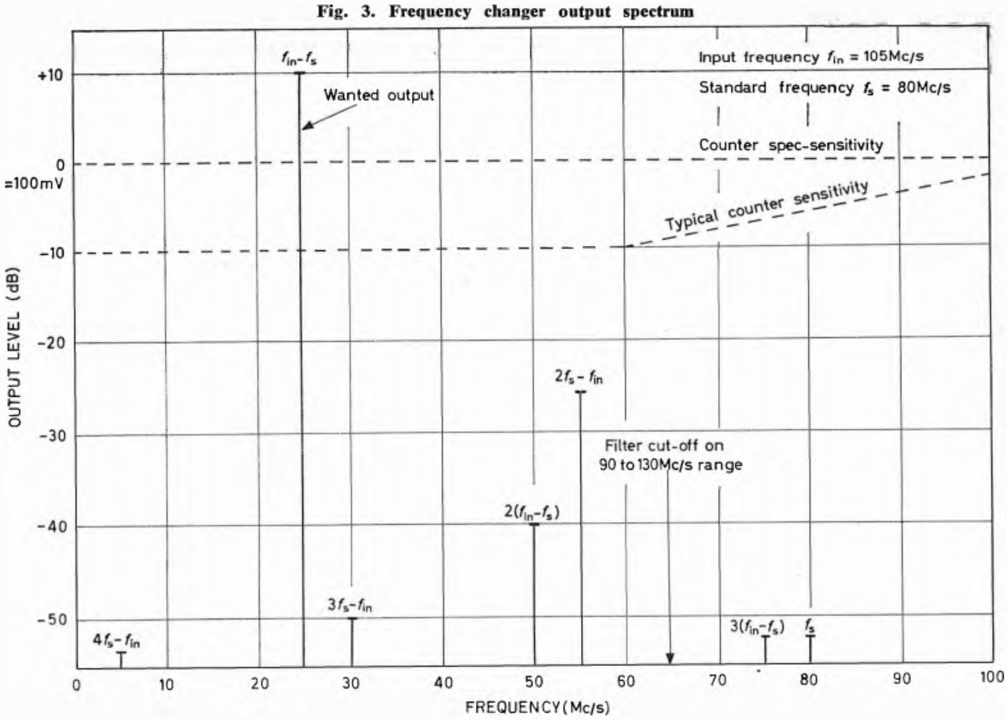


Fig. 4. Arrangement of frequency multiplier

is fed into the emitter at low impedance, while the high level standard frequency is applied at high impedance to the base. The transistor used is a T2029 which has an efficient base-emitter diode characteristic with low emitter input capacitance and provides a high conversion gain up to at least 100Mc/s. The collector load consists of a 50Ω low-pass filter in parallel with the 150Ω collector feed resistor and the transistor operates at a collector current of 1mA.

Fig. 3. Frequency changer output spectrum



The mixer produces output frequencies of the general form $nf_{in} \pm mf_s$ where n and m are integers; f_{in} is the unknown input and f_s the standard frequency. The wanted output is $f_{in} - f_s$ which must fall within the band 10 to 100Mc/s. Other signals generated by the mixer will also come within this band and experimental selection of operating bias and signal levels has resulted in these spurious outputs being more than 20dB below the wanted signal. Fig. 3 shows the relative amplitudes of wanted and unwanted signals at the output of the frequency changer under typical operating conditions.

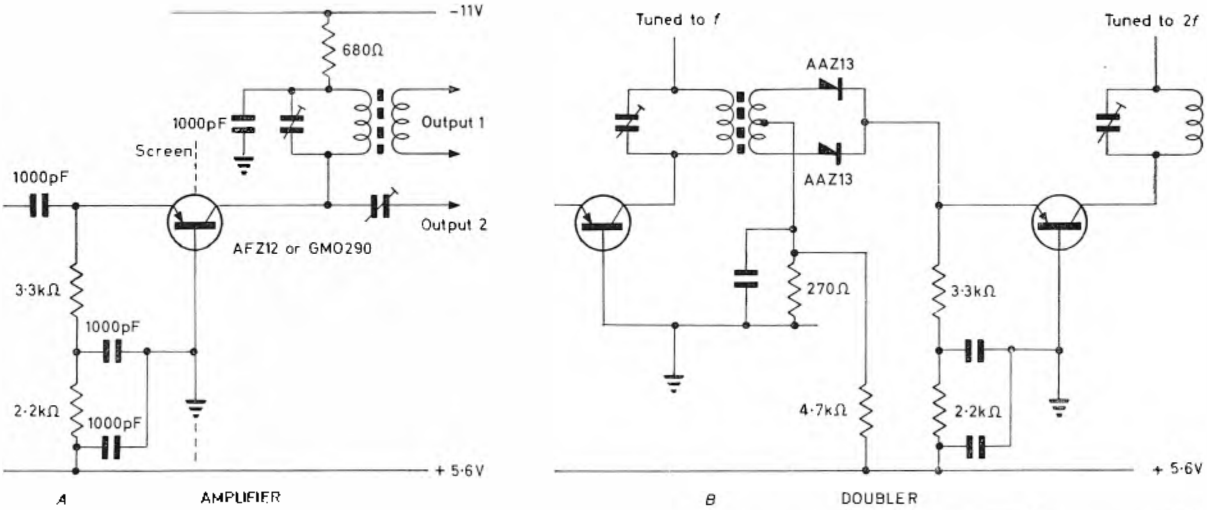
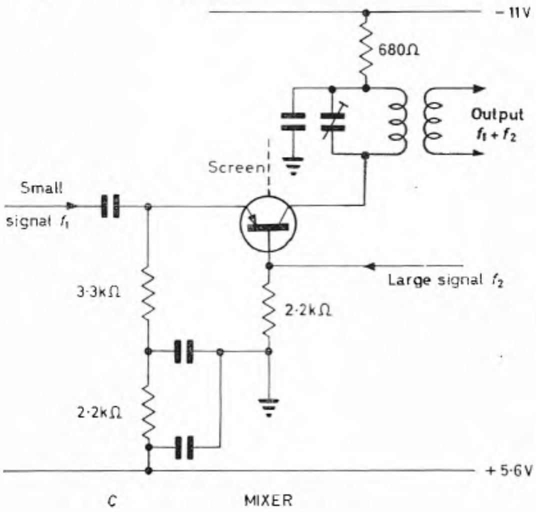


Fig. 5. Basic amplifier, doubler and mixer



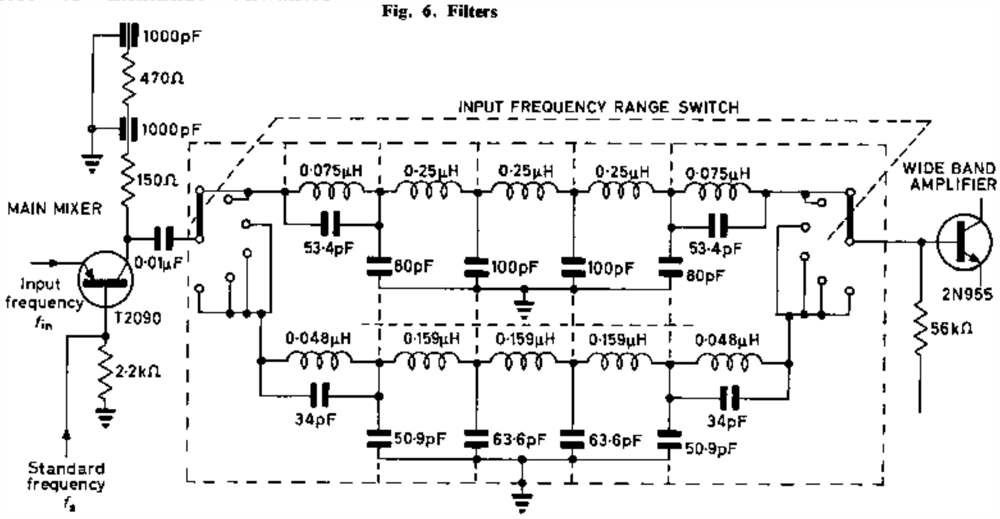
Frequency Multiplier

The standard frequency f_s fed to the main mixer is derived from the multiplier section and its value is chosen to avoid low values of $f_{in} - f_s$. It is undesirable for the counter to receive low frequency signals because of the difficulty of deciding whether the unknown frequency is just above or just below that of the standard. The lowest output frequency to the counter was chosen to be 10Mc/s. Some overlap of each range of input frequency is required for continuous cover. Standard frequencies were chosen at four 80Mc/s intervals with the first at 80Mc/s and an extra frequency at 120Mc/s. The latter is necessary because of the first range where the maximum wanted frequency after mixing must be less than 80Mc/s to avoid confusion with the 80Mc/s standard which the filters are designed to reject.

The multiplier system therefore provides frequencies of 80, 120, 160, 240, 320 and 400Mc/s from the 10Mc/s reference signal of the counter. They are generated at a level of 300mV r.m.s. and selected by a wafer switch for connexion to the main mixer. A block diagram of the multiplier circuit is shown in Fig. 4.

The complete circuit includes 17 transistor stages but only three basic configurations are used, namely a doubler, an amplifier and a mixer, each of which is shown in Fig. 5. The amplifier (Fig. 5(a)) is an unneutralized common base amplifier working between positive and negative rails. The base is connected directly to an earthed screen across the transistor to minimize unwanted feedback from output to input. At frequencies up to 80Mc/s transformers are employed to couple the input and output circuits to obtain highest efficiency (output 1, Fig. 5). Above this frequency use is made of small matching capacitors (output 2, Fig. 5) which present a high impedance across the collector tuned circuit even when connected to the low input impedance of the following stage. The reactance of the these capacitors is Q times the input resistance of the next transistor, where Q is the magnification factor of the unloaded tuned circuit in the preceding collector. The value of Q is obtained from the relation $1 + Q^2 = R_o/R_L$, where R_o is the collector output resistance, R_L is the load resistance. Small variable ceramic capacitors are used in these positions and are preset to adjust the output voltage for optimum level. Conventional wound coils with adjustable cores form the collector inductors in all amplifier stages up to 320Mc/s but the small inductance required in the 400Mc/s tuned circuits is obtained from pieces of

Fig. 6. Filters



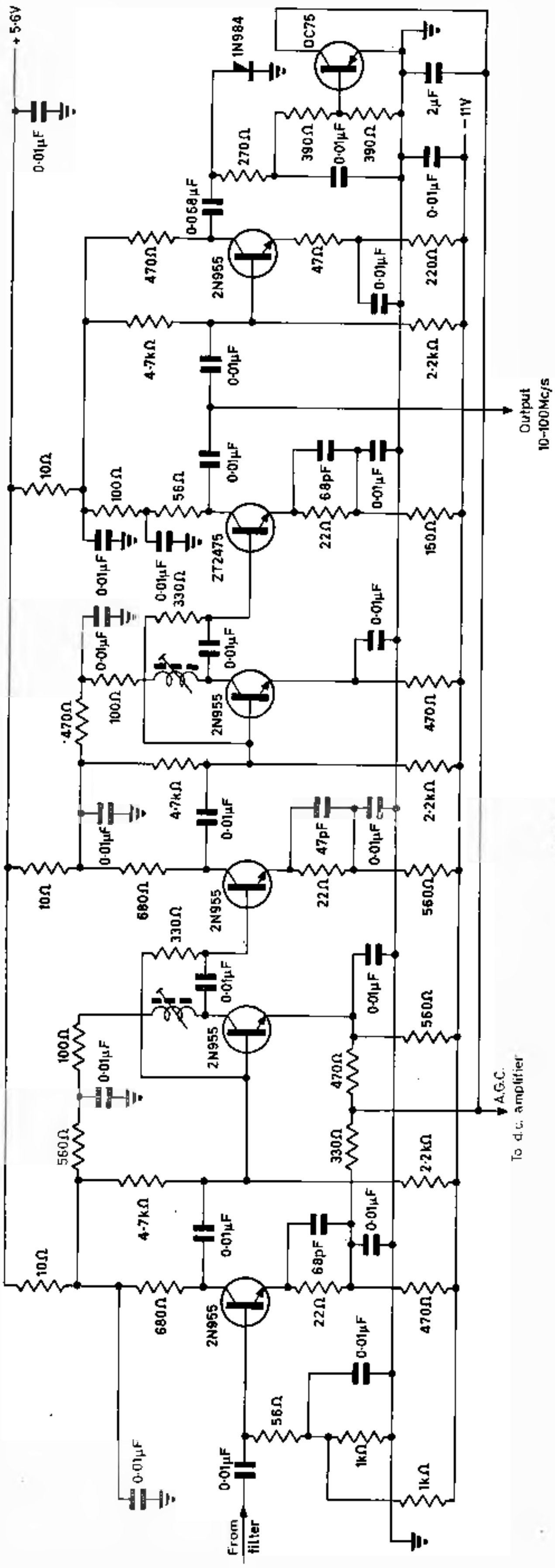


Fig. 7. Wide band amplifier (10 to 100Mc/s)

coaxial cable less than one quarter wavelength long, which are short-circuited at the far end. These present an inductive reactance to the collector given by $jZ_0 \tan 2\pi (l/\lambda)$ where:

Z_0 = characteristic impedance of cable
l = length of cable
λ = wavelength

The frequency doubler is shown in Fig. 5(b). Two diodes form a full wave rectifier and are fed from a centre-tapped transformer winding coupled to a tuned circuit in the collector of a preceding amplifier stage. The centre-tap has positive d.c. bias to offset V_{BE} of the following transistor which would apply reverse bias to the diodes.

In this arrangement the signal at the common point of the diodes is rich in even harmonics but the fundamental frequency is suppressed. It is amplified by the following stage whose collector circuit is tuned to the second harmonic by preset trimmer capacitors.

The basic mixer used in the frequency multiplier is shown in Fig. 5(c). The signals to be mixed are transformer coupled to the base and emitter of the transistor at low frequencies and by small matching capacitors at high frequencies. For efficient operation the signal levels applied are approximately 40mV at the emitter and 300mV at the base. The multiplier mixers have the same order of input impedances at the base and emitter as the main mixer but their collector circuits are tuned to the sum instead of the difference of the two input frequencies.

Filters

Two separate filters are used, one for the two lower ranges 80 and 120Mc/s and the other for the remaining ranges. The filters are conventional and consist of three constant-*K* sections with *m*-derived half-sections at each end.[4] An *m* of 0·6 is used for a good impedance match, and both filters are terminated at their outputs by their characteristic impedance of 50Ω. They are mounted in a screened sectional box to give minimum intersection capacitance coupling.

The low-pass filter used on the two lower frequency ranges has a cut-off frequency of 64Mc/s and an infinity point at 80Mc/s. The infinity point is chosen to coincide with the frequency of a high level unwanted output from the mixer on the 80Mc/s range. For the higher frequency ranges, which are spaced at intervals of 80Mc/s, the filter has a cut-off frequency of 100Mc/s and an infinity point of 125Mc/s. The pass-band here allows the full range of the counter to be used. The circuit diagram of the filters is given in Fig. 6.

Wide Band Amplifier

The wide band amplifier is composed of alternate series and shunt feedback stages in a circuit arrangement shown in Fig. 7. The input and output stages are both of the series type. High frequency compensation is provided by small capacitors connected across the emitter feedback resistors of the series stages and by inductors in series with the collector loads of the shunt feedback stages. The great advantage of this type of amplifier is that the impedance mismatch between stages gives small interaction between them which leads to easier design. The feedback defines the performance of each stage and its overall gain[5].

The first four stages operate at an optimum current of 10mA which gives a good compromise between gain-bandwidth product and power consumption. The output stage requires higher current in order to drive an effective load of 25Ω formed by a 50Ω resistor in the collector circuit and a 50Ω coaxial cable which is terminated by the high speed counter. The overall voltage gain is about

140 and the amplitude frequency characteristic is typically 3dB down at 5Mc/s and 110Mc/s.

The disadvantage of employing an amplifier of fixed gain is that considerable amplitude changes in output signal may occur even though the unknown input signal to the frequency changer is constant. These changes may be sufficiently great to upset the operation of the counter and are due to variations in mixer efficiency as the standard frequency is switched from range to range; to the frequency response of the input circuit and level monitor, and other circuit tolerances. They are overcome in this instrument by an automatic gain control system which reduces gain on large input signals and

gain is defined by the ratio of the emitter resistors in each of the pnp-npn pairs, but in the present arrangement it is modified by the low value meter resistance. Good temperature stability is obtained without the necessity to match the transistors since the output signal is independent of the forward current gain of the transistors[6].

A number of preset potentiometers are included in the circuit. One of these, in series with the meter, compensates for the highly non-linear amplitude characteristic of the rectifier diodes at low values of input signal, in conjunction with a diode connected across the meter movement. Another potentiometer controls the sensitivity by resistor coupling between the two emitters, while a third potentio-

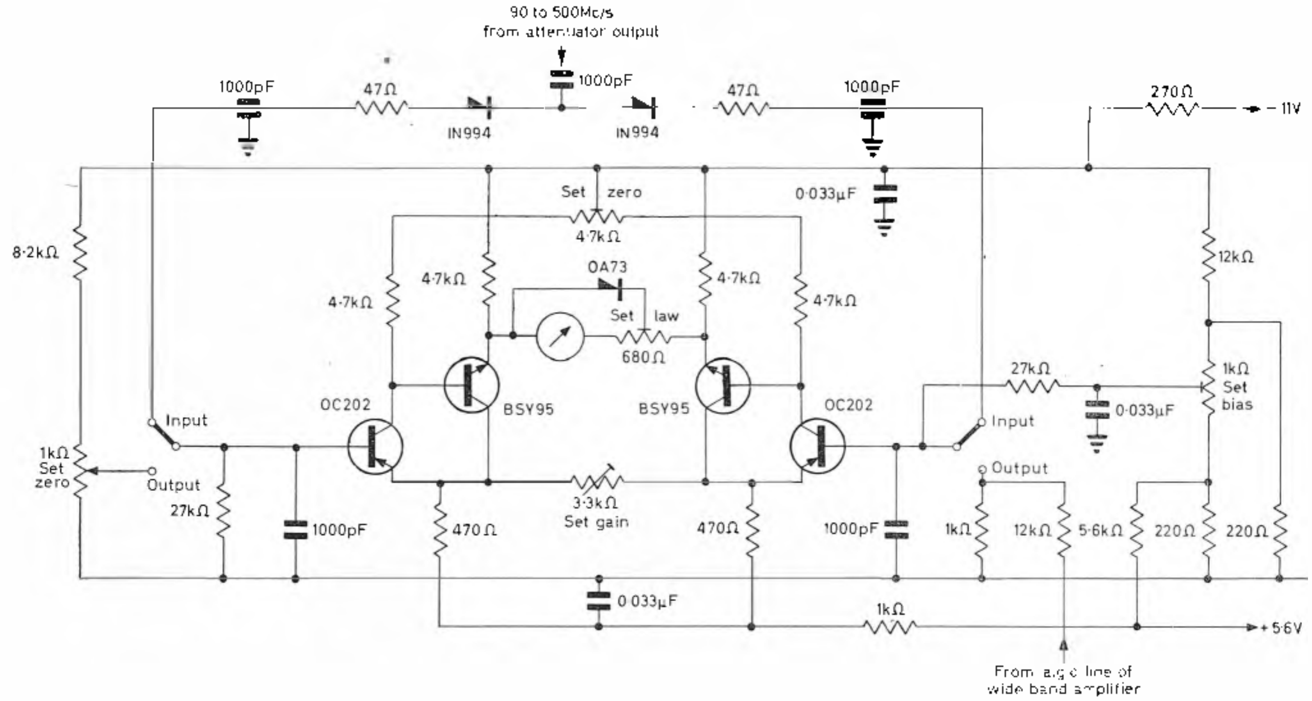


Fig. 8. Detector and d.c. amplifier

prevents excessive distortion. The a.g.c. is provided by an extra stage *RC*—coupled to the output of the wideband amplifier and having a gain of approximately three. It drives a diode detector. The rectified output controls a low frequency germanium transistor which draws current through the emitter resistors of the first two stages of the wideband amplifier, thereby reducing its overall gain.

Detector and D.C. Meter Amplifier

The circuit of the detector and d.c. amplifier is shown in Fig. 8. The detector monitors the level of the 90 to 500Mc/s unknown input frequency and feeds a d.c. amplifier which drives a meter. It makes use of germanium gold-bonded diodes with a recovery time of 2nsec to achieve a flat response to 500Mc/s. Small resistors in series with the diodes damp out resonances and careful attention to mechanical detail reduces stray inductance and capacitance to the minimum. The load resistance of the rectifier circuit is about 20kΩ and its value is a compromise between a high value which would be undesirable for the stability of the d.c. amplifier and a low value which would reduce the meter sensitivity. The d.c. amplifier is basically an emitter-coupled pair of pnp transistors modified by negative-feedback from collector to emitter by a pair of npn transistors. With this circuit the

meter provides a differential change in the value of the two collector loads to set zero on the meter. The remainder set the d.c. voltage levels to balance the circuit for correct operation.

The same meter circuit is used to monitor the changes in the a.g.c. circuit of the wideband amplifier. A second set-zero control operating only in this condition varies the base potential of one of the input transistors to restore the circuit balance. By monitoring the a.g.c. level it is possible to obtain a comparative measurement of signal output to the counter. Thus, when the input frequency is entirely unknown the correct range may be selected by noting the one on which the output level, as indicated by measurement of the a.g.c. voltage, is a maximum.

Conclusions

As mentioned above the feature of this instrument is the simplicity of its operation. It can be used by unskilled personnel for routine tests as well as for more sophisticated measurements in development or research laboratories. When the approximate value of the input frequency is known it is only necessary to adjust the input attenuator (or input signal level) to bring the input meter reading to a level indicated by a green band on the meter scale and then to switch to the appropriate frequency range. The range switch indicates the number of megacycles which are to be added to the counter reading to obtain the input frequency. When the input frequency is completely unknown (except that it is within the range of the instrument) the input attenuator is adjusted to the correct level as before and the meter then switched to read the output level. Each frequency range is selected in turn, starting at the lowest, and the first of the six ranges which gives a large output on the meter is the one required.

Although designed specifically for use up to 500Mc/s this frequency changer can, with care, be used to measure frequency accurately up to 900Mc/s. This is achieved by mixing the unknown frequency with the second harmonics of the 240, 320 or 400Mc/s standard frequencies. For example, if the 'unknown' frequency is 850Mc/s a reading of 50Mc/s will be obtained on the counter when the frequency changer is switched to the 400Mc/s range. This would not be confused with an input frequency of 450Mc/s since its level under these conditions would need to be the order of 1 to 2V.

A further facility obtained with this type of frequency changer is its use as a pre-amplifier. When the counter reference frequency (10Mc/s) is disconnected from the frequency changer no mixing takes place and the latter will provide a gain of 10 over the frequency range 10 to 100Mc/s, so enabling a 10mV signal to operate a counter which normally requires 100mV.

Frequency changers that have been available hitherto, have required careful manual tuning by the operator to select the correct harmonic of the counter reference frequency which has to be added to the counter reading. The design described above takes full advantage of the wide direct-counting frequency range of a modern high speed digital counter to eliminate this tuning operation, which requires great care and is a possible source of error. In the future, tunnel diode logic circuits are likely to extend still further the basic range of digital counters, and it is feasible that frequency changers will employ coaxial techniques, rather than lumped constant circuits. A simple frequency changer could then be developed in the gigacycle per second region, for which the transfer oscillator system is now widely used.

Acknowledgments

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A Small Size Computer

International Computers and Tabulators Ltd has produced a new full-scale computer for those organizations where the investment necessary to install a larger machine cannot yet be justified because of the limited size of the organization's operations. The new machine, the I.C.T. 1901, is easy to install and operate, requiring no special computer room, no false ceiling or trunking, no false floorings, and no voltage regulators. Simple air filtration and an ordinary single-phase electrical supply are all that is necessary for installation of the 1901 in any office, factory or laboratory.

A conventional magnetic-tape unit can be used with the 1901 computer but one of the several new peripheral devices which have been specially developed for this machine is a magnetic-tape unit employing a 'push in' cassette containing continuous loops of 1in magnetic tapes with eight recording tracks.

These tape units operate at 10 000 characters/second. The cassette principle permits easy loading and unloading. There is no threading or un-threading of tape; the cassette is simply pushed in or pulled out as required.

To aid the transition from punched-card to magnetic-tape filing, a 'unit record' system of storing information has been devised. The new tapes hold data in blocks, each containing 80 characters. One block can thus accommodate the data read from one 80-column punched card. Each group of four cassettes can hold three and a half million characters in block form and hence accommodate the data from 44 000 80-column cards.

Other new peripheral devices specially developed for the 1901 computer include: a 40-column card read operating at 600 cards/min, an 80-column card reader operating at 300 cards/min, a card reader/punch reading and punching at 300 cards/minute, a card punch operating at 300 cards/min, a line printer operating at 300 lines/min.

To simplify further operation of the I.C.T. 1901 or the transferring procedures from punched-card tabulator and calculator installations, a new commercial programming language has been developed. The language is known as NICOL (NIneteen hundred COmmercial Language). It has a simple format related to tabulator concepts which enables users readily to transfer their work without a high level of programming skill. In addition to NICOL, the PLAN, COBOL, RAPID-WRITE and FORTRAN programming languages are applicable.

A brief specification of the 1901 computer is given below.
Core store cycle-time, 6μsec
Core store size (words), 4 096: 8 192: 16 384
Surface interface channels, 3 to 6
Arithmetic times:

Add/subtract, 34μsec; multiply, 4 to 7msec; divide, approx. 7msec; jump, 21μsec.

Word length, 24 binary digits plus parity.

The cassette containing the continuous loop of magnetic tape being loaded into the reader unit

Field Effect Transistors for Biological Amplifiers

By R. E. Webb*

A brief statement of requirements for amplifying bioelectric potentials prefaces information on the use of field effect transistors for this purpose. Some comparisons between vacuum electrometer valves and field effect transistors are made and details of a high impedance amplifier are given. Circuit noise and its reduction is also briefly mentioned.

(Voir page 847 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 854)

THE purpose of a biological amplifier is to provide a high impedance and gain for bioelectric potentials. These may be either intracellular, i.e. measured across the walls of cells by the introduction of a micro-electrode into the cell, or extracellular, i.e. potentials associated with the conducting medium surrounding active tissues. Signal levels in the ranges $10\mu V$ to 10mV and 0·5mV to 200mV for extracellular and intracellular measurements respectively may be encountered. The resistance of a micro-electrode can be as high as $100M\Omega$ and signal rise times may be of the order of $20\mu sec$, therefore the input capacitance of the amplifier must be less than 0·2pF if distortion is to be avoided. In order to measure 'resting', i.e. steady state, potentials the amplifier must be d.c. connected. This requirement highlights the problem of amplifier input current which if excessive may modify the properties of a cell being investigated, polarize the electrode in use, generate excessive noise, etc. Depending upon the application, maximum permissible input currents range, from $10^{-9}A$ to $10^{-14}A$.

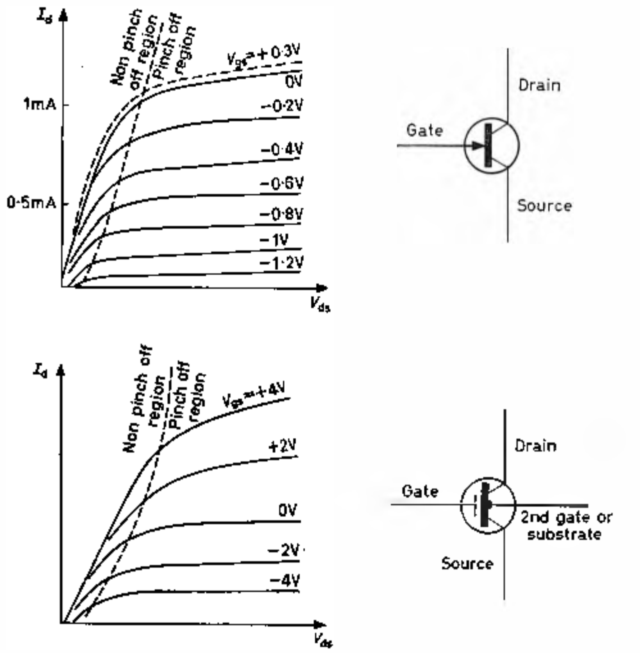

Fig. 1. Symbols and typical characteristics for a unipolar (upper) and an insulated gate f.e.t.

These high impedance requirements have, in the past, necessitated the use of vacuum electrometer valves[1] as the input stage of biological amplifiers. Other high impedance input arrangements[1], e.g. modulation by mechanical means, are unsuitable for the rise times of the signals involved.

The semiconductor counter-part of the vacuum electrometer valve is the field effect transistor[2] (f.e.t.). Potential advantages of this device are: lack of microphony, small size, operational simplicity, no heater requirements.

At the present time there are two main classes of f.e.t., unijunction[3] and insulated gate[4] types. They may be either single gated, i.e. three terminal, or double gated, i.e. four terminal, devices. The double gated type has a source (s), gate (g), second gate (g_2) and drain (d). These terminals can be associated with the cathode, control grid, screen grid and anode of a vacuum pentode valve. With the three terminal version there is no second gate. Gate 1 (g) is sometimes referred to as the modulator and the second gate (g_2) as the substrate. Static characteristics, and symbols (these have not yet been standardized) for the two classes of f.e.t. are shown in Fig. 1.

The input circuit of the unijunction type consists of a reversed biased silicon diode which must not be turned on, but the insulated gate device has a very high resistive layer separating the gate from the other electrodes and may be taken either positive or negative with respect to the source.

Attention is drawn to the pinch-off voltage (V_p) of an f.e.t.[5] This is the value of V_{gs} for near zero drain current when V_{ds} is zero. V_p, which for the insulated gate type of f.e.t. can be positive, zero, or negative, is also a function of V_{ds}. Manufacturers supply f.e.t. with nominal values of V_p within different ranges.

An equivalent circuit and typical element values for f.e.t. are given in Fig. 2 and Table 1 respectively.

Although not new devices interest in f.e.t. for electrometer purposes has been renewed recently with manufacturing advances which promise to make the f.e.t. characteristics superior to those of the vacuum electrometer valve. These advances have been associated with the insulated gate type, such as the Mullard 95BFY which has a typical input impedance of $10^{12}\Omega$ in parallel with 4pF. The capacitance is split between the gate-source and

TABLE 1

Typical Element Values for F.E.T.

ELEMENT	NOMINAL VALUES AT 20°C.		
	UNIPOLAR CLASS		INSULATED GATE CLASS
	3 ELEMENT	4 ELEMENT	
Gate leakage current	$<10^{-8}A$	$<10^{-9}A$	$10^{-12}A$
C_{gs}	10pF	10pF	3pF
C_{gd}	5pF	1·5pF	1pF
g_m	0·1mA/V	1 to 3mA/V	1mA/V
r_{ds}	0·3 to3MΩ	0·25MΩ	0·1MΩ

* *Queen Elizabeth College, University of London.*

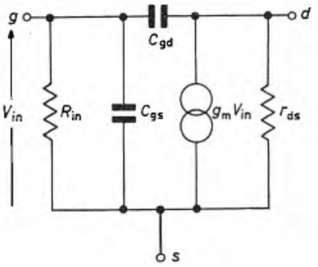

Fig. 2 An equivalent circuit for the f.e.t. in the pinch off region

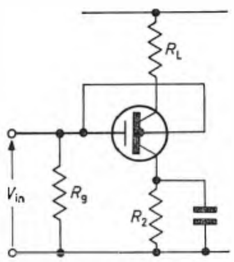

Fig. 3. Basic amplifier circuit

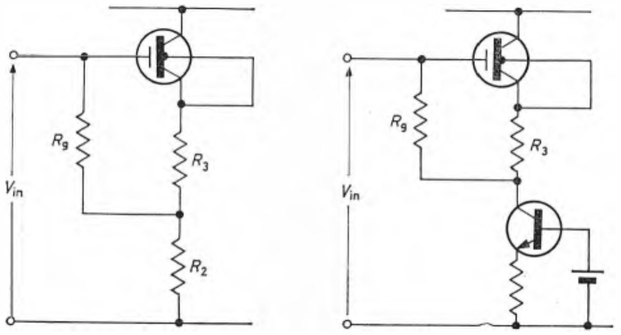

Fig. 4. Bootstrapped source followers

gate-drain in the approximate ratio of 3:1. Circuitwise the effective input resistance will be determined by the value of the gate leakage resistance R_g, see Fig. 3. The maximum value of this resistance is limited by changes in bias conditions resulting from gate leakage current changes with temperature. The presence of a permanent gate leakage resistance is necessary to prevent charges, which could destroy the device, forming at this electrode. For the same reason it is advisable to keep all electrodes connected together while soldering into the circuit.

To meet the requirements previously outlined for the high impedance amplifier it is necessary to select an f.e.t. with a sufficiently low leakage current and to reduce the input capacitance considerably. Also, if signal attenuation is to be avoided the effective value of R_g must be substantially increased. The last two requirements can be achieved by utilizing voltage negative feedback, and the simplest circuit configuration using this method is the boot-strapped source follower, see Fig. 4. This will have a gain:

$$A' < 1 \simeq \frac{g_m (R_2 + R_3)}{(1 + g_m R_3) + g_m R_2} \quad (r_{ds} \gg R_2 + R_3)$$

The input impedance will be:

$$Z_{in}' = Z_{in}(1 + \beta A)$$

where $\beta = 1$

Z_{in} = input impedance without feedback

A = internal gain

$= A'(1 - A')^{-1}$

If R_g is $10^7\Omega$ and care is taken with the circuit layout the boot-strapped source follower can provide an input resistance greater than 200MΩ shunted by approximately 1pF. Note that in Fig. 4 the substrate is connected to the source. This decreases the input capacitance at the expense of g_m.

Commonly used electrodes have an internal resistance of the order of 20MΩ. This resistance is sufficiently low to allow the use of the basic boot-strapped source follower. Unfortunately, to obtain a gain close to unity a fairly large value of R_2 is required. With a given voltage supply this will limit the current and therefore the g_m of the

TABLE 2

Measured Characteristics of Fig. 6 Circuit

Input resistance	$>10^9\Omega$
Input capacitance	<0·1pF
Leakage current	$<10^{-11}$A
Voltage gain	unity
Rise time (voltage source)	3μ sec.
Output resistance	25Ω
Noise at output (50 kc/s bandwidth)	80μV.

stage. To increase the effective value of R_2 without decreasing g_m a grounded base transistor can be used as the source load.

If an input impedance higher than that obtainable with the basic boot-strapped source follower is required then it is necessary to 'augment' the source follower by increasing the internal gain A, and make provision for reducing the gate to drain capacitance. The latter can be achieved by arranging for the drain to also follow the input signal. This is illustrated in Fig. 5 where it can be seen that if $V_1 = V_2 = V_3$ the input impedance is infinite. A circuit utilizing this principle and designed to meet the requirements of a 100MΩ probe is shown in Fig. 6. With a nominal internal voltage gain of 10^3 a theoretical input impedance of $10^{10}\Omega$ in parallel with 0·005pF is possible. These values are in fact limited by constructional considerations. Measured characteristics for the circuit are given in Table 2.

In Fig. 6 the input and output terminals are arranged to be earthy. A fine control on one of the stabilized power supplies can be used as a set zero control.

If the required input conditions for the amplifier are to be fully realized care must be taken with the circuit layout and choice of materials used to hold the first stage components. Providing this material and the f.e.t. header are not contaminated in any way the frequency response of the system will be determined by the electrode used and the cells being investigated. The former will have a high resistance and both distributed capacitance.

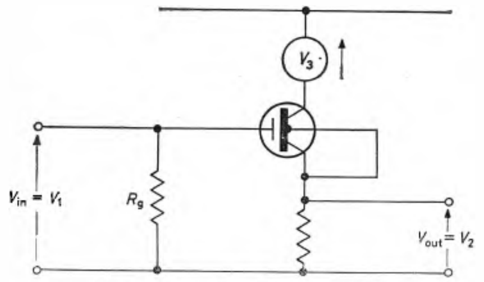

Fig. 5. Removing the effect of C_{gd}

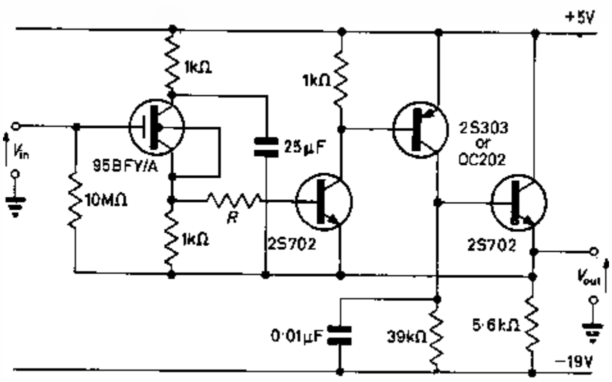

Fig. 6. High input impedance amplifier

R = temperature sensitive resistor, a.o.t. for zero drift at room temperature

Stray capacitance associated with the amplifier can be reduced by double screening the amplifier and input lead. The inner screen is connected to the amplifier output terminal. This reduces the capacitance to the outer (earthed) screen by approximately the same amount as the input impedance is increased.

Table 2 indicates that adequately small values of input capacitance can be obtained along with an input resistance which would not significantly shunt a high resistance micro-electrode, but the gate leakage current may be too large for some applications. The maximum expected value for the 95BFY is 10^{-10}A, with a nominal value of 10^{-12}A. Eventually Mullard hope to obtain leakage currents of the order of 10^{-15}A. This compares with 10^{-14}A which is of the order of grid current for a vacuum electrometer valve when operated at the usual zero slope point on its control grid characteristic.

To reject common mode signals, e.g. hum and drift; it is usual to operate two identical biological amplifiers differentially. The nature of noise forbids its reduction by this means, therefore to discern small signals precautions must be taken to ensure noise is at a minimum.

Noise is contributed by all circuit components[5] including the electrode, and is due to several causes. It is distributed in the frequency spectrum as shown in Fig. 7. Excess noise, αf^{-1}, is responsible for the increase at low frequencies and can be attributed to a diffusion phenomenon in granular type resistors, semiconductors, and to valves. In the latter case it is known as Johnson or flicker noise. Transistors are responsible for the increase at the higher frequencies; the turning point being dependent on their alpha cut-off frequency. Thermal noise, $\alpha(T\delta f)^{\frac{1}{2}}$, where T is the absolute temperature and δf the noise bandwidth, is contributed by all components.

It is evident that signal-to-noise ratios can be improved by using suitable components and by limiting the bandwidth of the amplifier. The latter procedure may result in distortion of some transient signals but this may be preferred to excessive noise. It should also be noted that a low noise transistor has a better noise figure than a low noise valve—both being operated under the required low noise conditions.

Noise reduction filters with variable cut-off frequencies can be effectively employed at both ends of the frequency band when the monitoring of d.c. levels is not required. They can be housed with the power supplies in a control unit some distance from the high impedance amplifier. Circuit wise, the filters should be placed after any high gain amplifiers in the chain.

Measurements indicate that the noise figure of an f.e.t. may be limited by its pinch-off mechanism[6], the noise figure being improved by operating the device in the non pinch-off condition. It is likely, however, that noise signals from other sources will be sufficiently greater than those contributed by the f.e.t. A 100MΩ electrode with a 10kc/s bandwidth would generate more than 100μV of thermal noise.

Capacitive neutralization and frequency equalization[7] as a means of reducing the effects of stray capacitance have been investigated. Both of these methods introduce extra circuit noise.

Conclusions

The insulated gate class of field effect transistors can exhibit input characteristics comparable to existing vacuum electrometer valves and has the advantage of mechanical ruggedness which leads to better electrical stability. With a constant supply voltage, zero drift of the device can be very small. By using conventional negative feedback techniques and an f.e.t. for the input stage adequately high values of amplifier input impedance can be obtained for monitoring biological potentials via the highest impedance probes. The complexity of the amplifier required will depend upon the characteristics of the probe to be used and the signals to be monitored.

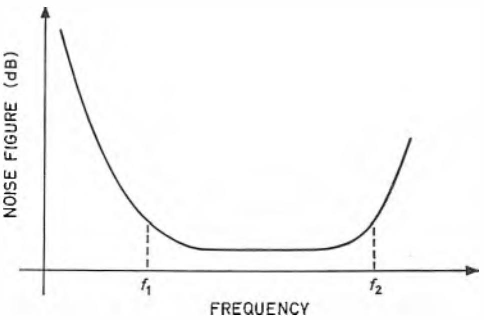

Fig. 7. Typical noise spectrum
For the circuits being considered, f_1 will be about 100c/s and f_2 about 100kc/s

Acknowledgment

The investigations were carried out at Queen Elizabeth College by the Electronics Department on behalf of the Department of Physiology.

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A New Vibrator Installation

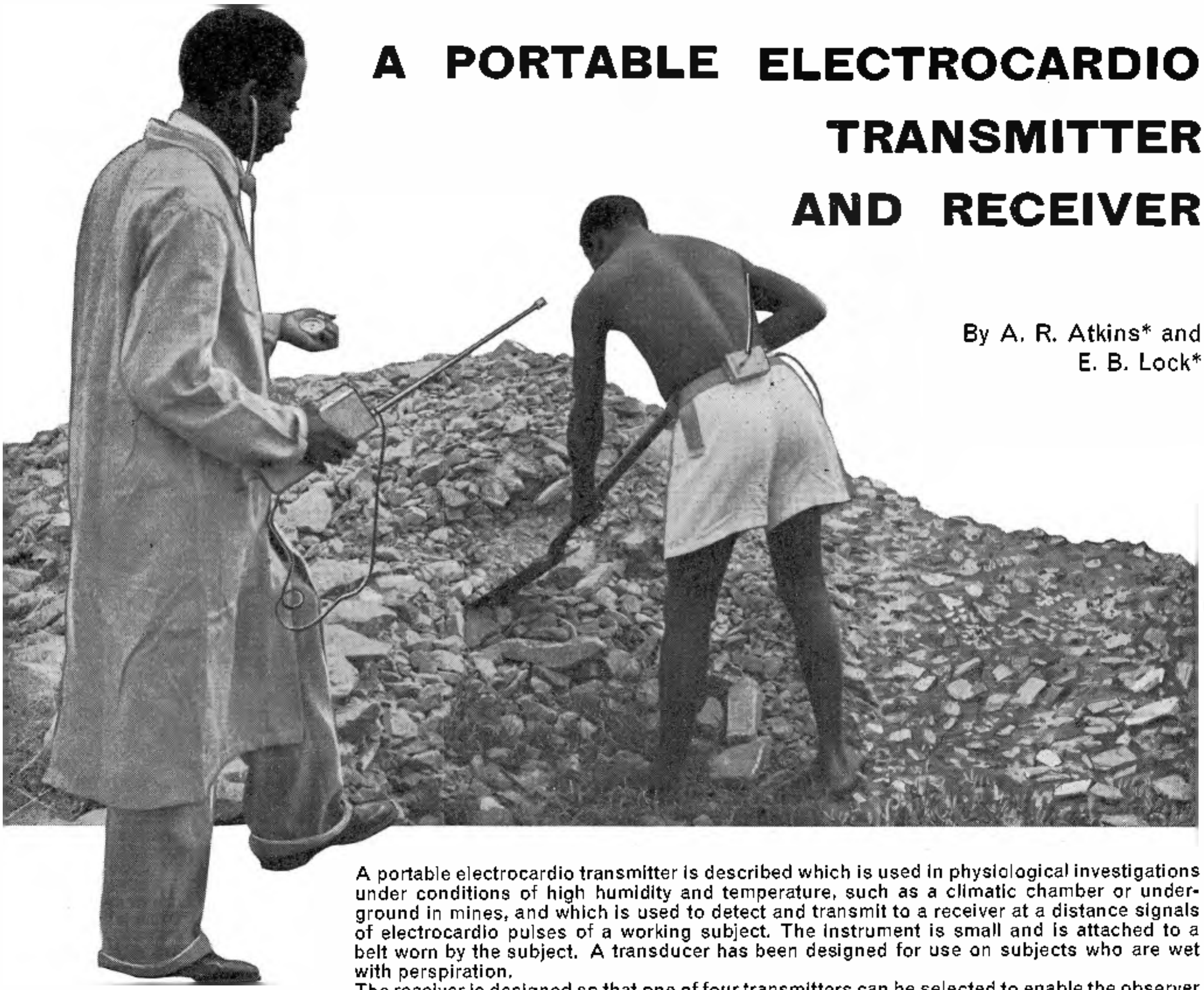
Components fitted to supersonic aero-engines have to withstand high operating temperatures and vibration conditions when in service, and to ensure adequate reliability it is essential that they are tested under equivalent conditions. For this reason, Vactric Control Equipment Ltd has just installed a new oil-cooled vibrator in its Environmental Test Laboratory at Morden in Surrey. This brings to a total of four the number of vibrators available for the resonance search and endurance testing of servo and other electro-mechanical components.

Made by Derritron Electronic Vibrators Ltd, the new vibrator has facilities for testing components at elevated temperatures with an amplitude of $\pm\frac{1}{2}$in at the lower end of the frequency range. The vibrator is driven by a 10kVA amplifier and controlled by an oscillator giving 132 ranges of sweep frequency. The vector thrust of the unit is 1950lb and the bandwidth capacity from 5c/s to 7kc/s.

In the design of the vibrator, provision has been made for mounting a climatic cabinet on to the vibrator table so that altitudes of up to 200 000ft with a temperature range of −85°C to +200°C can be simulated. The climatic cabinet has a built-in lifting arrangement and can be fitted to the table in either its horizontal or its vertical position. Components of up to 80lb in weight can be accommodated.

A reinforced concrete block of 125ft^3 isolated from the floor by a 2in thick insulating pad was provided for mounting the vibrator and it is housed in a separate room adjacent to the laboratory which is comprehensively equipped with other climatic and environmental testing equipment. The vibrator can be made available for testing other companies' products by negotiation.

A PORTABLE ELECTROCARDIO TRANSMITTER AND RECEIVER



By A. R. Atkins* and
E. B. Lock*

A portable electrocardio transmitter is described which is used in physiological investigations under conditions of high humidity and temperature, such as a climatic chamber or underground in mines, and which is used to detect and transmit to a receiver at a distance signals of electrocardio pulses of a working subject. The instrument is small and is attached to a belt worn by the subject. A transducer has been designed for use on subjects who are wet with perspiration.
The receiver is designed so that one of four transmitters can be selected to enable the observer to count the heart-beat rates of subjects by listening to the signals by means of headphones or by connecting the receiver to an electrocardiograph and making a recording.

(Voir page 847 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 854)

IN most types of physiological experiments the heart-beat rate is a value which is a necessary item of the data collected. The heart sound pulses are counted usually by means of a stethoscope or by means of an electrocardiograph, the leads of which are attached to the subject's chest. Where a working subject is involved, both these methods are unsatisfactory. If a stethoscope is used, the subject must stop temporarily to permit the heart-beat rate to be counted. If an electrocardiograph is used, the leads are a hindrance to the subject.

The physiological research carried out at the Human Sciences Laboratory of the Transvaal and Orange Free State Chamber of Mines involves a large number of observations on subjects working under humid conditions in a climatic chamber or underground and the equipment described here was designed so that heart-beat rates could be investigated under these conditions without the inconvenience caused by the use of a stethoscope or electrocardiograph.

The Heart-Beat Signal Transmitter

The main requirements of the transmitter for use under the conditions mentioned were that

(a) It should be small and light.

(b) It should be capable of operating at temperatures in the range of 0°C to 45°C.

(c) It should be operated by a battery with a life of at least five hours.

(d) It should be crystal controlled so that any confusion of signals from different transmitters selected by the receiver would be obviated.

(e) At least four transmitters could be used simultaneously, without confusion of signals, with a distance of fifty yards between transmitters and receivers.

(f) The e.c.g. amplifier should have a high input impedance (1MΩ) and be sensitive to a 100μV signal of frequency varying between 5c/s and 20c/s.

The Citizen band at 27Mc/s was chosen for operating as this provides a good ground range when a reasonably short aerial is used. The output power is only 30mW.

The transmitter circuit is shown in Fig. 1. The e.c.g. amplifier consists of transistors VT_1 to VT_5. The first two emitter-followers provide an input impedance of at least 2MΩ. Because this amplifier is isolated from earth a differential amplifier is not required. The input r.f. choke prevents any radio frequency from entering through the e.c.g. leads and distorting the input.

* *Chamber of Mines and Research Laboratories, S. Africa.*

The third and fourth transistors are voltage amplifiers and the fifth an emitter-follower output. There are only two coupling capacitors. The input $0{\cdot}1\mu$F capacitor must have a low value of leakage current and the 30μF capacitor isolates the amplifier VT_3, VT_4, VT_5, which is directly coupled. The high frequency response is limited by the 2μF and the $0{\cdot}5\mu$F shunt capacitors. Because only the heart-rate signal is of concern, the frequency response of the amplifier is best limited to the range 5 to 20c/s.

The e.c.g. pulse frequency modulates an audio tone in preference to direct modulation of the r.f. oscillator. By listening to the change in modulated tone from the receiver the pulse rate signal can be recognized and counted even when there is rather heavy e.c.g. and radio interference.

The transistors VT_6 and VT_7 comprise a multivibrator, the frequency of which is determined by their common base potential which is connected directly to the output of the amplifier. The normal frequency of the multivibrator is about 800c/s with a deviataion of 50 per cent for an input pulse of 0·25mV.

Transistor VT_8 amplitude modulates the r.f. oscillator by about 40 per cent. Some difficulty was experienced in amplitude modulating the crystal oscillator and when the modulation exceeded 80 per cent the oscillator stopped functioning. An additional r.f. power amplifier, however, would have required more space and greater power consumption.

The current for the oscillator is 6mA and for the remainder of the circuit 5mA. The small 9V type PM-3 dry-cell used has a life of six to eight hours.

The Heart-Beat Signal Receiver

The circuit for the beat-beat receiver, shown in Fig. 2, has a conventional superheterodyne receiver with crystal oscillator. An output plug provides a means for listening to the modulated tone pulses by means of a pair of headphones. An electrocardiograph may be connected to the receiver, for recording the heart-beat signals.

The transistors VT_7, VT_8 and VT_9 form the demodulator. The limiting amplifier VT_7 produces sharp

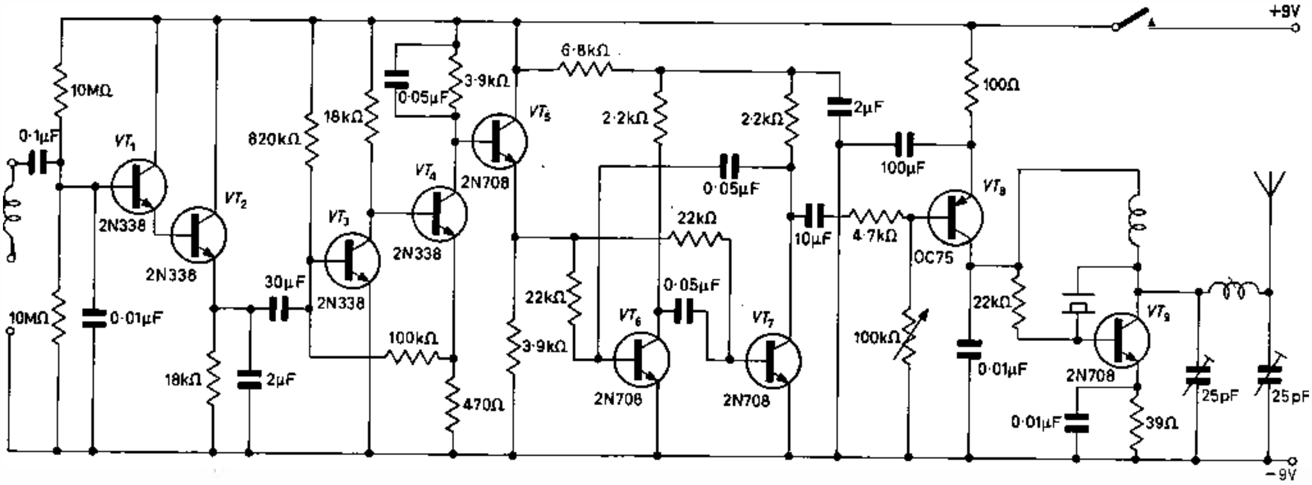
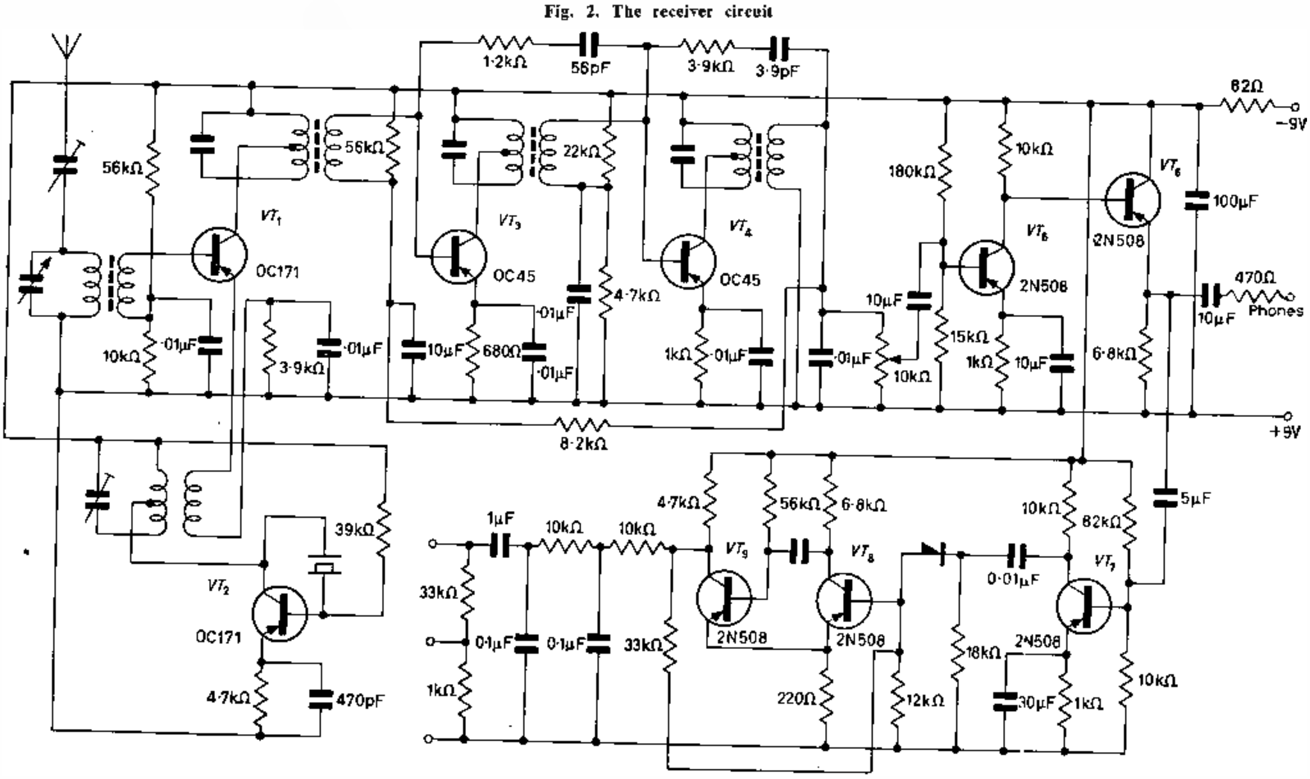


Fig. 1. The transmitter circuit

Fig. 2. The receiver circuit



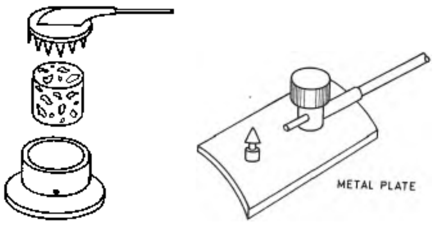


Fig. 3. The metal plate and cup electrodes

square waves which are differentiated and used to trigger the monostable multivibrator circuit VT_8 and VT_9. The circuit produces pulses of constant height and width, the frequency being the same as that of the tone. If this is filtered, the mean low frequency corresponds to that of the cardiopulses. Outputs are provided for recorder and e.c.g. The attenuated output is for the e.c.g.

The Cardio-Electrodes

Ideally, a cardo-electrode must be small and light, it must maintain good contact continuously with the skin without any induced e.m.f. such as polarization being produced, and it must adhere to the skin irrespective of the presence of perspiration or body movement. The electrode must also be inert to chemical action, must not irritate the skin and be easy to apply.

The plate electrode normally used with an e.c.g. makes only partial contact with the skin, especially when the electrode jelly is dissolved away by perspiration; and heavy breathing and arm movement by the subject moves the belt supporting the electrodes causing severe interference.

Three important requirements for the electrodes of this type are therefore, that

(a) The electrodes must be mounted individually on the skin.

(b) There must be no direct contact between the metal of the electrode and the skin.

(c) The electrode jelly should remain effective for at least four hours and should not dry out nor be leached away by perspiration.

The metal plate electrode and the preferred cup electrode are illustrated in Fig. 3. The cup-shaped electrode is made of Polythene and has a base diameter of 25mm and upper diameter of 12mm.

Application of Electrode to Skin

The skin is wiped clean of surface fat. It should not, however, be rubbed too much with a solvent because this dries the skin resulting in a poor electrical contact. Rubber-base cement is applied to the roughened face of the electrode and allowed to become tacky after which the face of the electrode is pressed on to the skin. The cup may be filled with a thick paste or, preferably, with a sponge slightly larger than the cup and saturated with electrode paste. A stainless metal spring-cap, to which the lead is attached, is then pressed into the top of the cup. The electrode lead should be light and flexible so that there is no pull on the electrode. The spring teeth of the cup provide a good contact with the electrode paste. Three small holes at the base of the cup allow the air to escape when the paste is applied to the electrode.

Experience has shown that this electrode has adhered to the skin for three hours even when the subject perspired freely. The entire electrode may be covered with a square of adhesive tape as a safeguard against it being pulled away from the skin.

It has been found that the electrode may be removed from the skin by means of a gentle pull, but a solvent may be used, if necessary. The thin layer of cement is easily peeled from the face of the cup, at the end of the experiment.

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Photo-electric Monitoring For Car Washing

Photo-electric monitoring equipment has been supplied by -DEP (MI Group) of Rugeley, Staffs, to The 5 Minute Car Wash Service at Wolverhampton.

The equipment, designed to count .nd record the number of vehicles)eing washed, consists of modulated)hoto-electric units with infra-red)eam, strip chart recorder, process imer and read-out counter with date tamp.

As each vehicle breaks the beam, he process timer is initiated causing . count to be stored within the equipment. At two-hourly intervals, the otal stored number is printed on the ead-out counter.

The position of the beam is nonitored by the strip chart recorder.

Should any unauthorized entry to he control cubicles be made, or interference with the beam take place, a eflexion of the recorder pointer ccurs on the chart, thereby presentng full details of the time, duration nd nature of the interference.

Although the recorder continuously nonitors the beam, the read-out ounter is inhibited for a pre-set eriod. By this means only the number of vehicles breaking the beam is detected and not the people engaged in cleaning.

At midnight the read-out counter automatically prints the cumulative total for the previous 24 hours, changes the date and re-sets to zero.

The car washing bay. The photo-electric equipment is shown in the foreground on either side of the bay

Varistor Networks as Analogue Computing Elements

By W. G. P. Lamb*, Ph.D. and Margaret L. Stones*, B.Sc.

Simple networks of non-linear and linear resistors designed to produce an element which accurately obeys a current-potential relation $i = Kv^n$ where n lies between one and three are described. An application relating to the diffusion of hydrogen ions in iron is outlined.

(Voir page 847 pour le résumé en français: Zusammenfassung in deutscher Sprache aut Seite 854)

IN analogue computing methods, the requirement for a unit which accurately obeys the current-potential relation $i = Kv^n$ has often been met by the use of a diode function generator employing a fairly large number of diodes and associated resistors. However, the relative complexity of such arrangements has led to some use being made of silicon carbide varistors as non-linear elements; in particular, Kovach and Comley[1] and Brown and Walker[2] have made use of a varistor in series with an ordinary ohmic or linear resistor in order to produce an element with n equal to about 2. Brown and Walker also corrected for some of the residual error by the use of a subsidiary network employing a further varistor and an operational amplifier circuit.

In addition Cluley[3] has explored the possibilities of a cadmium sulphide semiconductor element in order to produce an accurate square law element. This unit, however, operates under bias conditions and is limited to a somewhat restricted working range.

The authors of the present article were interested in a varistor unit which would be accurate to within a few per cent, have a large working range and would not require active elements such as operational amplifiers. Such units were required for use in the construction of networks for the solution of certain non-linear differential equations.

Correction by Use of a Single Series Resistor

Silicon carbide varistors obey a current potential relation $i = Kv^m$ where m varies gradually from unity to four as v is increased from zero over the working range, and it is found empirically that a relation of the type given by equation (1) can be used to describe the behaviour over a large part of the range. This equation is:

$$\log i = a + b \log V + c \log^2 V \quad \text{(1)}$$

For the simple combination of a varistor and a resistor R connected in series, the total potential E across the combination is given by:

$$E = V + iR \quad \text{(2)}$$

where V is the potential across the varistor when current i flows through it.

Fig. 1 shows a plot of log E versus log i. At very low values of current the characteristic has unity slope; this decreases with increasing current but returns to unity slope at large current values.

The error ϵ between the actual and the required characteristic is given by:

$$\epsilon = V + iR - (i/K)^{1/n} \quad \text{(3)}$$

where $i = Kv^n$ is the required characteristic, n being constant and independent of v.

An analytical analysis of the way in which this error varies with R etc. is difficult to undertake. However, from the nature of the partial derivative $\partial\epsilon/\partial R = i$ it can be seen there is no best value of R for minimizing ϵ. In fact, as R is increased the values of ϵ decrease as the three points A, B and C at which it is zero tend to coalesce. Thus for large values of R a close fit to a required power law may be obtained at the expense of working range, while for small values of R a large working range is available but with relatively large deviations or errors.

* Sir John Cass College, London.

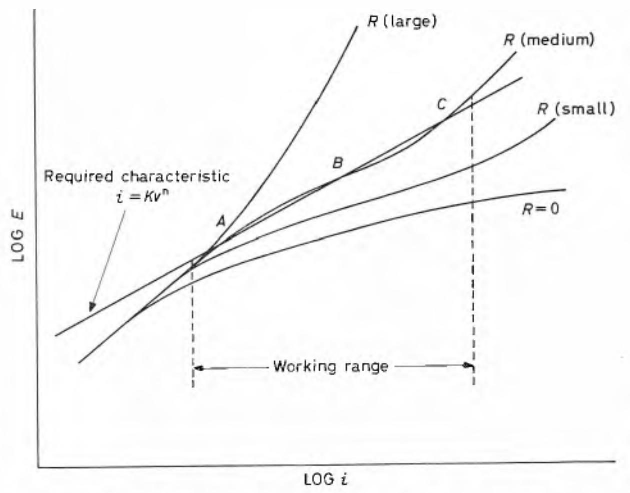


Fig. 1. Characteristics for a SiC varistor in series with an ohmic resistor R

Upper-end Correction

The working range consistent with a reasonable degree of error may be extended to large values of current by choosing a fairly large value of R, so that the intersection points A, B and C are brought quite close together, and then connecting an additional varistor across R. This additional varistor stops the return to unity slope above C and produces a further intersection at D as shown in Fig. 2. If now, a resistor R' is placed in series with this second varistor, yet another intersection is produced. It is possible to continue the addition of shunts consisting of varistors and resistors until the desired length of characteristic is made available. An accuracy of about ± 2 per cent can usually be produced by the use of only three varistors and their associated resistors. "Morganite" silicon carbide varistors type Y876/2 were used together with carbon "Hystab" resistors.

Specimen characteristics were constructed for $n = 3/2$, 2 and 3 with a working range of about 100V. A maximum fractional error of about 2 per cent within the working range was obtained. This was defined as $\Delta E/E$, where ΔE is the error at the value E.

A mathematical formulation of the behaviour of these circuits is rather awkward, but the practical determination of the characteristics is easily carried out with the aid of

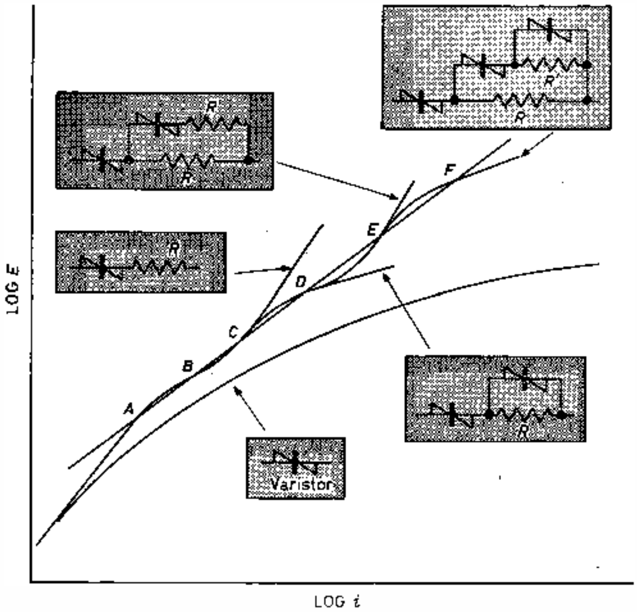


Fig. 2. Behaviour of varistors with different padding circuits

a suitable power unit and other simple laboratory instruments.

As an alternative, the behaviour of the squaring circuits was studied with the aid of the operational amplifier set-up shown in Fig. 3. A repetitive square wave was integrated to provide a sawtoothed wave which was then used to drive the varistor network under examination. The sawtoothed wave was integrated and the square-law wave so produced compared with the output of the varistor unit by a non-linear difference amplifier. This error signal was then displayed on a cathode-ray oscilloscope. Typical resistor values for the $n = 2$ case were $R = 5\,\text{k}\Omega$, $R' = 2{\cdot}2\text{k}\Omega$ and $R'' = 1{\cdot}2\text{k}\Omega$.

Lower-end Correction

For currents of less than about 0·1mA the behaviour of the varistor type Y876/2 tends to become ohmic as the current is further reduced. However, a considerable degree of correction to the lower end of the characteristic may be obtained by connecting pairs of semiconductor diodes in series with the main element and with a resistor across them. It was found that germanium diodes type OA81 in series with silicon diodes type FS10 provided useful correction for the $n = 3/2$ case, while the FS10 was used for $n = 3$. When $n = 2$, it was found necessary to use two varistors in series as the main element together with diodes OA81. Fig. 4 shows the circuit arrangement of the diodes for lower end correction, together with the upper end padding circuits.

The addition of the semiconductor diodes permitted extension of the lower limit of the working range from about 0·1mA down to about 0·0001mA with about the same fractional error as that present over the main part of the characteristic.

It was found that both types of correction could be employed together; the working range for the element for which $n = 3/2$ then extended from 0·0001 to 100mA for example.

The maximum variation of K with temperature for the completed units was found to be close to that of a single varistor namely 0·2 per cent/°C.

Application to the Solution of a Non-linear Diffusion Equation

Apart from function generation, multiplication etc., the availability of a passive and reasonably accurate non-linear element which will operate over a wide working range is of considerable assistance in extending the use of resist-

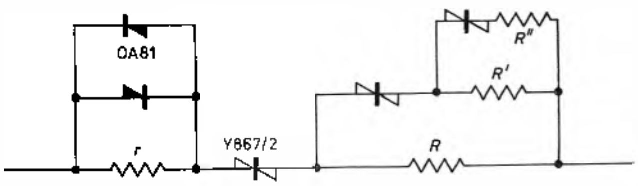


Fig. 4. Circuit arrangements for lower and upper correction

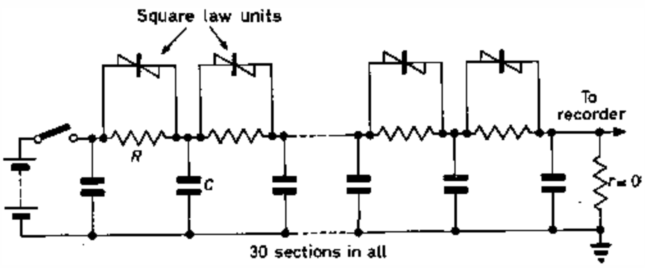


Fig. 5. Non-linear RC network used for studying diffusion of H+ in Armco iron

ance-capacitance networks to provide solutions of certain non-linear partial differential equations.

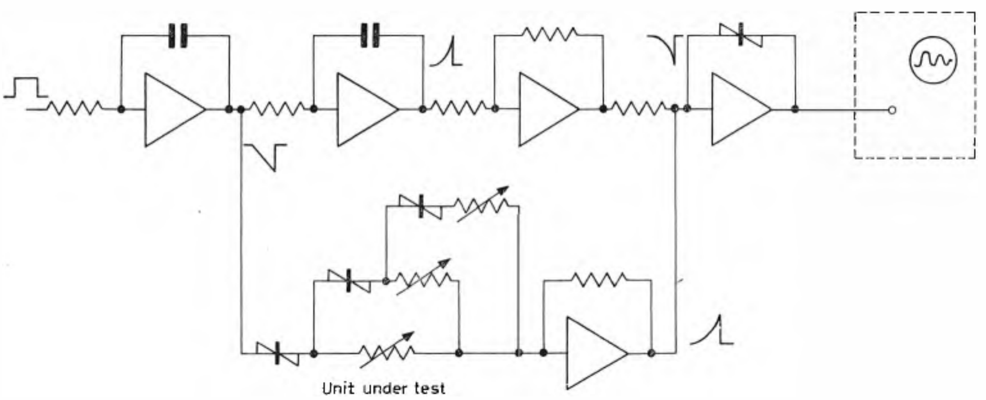
The authors were particularly interested in solutions of the non-linear diffusion equation:

$$\partial G/\partial t = \partial/\partial x\ (D\partial G/\partial x) \text{ where } D = a - (b\partial G/\partial x)$$

where G is the concentration at position x and time t and the diffusion coefficient D is a function of $\partial G/\partial x$, a and b being constants.

The equation is relevant to the diffusion of hydrogen ions in Armco iron. Usually, when D is not constant it is acceptable to take D as dependant on the concentration, writing for example $D = a + bG$. However, from experiments in which the steady state flow was measured for specimens of different thicknesses the authors have shown that, in this case, D appears to be a function of the concentration gradient rather than concentration.

Fig. 3. Circuit for display of error characteristics



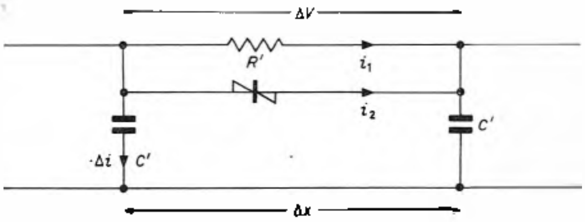


Fig. 6. Network element

Steady state values do not permit deduction of the values of a and b, but by studying the transient behaviour of a network analogue, and making comparisons with experimental results it was possible to calculate a and b, and to confirm that the dependance of D on concentration gradient gave the correct transient behaviour as well as the observed steady state flow.

The network used is shown in Fig. 5 and consisted of square law and ordinary resistors in parallel together with capacitors. Some thirty sections were employed. It can readily be shown that the network obeys the required differential equation.

Acknowledgment

Acknowledgment is due to T. Radnakrishnon of the Metallurgy Department of Sir John Cass College who carried out the experimental work on the diffusion of hydrogen ions through specimens of iron.

Appendix

Let R' and C' be representative components of the network and let them be associated with a length Δx as shown in the diagram, Fig. 6.

If for the square law resistor $i = Kv^2$, then

$$-\Delta V = K^{-\frac{1}{2}} i_2^{\frac{1}{2}} = i_1 R'$$

and

$$-(\Delta V/\Delta x) = (K^{-\frac{1}{2}}/\Delta x)\, i_2^{\frac{1}{2}} = (i_1 R'/\Delta x)$$

In the limit

$$\lim_{\Delta x \to 0} -(\partial V/\partial x) = i_1 R = k i_2^{\frac{1}{2}}$$

where $R = R'/\Delta x$ and $k = K^{-\frac{1}{2}}/\Delta x$

Similarly

$$-(\partial i/\partial x) = C\,(\partial V/\partial t)$$

where $C = C'/\Delta x$

And as $i = i_1 + i_2$

$$-(\partial i_1/\partial x + \partial i_2/\partial x) = C \partial V/\partial t$$

But $i_2 = (i_1 R/k)^2$ so $\partial i_2/\partial x = (2R^2/k^2)\,(i_1 \partial i_1/\partial x)$

Thus, $-(C\partial V/\partial t) = \partial i_1/\partial x\,(1 + (2R^2 i_1/k^2))$

Now, $i_1 = -(1/R)\,\partial V/\partial x$ and $\partial i_1/\partial x = -(1/R)\,\dfrac{\partial^2 V}{\partial x^2}$

So:

$$-(C\partial V/\partial t) = -(1/R)\,\partial^2 V/\partial x^2)(1 - (2R^2/k^2)\,.\,(1/R)\,(\partial V/\partial x))$$

or $\partial V/\partial t = \partial^2 V/\partial x^2\,(1/CR - (2/k^2 C)\,(\partial V/\partial x))$

It will be seen that the required equation is of this form since by substituting for D, then:

$$\partial G/\partial t = \partial/\partial x\,[(a - (b\partial G/\partial x))\,\partial G/\partial x]$$

which gives:

$$\partial G/\partial t = \partial^2 G/\partial x^2\,(a - (2b\partial G/\partial x))$$

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A Solid State A.D.F.

The first airborne automatic direction finder in the world which conforms fully with the latest airline specification, (ARINC 550) and also provides completely automatic, solid state, crystal controlled tuning, has been developed by The Marconi Co. Ltd.

This new equipment, type AD370, provides pilots with the fastest possible means of finding the bearing of any given radio station. The price is competitive with current American automatic direction finders, and the equipment will fit directly into all the latest American aircraft types, and also into the BAC 111. Its small size, light weight and ease of operation, also make it suitable for use in all types of military and civil aircraft, including helicopters and high performance fighters.

The new equipment is completely transistorized, and there are no moving parts in either the tuning stages or the range selection. It is believed to be the only a.d.f. available in the world in which mechanical switching and variable capacitors have been completely eliminated. For this reason alone, the AD370 should prove in service to be more reliable than any other automatic direction finder which is currently available.

The accuracy and stability of the frequency selection is ensured by only 13 crystals, and the need for fine tuning is completely eliminated. There is, in fact, no provision on the control unit for frequency trimming. In operation, the required frequency is set up on the control unit, using easily read decade dials. Tuning is then performed automatically and instantaneously by the solid state electronic system. In this way, the bearing of the selected station is provided easily, quickly and reliably.

The AD370 is a completely solid state equipment. Varactor diodes are used in the tuning of the i.f. and r.f. stages, and semiconductor diodes provide all range and crystal switching functions.

Moving parts have been completely eliminated from the receiver, with the exception of the goniometer and servo motor, which replace the rotating loop aerial in earlier direction finders, and which cannot be eliminated in a practical a.d.f. The associated synchro transmitter has also been retained to transmit the bearing information to the pilot's indicator. Two relays, used to switch the aerials when the equipment is not being used for automatic direction finding, are the only other moving parts in the receiver.

The circuits are contained in a short ½ ATR case, with a 'dog-house' on the front panel, containing the goniometer, servo motor, synchro transmitter and associated circuits.

Circuit components are mounted on tag boards, which are screwed to both sides of a central partition running the length of the receiver unit. Internal plugs and sockets have been eliminated, and direct soldered connexions are used to interconnect the tag boards. This design gives easy access to all components and enables interconnexions to be kept to a minimum.

The tuning system employs three mixers, with crystal oscillator inputs, which convert the r.f. signals to an intermediate frequency in the band 130·5 to 180kc/s. The i.f. unit is tuned accurately over this narrow band by varactor diodes. With this system, all the advantages of crystal control are obtained using only 13 crystals, instead of the usual complement of 36.

A feature of the AD370 is the use of three separate detectors, for servo amplifier, audio amplifier, and a.g.c. circuits. This permits the use of a beat frequency oscillator, without affecting the a.d.f. performance.

The decade frequency selector on the control unit, also conforms fully with ARINC Characteristic 550. The selected frequency is displayed in clear numerals, ½in high, and integral panel lighting is provided. The frequency controls select resistance networks which provide the voltages to control the varactor diodes in the r.f. and i.f. tuning stages.

The operating frequency is selected on an easily read decade dial, using three selector controls—for hundreds, tens and units and halves of kilocycles per second. The frequency scale is continuous from 190 to 1 799·5kc/s, in steps of 0·5kc/s.

In normal a.d.f. operation, the bearing of the selected station will be shown immediately on the bearing indicator. Alternatively, direction finding can be accomplished manually, with the function switch turned to the LOOP position. The sense aerial is then disconnected, and the search coil of the goniometer may be rotated by means of a switch on the controller. The null position is determined by listening to the audio output of the receiver. This facility is normally only used in conditions of abnormal static. The loop rotating switch may be used, however, in normal operation, to move the bearing indicator from the indicated reading, to check that it returns to the same position when the loop switch is released. This confirms the operation of the automatic direction finder.

Transistor Astable Multivibrators for General Purpose Logic Elements

By G. H. Stearman*, B.Sc., A.M.I.E.E.

The desirable features of an astable multivibrator circuit for use as a clock generator with a general purpose set of logic elements are first discussed. The shortcomings of the conventional circuit, of several variants upon it and of some unconventional circuits are pointed out, after which some new variants upon the basic configuration are introduced and shown to meet the specified requirements more closely than previously existing circuits.

(Voir page 847 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 854)

In the course of designing a set of general purpose solid state logic elements for use in a variety of digital control and indication applications, one problem encountered was the selection of a suitable astable multivibrator for use as a clock generator. During a similar, previous exercise a conventional circuit, as shown in Fig. 1, had been adopted but had proved to be not entirely satisfactory for several reasons. The author therefore began looking for alternative circuits more closely compatible with the particular requirements, which are discussed in the next section, and this article describes the results of the search.

The Requirements

The characteristics of an astable multivibrator likely to be of interest are discussed here together with their relative importance to the application in mind.

Rise and Fall Times

The general shape of the output wave of the circuit of Fig. 1 is well known, the positive-going edge being appreciably faster than the negative-going. In the set of elements in question only positive-going edges were to be used for triggering bistable and monostable circuits so that the slow negative-going edge was not a serious disadvantage.

Voltage Levels

The basic logic element of the set was the diode-saturated transistor NAND circuit of Fig. 2 whose binary states are represented by the voltage levels shown. It is clearly desirable that the astable circuit should produce outputs having the same levels, and in particular that one level should be very close to zero. The basic circuit of Fig. 1 fulfils this requirement very well.

Self-starting

The circuit of Fig. 1 is not inherently self-starting if both transistors are heavily saturated because the loop gain with both transistors conducting may be much less than unity. It is not necessarily sufficient, to ensure starting, to bring one stage out of saturation since the gain of the other may still be so small as to keep the loop gain less than unity; generally both stages must be brought out of saturation. This problem is severe when the circuit is to be used to drive a number of resistive loads connected between collector and V_{cc}, which was so in this instance, and where a worst-case design procedure has been adopted to ensure static stability, as was also true here. In these circumstances the transistors of an unloaded astable may well be oversaturated by a factor of 10 or more, making starting a very real problem.

Oscillator Period

Because resistors R_t must be small enough to ensure that the base current of the conducting transistor is sufficient to give the required degree of saturation, the time-constant C_tR_t can be made large only by increasing C_t. Periods of a few milliseconds are readily obtained with moderate values of C_t but periods of the order of seconds are inconvenient to produce. It is not difficult to overcome this problem with additional components, as discussed later, but very long periods were not required for the application in mind.

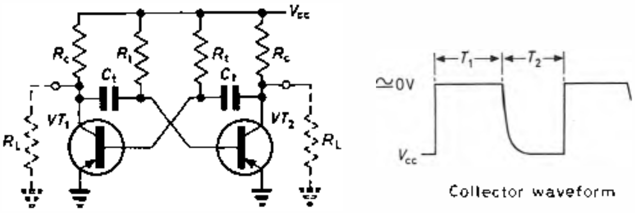

Fig. 1. Basic conventional astable multivibrator

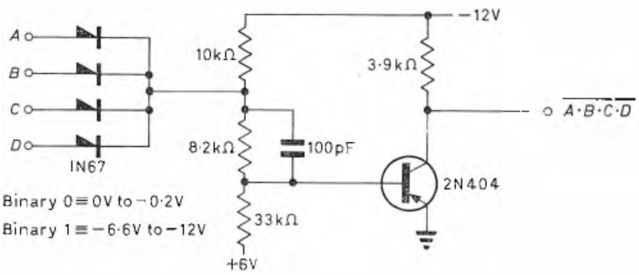

Fig. 2. Diode-transistor NAND circuit

Mark-to-space Ratio

In general, high mark-to-space ratios are not required for use in logic networks but certain special applications may demand them. The basic circuit is somewhat limited in this respect, as is well known, because if different value timing capacitors are used the larger capacitor has to be recharged during the shorter period. Variation of resistors R_t is not feasible due to the requirement of saturation already mentioned.

Stability of Period

The astable multivibrator is at best a relatively poor device from this point of view and, as correctly observed by Hurst[1], it is only in certain special applications (e.g. frequency division) that it makes sense to synchronize the circuit to a crystal or other highly stable oscillator. Nevertheless with care in design the frequency may be made adequately stable for many purposes. Basically each half-period of the circuit of Fig. 1 is a function of the following parameters (see Fig. 3).

(a) The amplitude of the collector waveform, which determines V_o.

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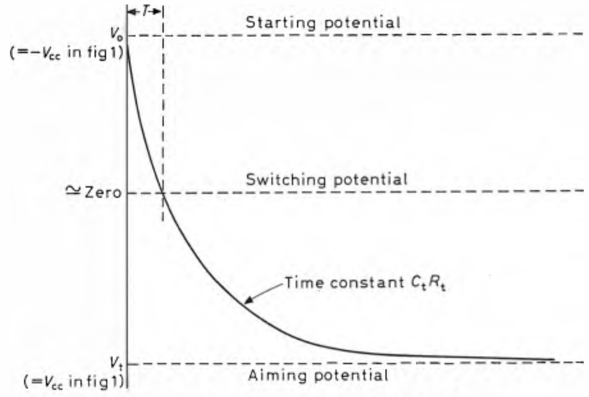
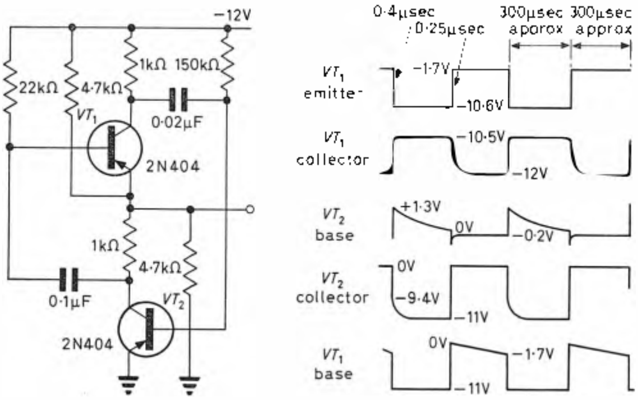


Fig. 3. Timing wave of the basic circuit

Fig. 4. A serial circuit (Smith)



(b) The aiming potential V_t, in this case $= V_{cc}$.

(c) The time-constant C_tR_t.

(d) To a second order, the base-emitter potential at which conduction commences.

The parameter in (d) is generally very small relative to V_o and V_{cc} and may be assumed to be zero in practice. The aiming potential can be varied, as described by Hurst, to give frequency control, but is usually fixed at V_{cc} for convenience and economy of supply lines. The potential V_o, for an unloaded circuit is equal to $-V_{cc}$, since this is then the collector waveform amplitude; the half-period is given by the equation:

$$T = RC \log n \left[\frac{|V_o| + |V_{cc}|}{|V_{cc}|} \right] \quad \ldots\ldots\ldots\ldots\ldots \quad (1)$$

Putting $|V_o| = |V_{cc}|$ it is found that $T = RC \log n\, 2$ and is thus independent of V_{cc}. However, a set of general purpose logic elements will frequently contain one or more circuits likely to impose upon the multivibrator outputs resistive loads from collector to ground, as shown dotted in Fig. 1, causing the negative voltage level to be reduced to $V_{cc}R_L/(R_L + R_C)$.

Potential V_o in Fig. 3 is reduced to this same value, giving rise to a reduced half-period (which is still nearly independent of V_{cc}). The effect is quite significant and means of minimizing it are worth consideration.

Complementary Outputs

In logic systems it is very convenient to have available two complementary outputs, a feature provided by the basic circuit.

Resistive and Capacitive Loads

One effect of resistive loads to ground has just been pointed out. The reduction of waveform negative voltage level is not usually of great significance in itself because the elements in the set have to be designed to operate satisfactorily over a fairly wide range of levels. The element of Fig. 2, for example, operates with negative (binary 1) inputs down to about 55 per cent of the unloaded value (V_{cc}).

Resistive loads to V_{cc} are in parallel with the collector load resistor and partly account for the degree of oversaturation required in a conducting transistor but otherwise the only effect is to improve the negative-going edge of the waveform by reducing the timing capacitor recharge time-constant.

Capacitive loads do not seriously affect the positive-going edge (usually the more important) but deteriorate the negative-going edge appreciably.

Alternative Circuits

From the above discussion it is clear that the basic circuit is to a large extent suitable for the purpose in mind but suffers from some faults, primarily dependence of the period upon resistive loading to ground and difficult starting. Before considering what modifications might be applied to remove these faults the literature was scanned to determine what alternative circuits were available which might more closely satisfy the various requirements. These alternatives are briefly discussed here.

The literature on astable multivibrators contains descriptions of several unconventional designs and a number of variations upon the basic circuit of Fig. 1. Most authors were concerned mainly with improvement of the output waveforms although some also discussed the achievement of high mark-to-space ratios and of high ratios of period control. Fig. 4 is a circuit due to Smith[2], the waveforms shown, like all those given in this article, having been measured by the present author.

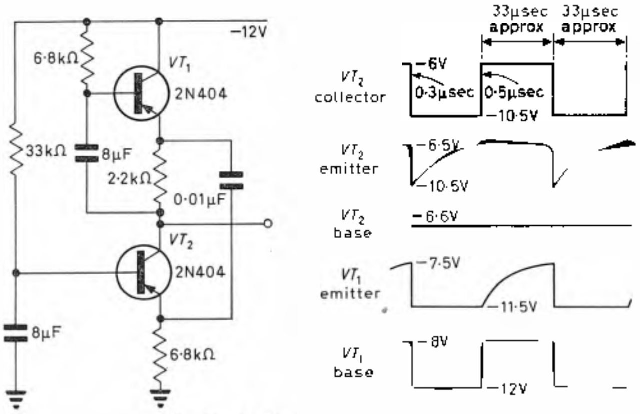
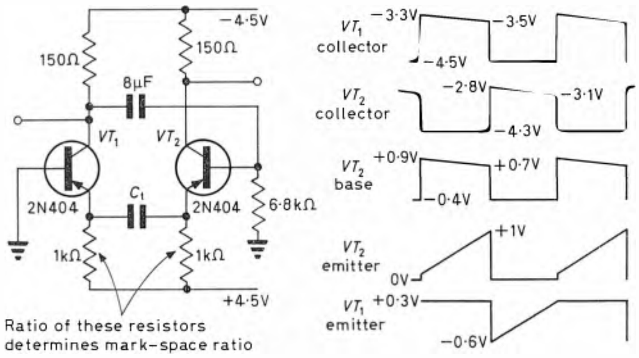


Fig. 5. Another serial circuit (Ristic)

Fig. 6. An emitter-coupled circuit (Johnson)



The circuit is a serially switched version of the basic circuit and its waveforms are easily predictable. The output waveform is well-shaped and does not contain exponentials due to the recharging of timing capacitors. However, it is unsuitable for the present purpose since there is one output only, neither of whose voltage levels is nominally at zero and its period is dependent upon resistive loads to ground or to V_{cc}. In addition the period is strongly dependent upon V_{cc} itself. The circuit is self-starting normally but if stopped artificially (e.g. by grounding the output) will not restart automatically. (This ability to restart will be referred to hereafter as self-restarting).

An ingenious variant of the serial connexion is described by Ristic[3] and is shown with its waveforms in Fig. 5. This circuit has many unusual features; like Smith's circuit it has one output only and neither level is at zero, but unlike the latter the voltage levels and the half-periods obtained

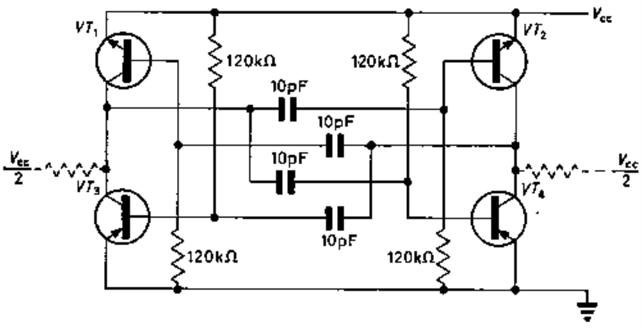


Fig. 7. A complementary circuit (Jovanovic)

can be predicted only by a tedious iterative process. Ristic does not make this clear in his article but reference to his analysis will show that this is so. The output waveform is good although the switching action on the positive-going edge is cumulative over a small portion (0·1V) only of the swing so that this edge has a relatively long rise-time. One timing capacitor only is used, supply current is drawn during one half-period only (for the mode of operation see Ristic's article) and the circuit is self starting and restarting. However, resistive loads to V_{cc} strongly affect both period and amplitude, while loads to ground alter the mark-to-space ratio but not the amplitude or period. For these and the other reasons mentioned already the circuit is unsuitable for the present purposes.

Johnson[4] describes an emitter-coupled circuit (Fig. 6) whose main advantage is the high mark-to-space ratios obtainable. In most other respects it is unsuitable since neither transistor saturates and neither voltage level is at zero. The rise and fall times are satisfactory but both outputs suffer from a distinct sag of the positive level; the circuit is self-starting and restarting.

Another unconventional circuit having many good features is that of Jovanovic[5] shown in Fig. 7. This uses complementary transistors and, being perfectly symmetrical, should give equally good rise and fall times. In addition, provided that the conducting transistors are maintained in saturation the period must be quite independent of resistive loads either to ground or to V_{cc} since the output levels are firmly set at zero and V_{cc}. Capacitive loads can be tolerated and affect both rise and fall time equally. The only serious shortcoming of the circuit is the rather complex cross-coupling network which makes period variation an awkward procedure, but for a fixed period it is all but ideal for the purpose in mind. Unfortunately the author had no npn transistors available at the time and place at which this work was carried

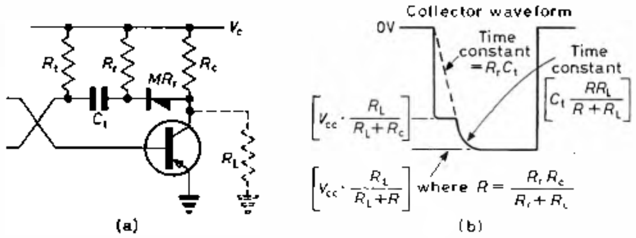


Fig. 8. Use of isolating diodes
(a) Circuit
(b) Effect of resistive load to ground

out so that not only could no details of the waveforms be measured but it was necessary to go on with the search for alternative solutions. Probably this circuit is self-starting and restarting due to its symmetry but this could not be verified.

The remaining literature studied dealt with variations of the basic circuit, the two most important of which are shown in Figs. 8 and 9. The use of an isolating diode (Fig. 8(a)) to improve the negative-going edge was suggested by Rozner[6] and succeeds in removing the recharging exponential from the output waveform. Naturally external capacitive loads will still deteriorate this edge; furthermore, resistive loads to ground still affect the output amplitude and hence the period but in addition give rise to the distorted waveshape shown in Fig. 8(b), since the diode recommences to conduct at a potential $V_{cc} \cdot R_L/(R_L + R_c)$, whereupon the effective collector load becomes (neglecting diode forward resistance) the parallel combination of R_c and R_r so that the output moves (on a time-constant $R_{eff}C_t$) towards a final potential of $\dfrac{V_{cc}R_L}{R_L + \dfrac{R_cR_r}{R_c + R_r}}$ where R_{eff} is the parallel combination of R_r, R_c and R_L.

Although this waveform is an unusual shape this does not necessarily mean that it is unsatisfactory. The pursuit of ideal waveforms for their own sake has less to do with engineering than with aesthetics and provided the essential features of a waveform are present other irregularities can often safely be ignored.

The second important variant of the basic circuit is

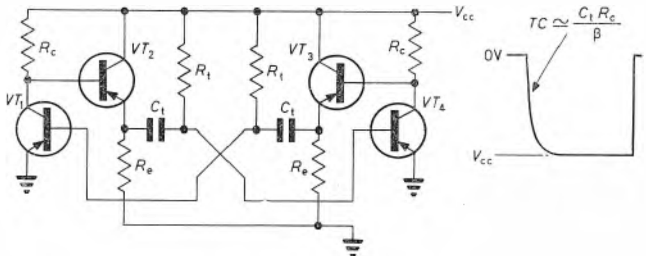
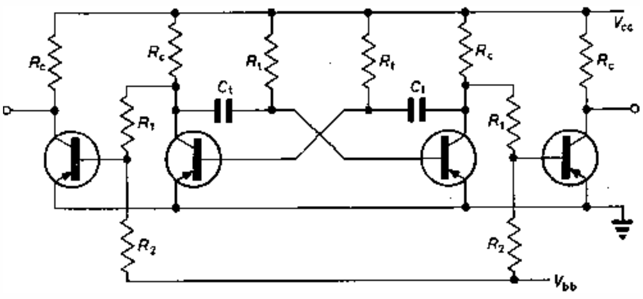


Fig. 9. Use of emitter-followers to recharge C_t

Fig. 10. Use of buffer stages to reduce loading effects



the addition of emitter followers to provide a low-impedance recharging path for the timing capacitors (Fig. 9). This arrangement was described by Rakovic[7], although this may not have been the first publication of the circuit. The capacitive current of the recharge is reflected, at much reduced amplitude, in the output circuit so that an exponential still exists but of much smaller time-constant. Thus, for the same rise time much larger capacitors and hence longer periods may be used. The period is still affected by resistive loads to ground and self starting is not improved.

Some New Variants of the Basic Circuit

After study of the various circuits available it was concluded that the basic circuit possessed most of the desirable features except constancy of period under load and self-starting. It was decided therefore to examine methods of improving these two factors while at the same time attempting to achieve an adequately good waveform.

One obvious method to avoid loading effects is the addition of two transistors as simple buffers (Fig. 10).

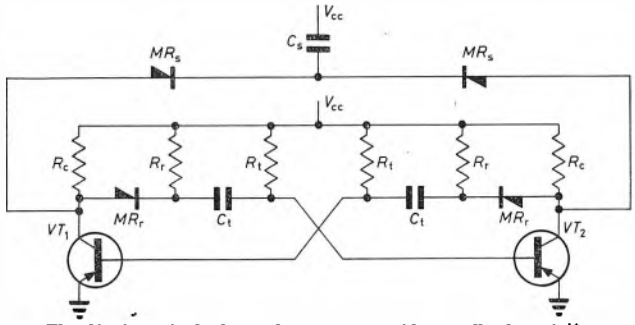


Fig. 11. A method of ensuring start on sudden application of V_{cc}

Although this achieves constancy of load and therefore of period the waveform of the output is not appreciably improved and the recharging exponential of the negative-going edge appears as a deterioration of the positive-going edge of the output, an undesirable feature. Furthermore, since the degree of over-saturation of the multivibrator transistors is largely determined by tolerance of supply lines, resistor values and transistor gain, especially the last, isolation from external loads does not materially improve the self-starting capability of the circuit. If additional transistors are accepted as necessary there are better ways of employing them than as simple buffers.

Attention was initially concentrated upon the achievement of self-starting and an existing circuit for this purpose was examined. The circuit (Fig. 11) uses two diodes and a capacitor additional to the basic configuration plus isolating diodes. When V_{cc} is first applied (suddenly) capacitor C_s begins to charge via both diodes MR_s and both transistors. In effect diodes MR_s are temporarily in parallel with their respective collector resistors so that the degree of saturation is reduced and the loop gain raised sufficiently to permit starting. This arrangement is only partly successful because if the degree of over-saturation is, say, 10 then the diode forward resistance must be less than $R_c/10$, a condition not always easy to satisfy, especially with signal diodes. Furthermore, the method is restricted to a suddenly applied supply and will not function if V_{cc} is increased slowly or if the circuit is artificially stopped (i.e. the circuit is not self-restarting).

Fig. 12 shows a better arrangement of the same three components. Suppose that, after applying V_{cc}, both stages are in saturation so that both collector potentials are at zero. Then no potential is applied to resistors R_t so that

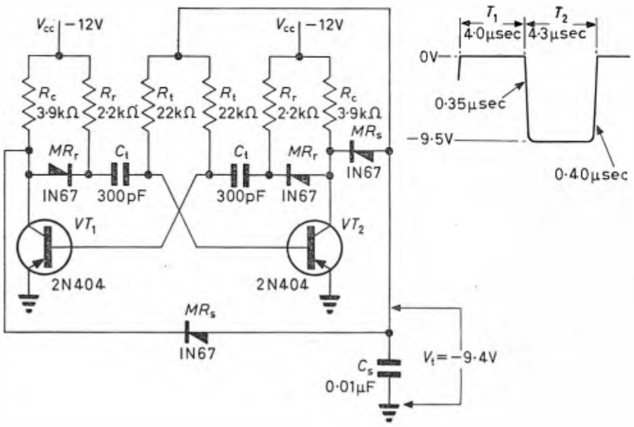
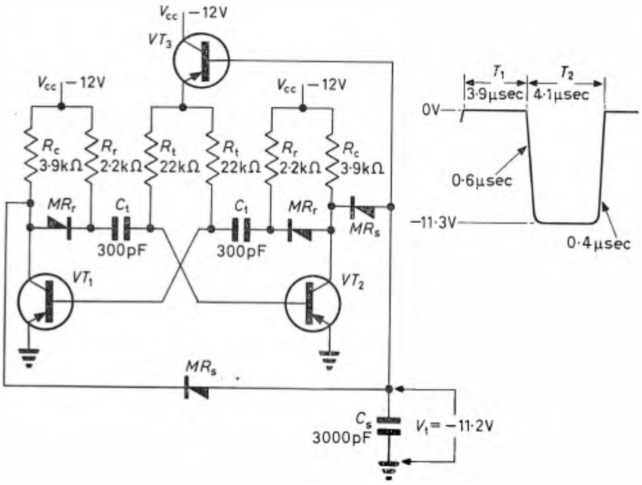


Fig. 12. An improved circuit using the same additional components as that of Fig. 11

the saturated state cannot be maintained. Both collectors therefore tend to go negative until cumulative action commences and the oscillation starts. Once started diodes MR_s with C_s act as a full wave rectifier to generate a steady aiming potential V_t for application to resistors R_t. The output waveform is not deteriorated in respect of rise and fall times although the negative level is reduced by the loading effect of resistors R_t upon the collector circuits. The starting mechanism is not dependent upon time effects and thus is effective when V_{cc} is applied slowly. Furthermore, the circuit is self-restarting because if, for example, one collector is accidentally grounded, any supply for R_t must arise from the other transistor, whose collector potential must be negative and which is therefore out of saturation. On removing the ground the first transistor will also be out of saturation (assuming equal gains) so that the oscillation re-commences. If the gain of the grounded stage is in fact appreciably greater than that of the other it is possible for this stage to be saturated when the ground is removed but in practice the degree of saturation will not generally be sufficient to reduce the loop gain below unity. This would demand a very high gain ratio between the two transistors. This circuit, therefore provides completely reliable starting and restarting under all conditions. The addition of a third transistor as emitter-follower to supply resistors R_t reduces the loading effect appreciably and permits a smaller smoothing capacitor C_s to be used (Fig. 13).

In neither circuit (Fig. 12 or 13) are the isolating diodes

Fig. 13. Addition of an emitter-follower to the circuit of Fig. 12 to reduce loading.



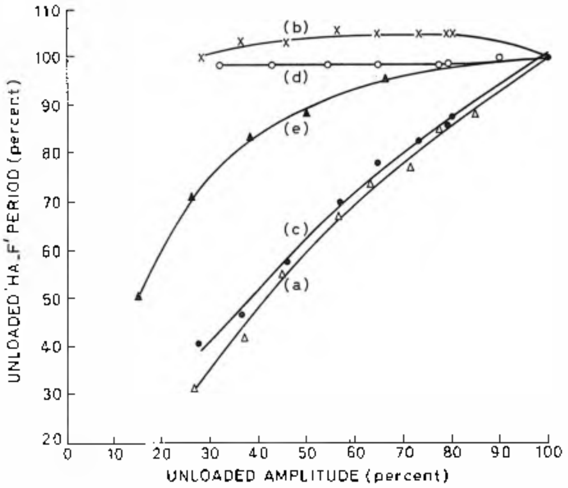


Fig. 14. **Variation of 'Half' period with output amplitude for:**
(a) Basic circuit (balanced load)
(b) Circuit of Fig. 13 (balanced load)
(c) Circuit of Fig. 13 (unbalanced load)
(d) Circuit of Fig. 15 (unbalanced load)
(e) Circuit of Fig. 16 (unbalanced load)

essential and they may be replaced by emitter-followers as in Fig. 9 or dispensed with entirely.

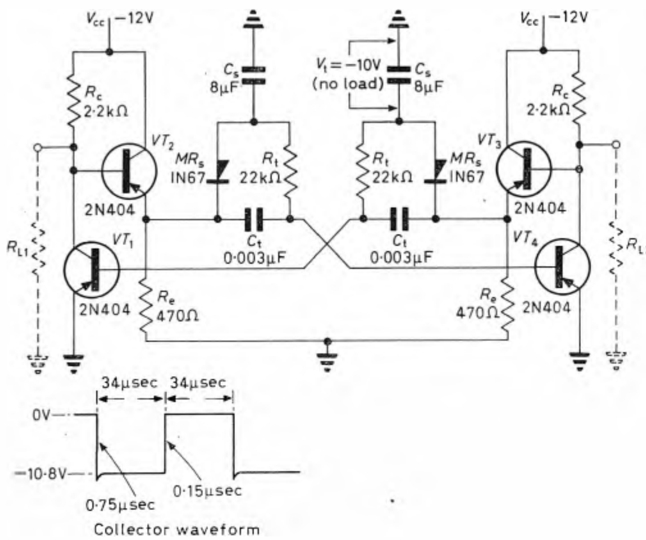
Constant Period Circuits

In the circuits of Figs. 12 and 13 it will be observed that V_t is proportional (in fact, nearly equal) to the negative voltage level of the output signal. Thus the aiming potential V_t may be replaced by kV_o in calculating the 'half'-period* giving:

$$T = RC \log n \cdot \left[\frac{|V_o| + |kV_o|}{|kV_o|} \right] = \text{constant}$$

Thus, provided both outputs have the same amplitude (i.e. the resistive loads to ground are equal) the period is independent of output amplitude. This is illustrated in Fig. 14 where the 'half'-period is plotted against the output amplitude for (a) a basic circuit with isolating diodes and (b) the circuit of Fig. 13, both with balanced loads. In these cases the full period is, of course, double the half-period. Also, curve (c) in Fig. 14 illustrates the change of half-period of the circuit of Fig. 13 when load is applied to one side only.

Fig. 15. **A cross-coupled constant period circuit for unbalanced loads**



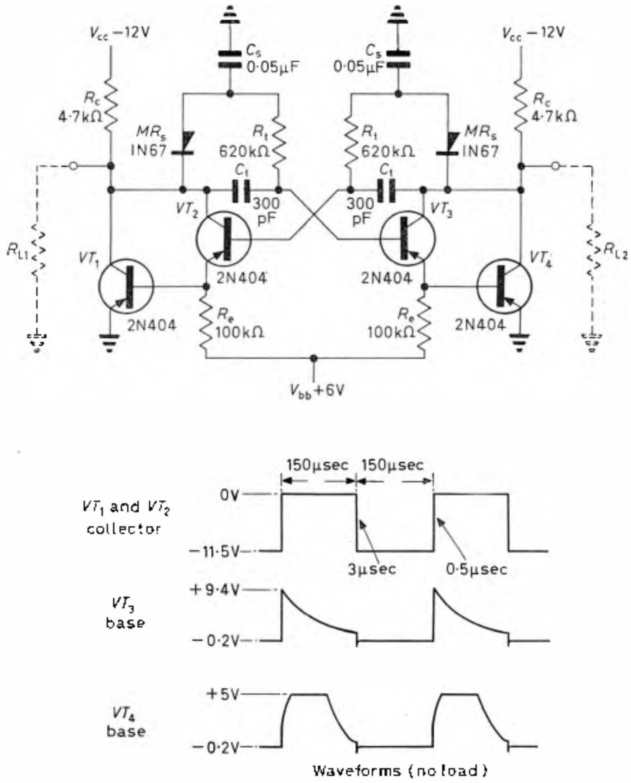


Fig. 16. **Compromise circuit giving optimum benefit from the additional transistors**

The inherent constancy of period under balanced loading of this circuit suggested that it might be possible to devise a further modification which would give constant periods for unbalanced loads. A little thought shows that to achieve this the aiming potential for one half-period must be determined by the negative potential to which the collector of the non-conducting transistor goes during the other half-period. The diodes must now be used as half-wave rectifiers and if emitter-followers are used to reduce loading effects it is possible to make them serve a double purpose, as in Fig. 15, where they also serve to recharge the timing capacitors. Curve (d) in Fig. 14 shows the variation of half-period with output amplitude when loaded on one side only and it will be observed that the change is very small. Unfortunately this circuit is not self-starting because, as the layout of Fig. 15 clearly shows, it has a configuration in which bistable action is inherent via resistors R_t. Thus, unless large values of smoothing capacitor C_s are used this circuit fails to self-start (but for a different reason from that previously met) and even then does not start reliably and does not restart. It is interesting to note that capacitor C_s does not otherwise need to be large compared with the timing capacitor C_t for, if the behaviour of the network comprising C_s, R_t and C_t is analysed it is easily found that the condition to be satisfied is:

$$2V_t \exp(-T/CR_t) + 2V_t C/C_s \left[1 - \exp(-T/CR_t)\right] - V_t = 0$$

where V_t = negative excursion of the collector waveform,

$$T = \text{half-period and } C = \frac{C_s C_t}{C_s + C_t}$$

Thus T is independent of V_t if $V_t \neq 0$, regardless of the relative values of C_s and C_t.

* *By 'half'-period is meant the period of either of the two timing actions, but this does not necessarily imply that the full period is twice this value.*

A Compromise Circuit

The circuits presented so far have displayed on the one hand (Fig. 13) reliable starting together with constancy of period for balanced but not unbalanced loads, and on the other (Fig. 15) constancy of period for balanced and unbalanced loads but unreliable starting.

The circuit of Fig. 16, in which the emitter-followers are placed in a different position, was found to make the best overall use of two additional transistors and at the same time to give reasonable constancy of period for unbalanced loads and to be reliably self-starting. The emitter-followers are in fact used in the well-known 'super-alpha' configuration and resistors R_b are added to ensure rapid cut-off of VT_1 or VT_4 during the positive-going excursion at the base of VT_2 or VT_3 respectively. Due to the enhanced gain of the 'super-alpha' arrangement much higher values can be used for R_t, allowing smaller C_t for a given period and hence faster negative-going edges. In addition much longer periods can be achieved without inconveniently large capacitor values and the loading effect of R_t upon the collector is small enough to be neglected. A further advantage is that mark-to-space ratios of about 10:1 can be obtained (by reducing one of the R_t) since the recharging time for C_t is so much shorter than the half-period.

Added to these various advantages are a reasonable constancy of period with balanced or unbalanced load variations, as indicated by curve (e) of Fig. 14 and reliable self-starting and restarting. The reason why this circuit is self-starting while the similar one of Fig. 15 was not appears to be that the leakage current of, for example, VT_2 multiplied by the gain of VT_1 prevents VT_1 from being fully cut off under quiescent conditions so that proper bistable action cannot occur. VT_1 is, however, fully cut off during normal operation since a potential actually positive to emitter is applied to its base by the opposite collector via C_t and VT_3. This circuit, using pnp transistors only, displays all the features specifically regarded as desirable for use with a set of logic elements plus several which would make it suitable also for other applications. In fact, its low frequency capability gives it an advantage over the complementary circuit of Fig. 7, which was considered nearly perfect for the logic application, since this latter circuit suffers from the same limitation upon the upper value of R_t as the conventional circuit when high degrees of saturation must be maintained.

Conclusions

Some novel variants of the basic conventional astable multivibrator have been described and shown to fill satisfactorily the requirements listed as being desirable for use in conjunction with a set of general purpose logic elements. In particular, solutions to the problem of self-starting and constancy of period have been obtained in the circuits of Figs. 12, 13, 15 and 16. The circuit of Fig. 16 combines many desirable features which make it useful in a wider variety of applications.

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Printed-Circuit Boards Tested Automatically

Eighteen months ago, all printed-circuit boards used in I.C.T. 1900 Series computers were tested by a semi-automatic method involving the use of manually-operated testing devices and oscilloscopes. Testing one board could take anything from one to fifteen minutes, without a fault necessarily being discovered until the final check was made and, owing to the length of time required for testing, a careful preliminary inspection of each board was essential. This inspection located obvious errors such as diodes connected the wrong way round and resistors and capacitors having incorrect values.

Although this method of testing was extremely accurate, the process was too lengthy to cope with the rapidly increasing number of boards passing through the test department.

In order to streamline the test procedure, I.C.T. designed and produced two fully automatic board-testing devices. The first is used to test the a.c. characteristics of each board, including verification that circuit delays and rise-times are within prescribed limits. The second applies a straightforward d.c. test and so detects such faults as short-circuited diodes and the failure of transistors to saturate or switch off completely. Between them, the two devices provide an exhaustive range of automatic tests. The operator merely has to plug the board into a socket provided and observe that the 'pass' lamp comes on. He then removes the board and inserts the next one. Should the 'fail' lamp glow, further lamps simultaneously indicate the location of the fault on the board.

Each device is designed to test six different types of board, which together account for about 80 per cent of the logic circuits in I.C.T. 1900 Series computers.

Passing a board through both automatic testing machines in sequence takes less than 20sec. One pair of automatic machines can therefore test boards at the same rate as seven or eight of the previous manual devices. The brevity of the test obviates the need for an exhaustive preliminary visual inspection. All that is required is a check that the overall standard of soldered joints is adequate.

More detailed quality appraisal is carried out on a sample of boards, say 15 per cent, and the manual test equipment is used for this purpose. This is done so that wayward trends in circuit or component parameters are recognized early as a result of a prompt statistical analysis of each sample. Information can then be passed to the component specifier who is able to modify his specification accordingly and so avoid taking delivery of large numbers of inadequate components. Thus, apart from detecting faulty boards, the automatic testers are helping to reduce the number of faulty ones at source by fostering compatibility between the respective board and component specifications. As a result of this feedback process, the failure rate of boards under test has been reduced by 75 per cent.

The introduction of the new automatic board testers has increased the testing rate to 2 500 boards a week on single-shift working. A second pair of automatic testing machines will shortly be in operation and it is then hoped to increase the rate to over 6 000 boards a week. Eight manual devices and associated oscilloscopes—with operators—would be necessary to handle this quantity of boards. Yet, on average, one automatic device is no more expensive than one manual one and the labour requirement for automatic testing is considerably less.

Later on, should the increase in quantity warrant, a new type of automatic tester will be designed and manufactured to combine the facilities provided in the a.c. and d.c. testers so that only one plugging-in operation and one lamp observation is necessary instead of the present two.

An A.F. Power Amplifier with Crossover Feedback

By A. R. Bailey*, Ph.D., M.Sc., A.M.I.E.E.

The design of conventional a.f. power amplifiers for high-fidelity reproduction is inadequate for many laboratory uses. In particular the distortion is frequently excessive over part of the frequency spectrum, and stability is not often maintained on capacitive loads. This article describes the design philosophy and performance of an amplifier designed for laboratory use and having a much better specification than that which is customarily accepted.

Full rated power is obtainable over the whole audio frequency range at distortion levels of less than 0·1 per cent. In addition the amplifier is unconditionally stable for all load values and has good overload recovery performance.

(Voir page 847 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 854)

LABORATORY power amplifiers have design requirements that are considerably more stringent than those of power amplifiers feeding fixed loads. Full output current may be required into near short-circuit loads, often at very high frequencies. In addition the stability must be exceptionally good for the amplifier to be capable of operating into pure capacitive loads. If a very low distortion is also required over the audio frequency range, then the design problem becomes very difficult. Due to this it was decided to use valves rather than transistors, particularly as valves are far more robust when driving active loads or those possessing high energy storage.

The amplifier will need a large amount of overall feedback to reduce distortion and the output impedance to low levels. This in turn gives rise to instability problems so it was considered essential that all the amplifier stages were designed with the maximum possible bandwidth. In fact the design problems are very similar to those in video amplifiers.

The first amplifying stage must possess low noise and microphony as well as a wide bandwidth so the cascode circuit became the obvious choice. In order that adequate gain could be obtained with wide bandwidth a frame-grid valve was used.

The phase-inverter presented something of a problem. A high gain was required as well as a low output impedance and fairly low input impedance. The author had previously used a triode-pentode phase-splitter[1], but this proved to give inadequate drive and gain when used with low values of anode resistor. The low output impedance required (12kΩ) is necessary to avoid spurious gain losses at high frequencies due to the Miller effect in the output valves.

The circuit finally used is shown in Fig. 1. This is a pentode long-tailed pair and uses frame-grid valves. Ample drive can be obtained for the output valves along with quite high gain and low input capacitance. The circuit has the added advantage of the long-tailed-pair configuration in being suitable for d.c. coupling to the previous stage. This removes one l.f. time-constant with consequent improvement in l.f. stability.

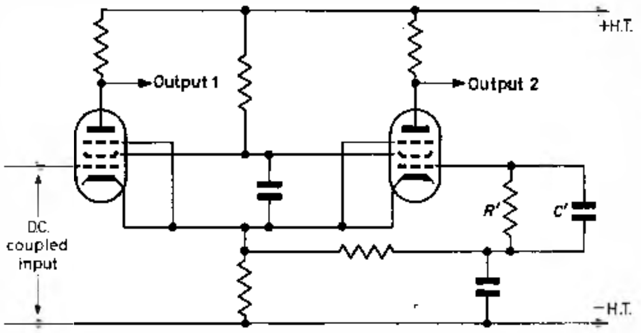

Fig. 1. Basic phase-splitter circuit

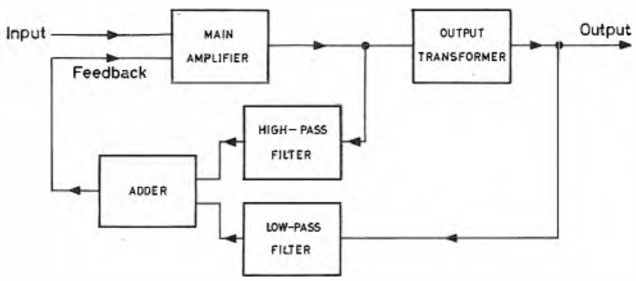

Fig. 2. Crossover-feedback system

The *RC* network *R'* and *C'* is included in the grid circuit of the grounded-grid section so as to equalize the a.c. impedances to the grids of the two sections. This greatly improves the overload performance due to the symmetrical characteristics that result.

The output valves are operated in the Blumlein 'ultra-linear' connexion so as to give minimum distortion. The capacitance between screen-grid and control-grid then gives rise to the high input capacitance mentioned previously. The output transformer must be of multi-section design to give both maximum bandwidth and good primary balancing at high audio frequencies. This latter requirement is more stringent in the tapped-load configuration used, as the screen tapping points must remain accurate at very high frequencies.

The bandwidth of the previous circuits can be made to extend up to the megacycle per second region, so the output transformer is obviously the limiting factor. If pure capacitive loads are placed on output transformers and the overall Nyquist plot produced, then it immediately becomes obvious that the secondary phase at very high frequencies can shift large amounts and still give

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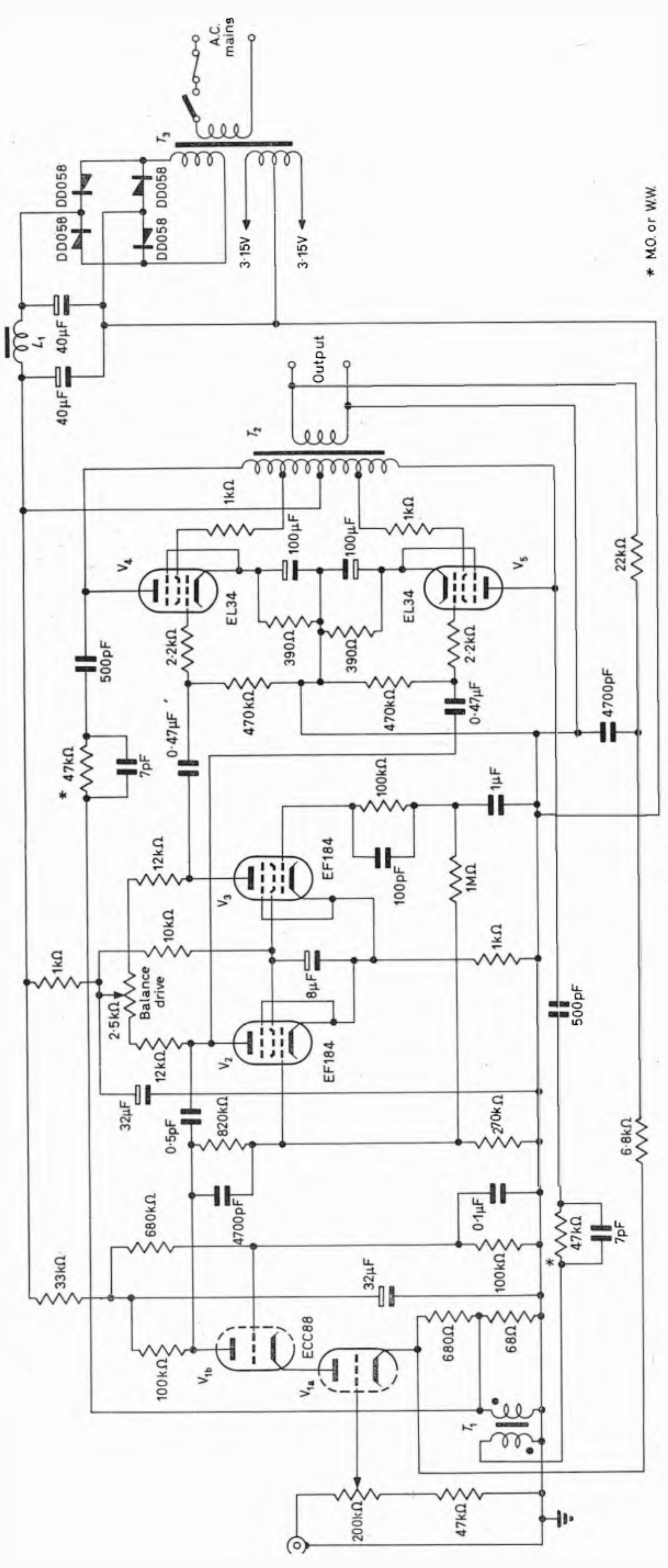


Fig. 3. The complete amplifier

considerable output. The means of overcoming this defect was suggested by Mr. D. Hafler of Dynaco Inc. Philadelphia. This was to the effect that if the feedback is taken from a separate transformer fed from the output transformer primary, then the stability is considerably improved under capacitive load conditions.

After some thought the reason became clear that the maximum phase-shift on the transformer primary can be only 90° as this is an impedance function. The secondary phase-shift can, however, have no final limits as this is a transfer-function comprising many elements.

Secondary feedback is necessary at low frequencies to reduce the iron distortion due to the transformer core. At high frequencies there is no longer this restriction so the feedback can be taken from the transformer primary. The crossover of feedback from primary to secondary can be made by complementary filters as shown in Fig. 2.

Unfortunately, the primary of the output transformer is driven in push-pull. The h.f. feedback must be derived from both anodes or the asymmetry results in high h.f. distortion and instability under peak drive conditions. After several experiments the best performance was obtained by using a bifilar mixing transformer to add the two h.f. anode voltages. Due to the crossover frequency being towards the top end of the audio spectrum, the transformer can be very small and iron distortion negligible.

Simple *RC* filter circuits are used for the crossover as these are found to be quite adequate.

Comparison of the amplifier performance with all the feedback from the secondary, with that of the crossover feedback, showed the crossover system to be far superior. Under resistive load conditions the amplifier stability was much improved due to the reduced h.f. phase-shift in the feedback loop. On open-circuit and capacitive loads the effect was even more marked as the load had then no damping effect on the transformer resonances.

The high gain with stability made it possible to utilize very high feedback values (40dB) and at the same time avoid the necessity of phase-advance stabilization of such magnitude that the amplifier response and rise-time suffered.

In addition, the effective bandwidth of the feedback over the output stage is so high that the stabilizing dominant lag of the whole circuit can be a simple *RC* time-constant starting at the extreme edge of the audio frequency range. This avoidance of a step-network inside the audio frequency spectrum prevents the falling feedback, and consequent distortion rise, that is such a feature of normal a.f. amplifiers.

The circuit is therefore much more 'designable' in that the deficiencies in the output transformer are removed from the feedback path. Production variations in h.f. resonances will therefore only have a small effect on the transient performance of the amplifier as a whole. Owing to the increased feedback that can be applied, it is now possible to obtain much better performance from amplifiers with output transformers.

The circuit of the complete amplifier is shown in Fig. 3 with the component values used in a practical circuit. The performance of the amplifier at full rated power output can be seen from Fig. 4, full output at less than 0·1 per cent distortion being available from 19c/s to 26kc/s.

Where very low distortion is not important, the amplifier will deliver considerably more power over a wide frequency range. Fig. 5 shows the performance obtained at a fixed distortion level of 1 per cent in the output.

Due to the large amount of feedback the amplitude response would be expected to be very flat. This is in fact the case as can be seen from Fig. 6, the audio range being covered with a gain tolerance of $\pm 0{\cdot}1$dB. Beyond this region the initial rate of gain fall is low to prevent ringing on transients. Some idea of the transient performance can be obtained from the square-wave output photographs shown in Figs. 7, 8 and 9.

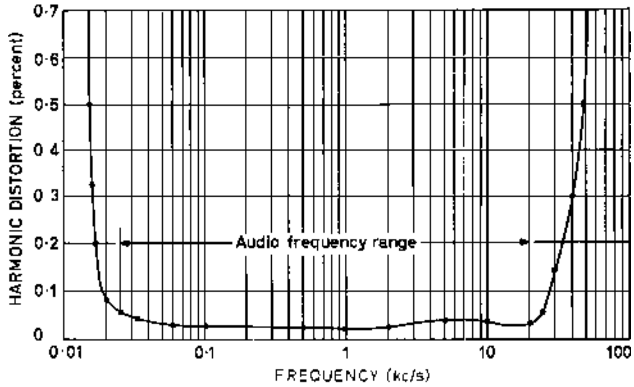


Fig. 4. Harmonic distortion of complete amplifier

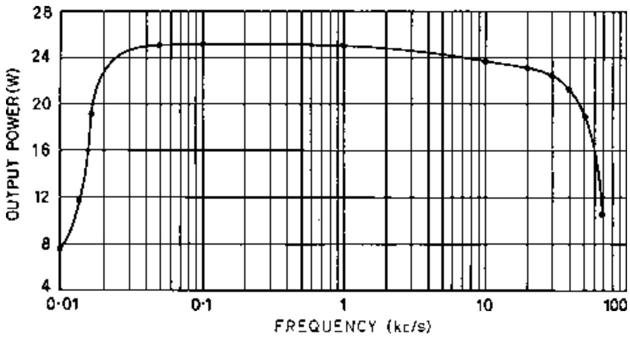


Fig. 5. Power output at fixed distortion level

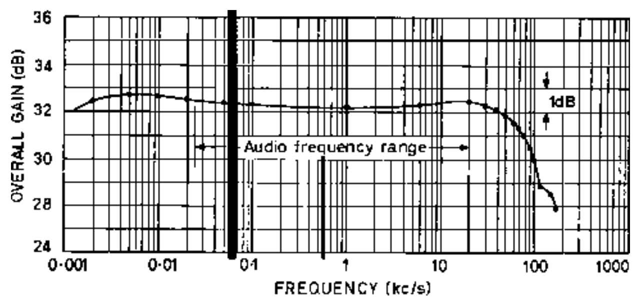


Fig. 6 (above). Overall frequency response

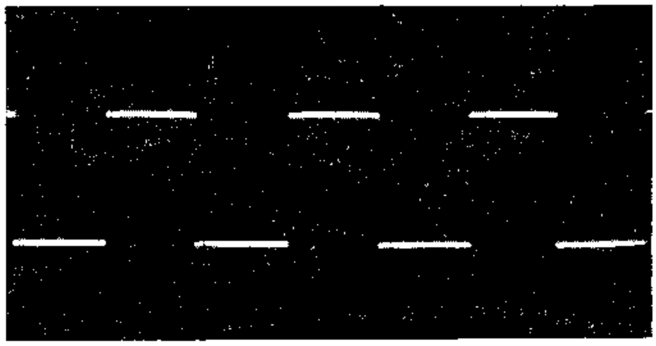


Fig. 8. Response to 1 000c/s square-wave input, 16Ω resistive load

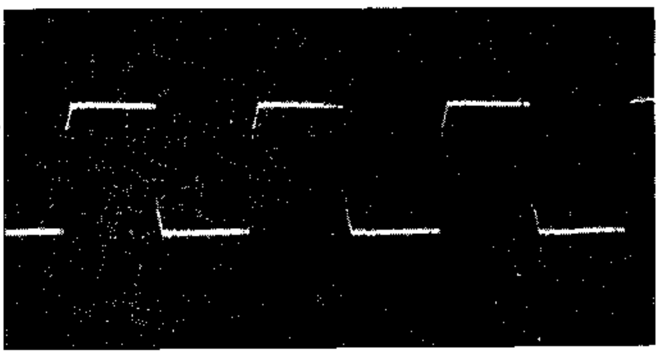


Fig. 9 (above). Response to 10kc/s square-wave input 16Ω resistive load

Fig. 7 (below). Amplifier response to 50c/s square-wave input, 16Ω resistive load

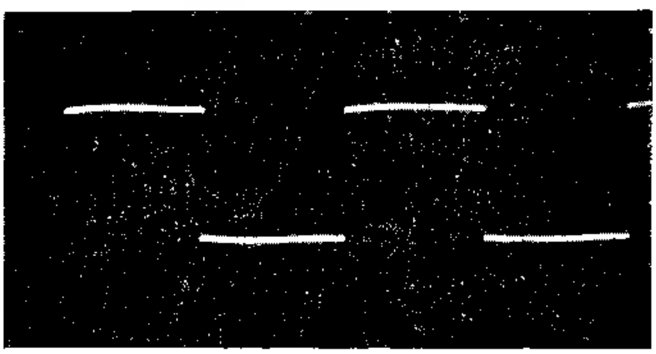
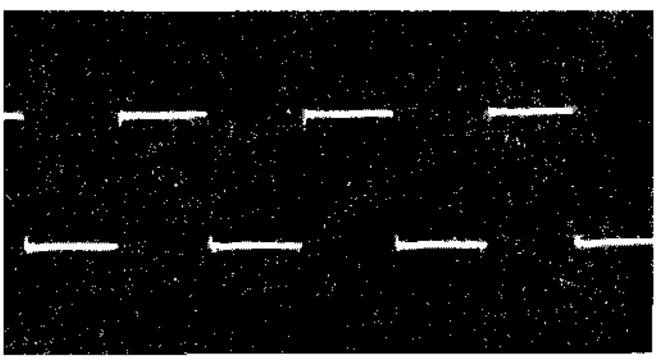


Fig. 10 (below). Response to 5kc/s square wave, load of 16Ω resistive in parallel with 0·05μF capacitance



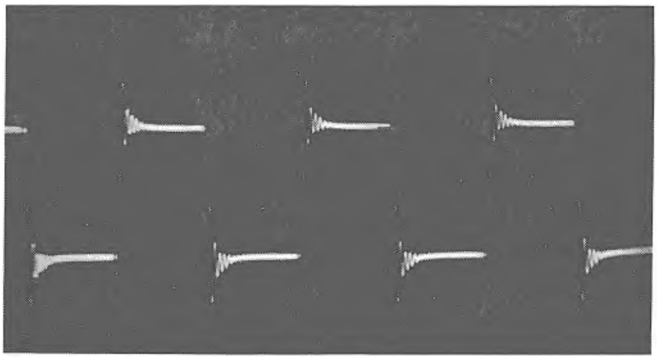


Fig. 11. Response to 5kc/s square wave, load of 0·35μF pure capacitance

The stability on capacitive loads was expected to be good due to the h.f. feedback being taken from the transformer primary. This is the case as is shown in Figs. 10 and 11, the capacitive load being chosen to give the maximum overshoot to the square wave.

Other details of the amplifier are:

Main loop feedback	40dB
Rated power output	15W
Input for rated power output	360mV
Intermodulation distortion (S.M.P.T.E.)	0·055 per cent
Hum and noise (Unweighted)	80dB on full output

Unconditionally stable for all conditions of drive and load. The feedback system is the subject of a patent.

Acknowledgment

The author would like to acknowledge the facilities of the Bradford Institute of Technology where much of the development was undertaken.

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A Digital Simulator for Naval Training

A Surface Tactical and Blind Pilotage Trainer has recently been designed and manufactured by the Military Systems Division of Solartron Ltd and is due to be installed at the Navigation and Direction School, Portsmouth. (H.M.S. Dryad).

The trainer employs a high speed silicon transistorized digital computer so linked with analogue computing elements in some of the peripheral functions to optimize the advantages of both the digital and analogue computer. The system will allow training in blind pilotage navigation, fleetwork, and manoeuvres —as well as tactical training of Command teams—under simulated, but essentially realistic conditions. The facilities provided are as follows:

(1) Six steerable radar-carrying ships. The radar picture from any one ship shows all other ships and targets within radar range (aerial height is adjustable to provide different detection ranges), and a high accuracy simulated radar coastline for training in blind pilotage navigation.

(2) Eighteen mobile targets which can be ships, aircraft, helicopters or submarines. All these craft are steerable with speeds up to 1 000 knots.

(3) Six fixed targets which can represent buoys or other sea marks.

(4) An operating area with sides 200 nautical miles long, and a positioning accuracy of 30 yards.

The outputs from each radar-carrying ship are fed to a separate operations room model where they feed the surface radar displays, automatic plotting tables and course and speed indicators.

The versatility of the trainer is such that in one period it is possible to carry out, say, basic training, blind pilotage navigation, fleetwork, team training and other exercises at the same time in different parts of an exercise area. In another period all six models can be fully manned to simulate two opposing fleets for tactical training of Command teams.

The following main items are incorporated in the tactical trainer:

A high speed special purpose digital computer: This digital computer has been specifically designed to derive from any two true positions the relative range and bearing in less than 15μsec. Essentially a cartesian/polar convertor, it continuously up-dates all cartesian positions, and rapidly cycles to establish relative polar co-ordinates.

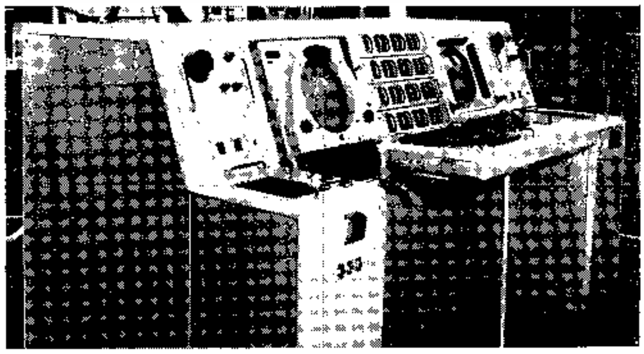
The computer proper, which has a built-in programme has 10 000 transistors and 60 000 diodes. It has an addition time of 1μsec, and the store is 1 024 words, 40 bits, with cycle time of two microseconds.

Control Desks: In the main simulator auditorium at HMS Dryad will be situated, facing a large-scale general situation display, two large desk consoles for use by control staff. The master and assistant controllers desk houses a normal surface radar display which can show the radar picture from any one of the six radar-carrying ships, a tactical display monitor, range and bearing indicators, repeater instruments, and various other controls. A larger console is for supervision of anti-submarine exercises and accommodates two controllers, helicopter pilots and tellers.

General Situation Display: A 35mm rapid processing photographic projector is used in conjunction with a tactical display to project the progress of the exercise on a large-scale screen.

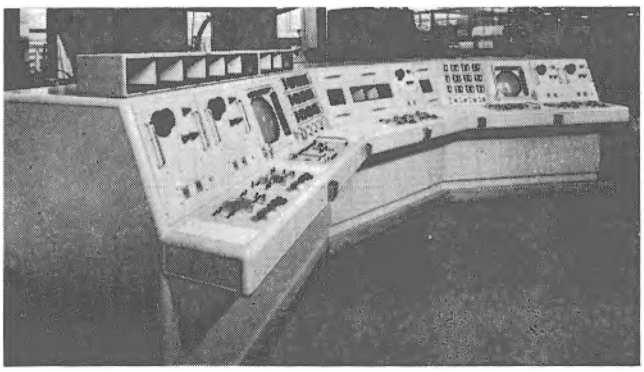
Craft Control Units: The control units for six radar-carrying ships are sited behind the chief controller. Facilities are provided to select ship or submarine characteristics for turning circles, acceleration etc., and settings such as size or targets, course and speed. Eighteen mobile targets and six fixed targets are also at the disposal of control staff.

Coastline Generators: Each of the six radar-carrying ships is provided with its own coastline generator where the movement of the ship is represented by movement of a carriage on which is mounted a photographic plate of the desired coastline.



Master and Assistant Controllers' Console. The desk is flanked by digital range and bearing indicators, and houses a radar display, tactical monitor, repeater instruments and various controls.

Anti-Submarine and Helicopter Controllers' Console, accommodating two anti-submarine controllers and three helicopter pilots, each controlling two helicopters.



The Choke-Zener Diode Filter Circuit

By J. M. Diamond*

Advantages of the choke-Zener circuit are discussed. These include short-circuit protection, regulation and hum reduction. When followed by a conventional degenerative regulator the choke-Zener circuit contributes a simple and foolproof form of short-circuit protection, and makes possible extremely low values of line regulation and hum output. The circuit is analysed for dissipation in the various elements, including the pass element of the degenerative regulator, under worst line and load conditions.

(Voir page 847 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 854)

FIG. 1 shows the choke-Zener diode circuit. It is similar to the ordinary choke-input circuit, except that the usual filter capacitor after the choke is replaced by a Zener diode, and the current-limiting resistor R is added. The Zener diode can replace the capacitor, since it can provide a low a.c. impedance after the choke. The capacitor shown with dotted connexions may be added, if needed, for additional filtering. While it is not an efficient circuit, because of losses in R and in the Zener diode, it may be very useful in low power, low voltage applications, such as transistor supplies. It offers the following advantages:

(1) The Zener diode replaces the presumably less reliable filter capacitor.
(2) The choke and power transformer may have high resistances, which then form part of R.
(3) As compared with a capacitor input filter, the circuit offers much better filtering, because it has two filtering elements. The form factor of current in the power transformer is better, which offsets somewhat the inherent inefficiency of the circuit. The choke is heavier, but not necessarily larger than the filter capacitor that would be required for capacitor input. As usual with choke input circuits, it is easy to specify the power transformer, because output regulation does not depend critically upon transformer impedance.
(4) The Zener diode gives a considerable degree of regulation against line and load changes.
(5) With proper design the circuit is short-circuit proof.
(6) The choke-Zener circuit is very convenient when followed by a conventional degenerative regulator, since it contributes a simple and foolproof form of short-circuit protection, and makes possible an extremely low overall line regulation, as well as extremely low hum output. It also supplies a well-defined input to the degenerative regulator, which permits a rather small voltage drop to be used in the pass element (usually a transistor), resulting in low pass element dissipation. The question of dissipation in the various circuit components will be dealt with in a later section, particularly under abnormal loading conditions.

Minimum Inductance and Current Requirement

For the ordinary choke input filter, it is well known that the instantaneous current in the choke must not reach zero. This requirement leads to the relation $\omega L = E_{dc}/3I_{min}$ for the full wave single phase rectifier, where ω is the fundamental angular frequency, E_{dc} is the rectified d.c. voltage, and I_{min} is the minimum d.c. through the choke. If the frequency is 50c/s, then:

$$L \geqslant (E_{dc}/942\, I_{min}) \quad \ldots\ldots\ldots\ldots\ldots\ldots (1)$$

For the choke-Zener circuit, equation (1) must hold as

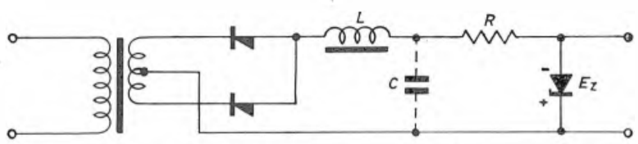


Fig. 1. Choke-Zener diode circuit

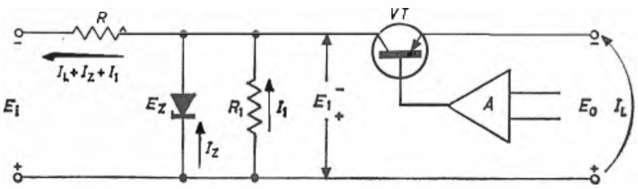


Fig. 2. Zener circuit followed by degenerative regulator

before, and in addition the current in the Zener diode must not reach zero. The current in the Zener diode, like that in the choke, is made up of a d.c. component and harmonic components.

If no capacitor is used (Fig. 1), the harmonic components of current in the Zener diode are the same as those in the choke. This leads to the same condition as before, as expressed by equation (1), except that now I_{min} is the minimum d.c. through the Zener diode.

Since minimum Zener current will occur at minimum input voltage, equation (1) may be written in the following form for this case:

$$I_{z(min)} = (E_{i(min)}/942L) \quad \ldots\ldots\ldots\ldots (2)$$

If this requirement is satisfied, then continuous current in the choke is also assured, since the d.c. in the choke is at least as great as that in the Zener diode. If a capacitor is added, the harmonic currents through the Zener diode are reduced greatly, and only a small minimum Zener current is needed, to ensure proper Zener breakdown.

Power Relationships in the Choke-Zener Circuit

Fig. 2 shows a Zener circuit followed by a conventional degenerative regulator. The latter uses a pass element, shown as a transistor, and a drive amplifier A.

* *Atomic Energy Commission, Denmark.*

The calculations to be made will also apply to the choke-Zener circuit standing alone.

As one of the advantages of the Zener circuit is its protection of the pass transistor VT during short-circuit, it is desirable to know the dissipation in VT for the worst loading condition.

This will be found, as well as the dissipation in the other circuit elements. The current I_1, flowing into R_1, represents a load connected before the pass transistor. An example of such a load is the current which may be supplied to the internal circuits of A.

The current and voltage symbols in Fig. 2 represent general quantities. Where needed, the subscripts 'min', 'max', and 'o' will be added to indicate minimum, maximum and design-centre values.

E_1 is the voltage across the Zener diode, while E_z is the Zener voltage—they will be equal when the diode is conducting.

Two additional parameters will be used: α, the permissible fractional variation of input voltage, and k, the ratio of design-centre input voltage to Zener voltage. Therefore:

$$E_{i(max)} = (1 + \alpha) E_{i(o)} \quad \ldots\ldots\ldots\ldots (3)$$

$$E_{i(min)} = (1 - \alpha) E_{i(o)} \quad \ldots\ldots\ldots\ldots (4)$$

$$k = E_{i(o)}/E_z \quad \ldots\ldots\ldots\ldots (5)$$

The principal circuit equation is:

$$E_i - E_1 = R\,(I_L + I_z + I_1) \quad \ldots\ldots\ldots\ldots (6)$$

The value of R which will permit regulation to be maintained down to an input $E_{i(min)}$, with a maximum specified load current $I_{L(max)}$, may be found from equation (6):

$$R = \frac{E_{i(min)} - E_z}{I_{L(max)} + I_{z(min)} + I_1} \quad \ldots\ldots\ldots\ldots (7)$$

$I_{z(min)}$ is determined by the considerations already discussed. Equation (7) may also be written:

$$R = \frac{E_z\,[(1 - \alpha)\,k - 1]}{I_{L(max)} + I_{z(min)} + I_1} \quad \ldots\ldots\ldots\ldots (8)$$

Maximum Pass Element Dissipation

If a short-circuit is applied to the circuit of Fig. 2, the control amplifier A will drive VT harder, in an attempt to maintain the output voltage. It is an essential design requirement that A still functions during short-circuit, to the extent of driving VT to saturation. If so, the voltage drop across VT, and the dissipation in VT, will be very small. The Zener diode will cut off, and the entire input power will be absorbed by R. If the supply need only operate safely at normal load and at short-circuit, it is sufficient to check the dissipation in VT at normal load. On the other hand, if safe operation is desired at all loads from normal to short-circuit, then the worst loading condition must be found.

An argument will now be presented to show that the worst loading condition occurs when all the Zener current I_z is diverted to the output. Consider the load current being raised gradually. It is clear that the dissipation in VT will increase as I_L increases, as long as I_z is greater than zero, since both E_1 and E_o, and therefore the drop $E_1 - E_o = E_z - E_o$, will remain constant. When I_z reaches zero, the Zener diode cuts off, and the situation changes radically. Further increases in I_L will cause E_1 to drop, because of the source resistance R. Eventually E_1 will drop so close to E_o that VT will saturate, after which the dissipation in VT will remain very low. Between these two conditions—Zener cut-off and saturation of VT—the

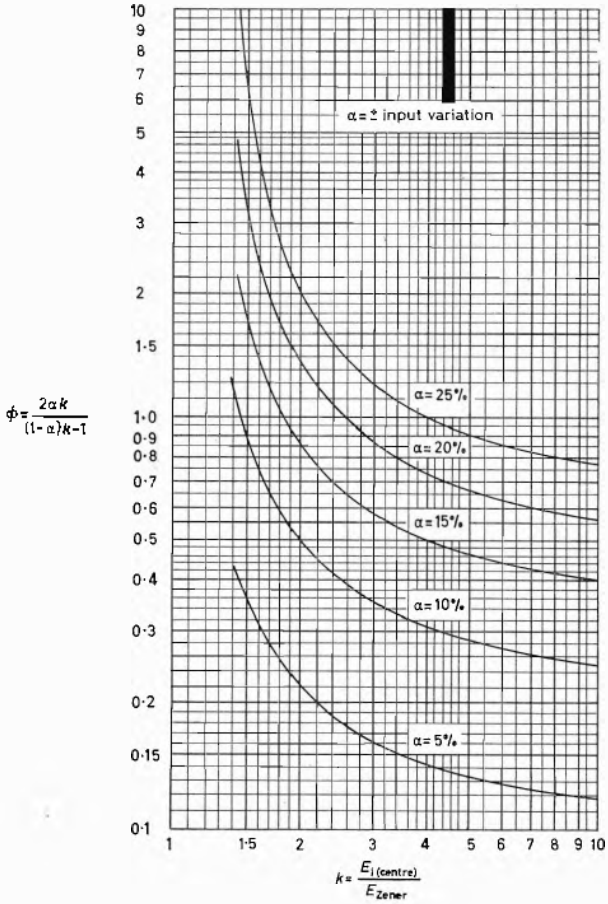


Fig. 3. ϕ **as a function of** α **and** k

circuit may be considered to consist of a source voltage $E_i - E_o$, a source resistance R, and a load VT. Maximum dissipation in VT will occur when VT is matched to the source, that is, when the voltage drop across VT is $\frac{1}{2}(E_i - E_o)$. The maximum voltage across VT occurs when the Zener diode is conducting, the value being $E_z - E_o$. If

$$E_z - E_o < \tfrac{1}{2}\,(E_i - E_o) \quad \ldots\ldots\ldots\ldots (9)$$

holds, then the 'matching' maximum will never be reached, and maximum dissipation in VT will occur at the moment of Zener cut-off. To show that this latter case is the one of practical interest, it may be observed from equation (9) that:

$$E_i > 2E_z - E_o \quad \ldots\ldots\ldots\ldots (10)$$

Equation (10) will certainly hold if E_o is nearly as large as E_z, which is the case for a supply designed for a fixed output voltage. Even if E_o is zero, which would require $E_i > 2E_z$, the requirement is reasonable, since a smaller input voltage will result in large swings in Zener current with changes in E_i. Finally, even if equation (10) is violated somewhat, the calculation of dissipation in VT is fairly accurate, because the 'matching' maximum, which then applies, is flat-topped, and has almost the same value as the dissipation at the moment of Zener cut-off.

The maximum dissipation in VT can now be calculated. From equation (6), the following is found for $I_L + I_z$, which is the current through VT at maximum dissipation, when $E_i = E_{i(max)}$:

$$(I_L + I_z)_{max} = \frac{E_{i(max)} - E_z}{R} - I_1 \quad \ldots\ldots\ldots\ldots (11)$$

Therefore the maximum dissipation in VT is:

$$P_{VT(max)} = [E_z - E_o]\left[\frac{E_{i(max)} - E_z}{R} - I_1\right] \quad \text{.. (12)}$$

If R has the value given by equation (7), then:

$$P_{VT(max)} = [E_z - E_o]\left[\frac{E_{i(max)} - E_z}{E_{i(min)} - E_z}(I_{L(max)} + I_{z(min)} + I_1) - I_1\right] \quad \text{.......... (13)}$$

Using equations (3), (4) and (5):

$$P_{T(max)} = [E_z - E_o]\,[(1 + \phi)(I_{L(max)} + I_{z(min)}) + \phi I_1] \quad \text{............. (14)}$$

Note that $I_{L(max)}$ is the maximum normal load current, not the current in the worst loading condition.

ϕ is defined as:

$$\phi = \frac{2\alpha k}{(1 - \alpha)k - 1} \quad \text{............ (15)}$$

ϕ is plotted as a function of α and k on Fig. 3.

In conjunction with equation (14), Fig. 3 shows the effect of varying k and α. In particular, it verifies the statement made above, in connexion with equation (10), that it is desirable for E_i to be greater than $2E_z$ ($k > 2$).

While equation (14) is a convenient explicit solution in terms of load current, I_1, and $I_{z(min)}$ (determined as discussed previously), it assumes that E_z is known accurately, and that the exact value of R, as given by equation (7), is used. To include the effect of tolerances on E_z and R, and of standard value selection for R, first find R from equation (7), then choose the closest standard value for R smaller than the value found. Adjust this value downward according to the tolerance of the resistor, and substitute this value into equation (12). The value of E_z is selected so that, with the Zener diode at the low tolerance limit, there is sufficient voltage to drive VT adequately.

Dissipation in the Zener Diode

Maximum Zener diode dissipation will occur at maximum input voltage and minimum load current. If the supply must operate safely with the load disconnected, $I_{L(min)}$ is zero. Thus:

$$P_{z(max)} = E_z\left[\frac{E_{i(max)} - E_z}{R} - I_1 - I_{L(min)}\right] \quad \text{.. (16)}$$

If R has the value given by equation (7):

$$P_{z(max)} = E_z\,[(1 + \phi)(I_{L(max)} + I_{z(min)}) + \phi I_1 - I_{L(min)}] \quad \text{............ (17)}$$

Equation (17) will be applicable when R is adjusted to take up the tolerance in E_z. In equation (17) E_z is taken at the high tolerance limit. If R is not adjustable, the value used in equation (16) is the same as that discussed in the previous section in the corresponding case. On the other hand equation (16) is not very sensitive to variation in E_z, although the maximum dissipation will usually be found at the high tolerance limit.

Maximum Dissipation in R

Maximum dissipation in R will occur during short-circuit, with maximum input voltage:

$$P_{R(max)} = \frac{E_{i(max)}^2}{R} \quad \text{.............. (18)}$$

If it is not necessary to provide for a steady short-circuit:

$$P_{R(max)} = \frac{(E_{i(max)} - E_z)^2}{R} \quad \text{.......... (19)}$$

Tabulated Design Equations

The results described above are tabulated in Table 1, together with equations covering normal operation.

Example

A choke-Zener filter is to be followed by a degenerative regulator. $E_o = 7{\cdot}5$V, $I_L = 160$mA, $\alpha = 20$ per cent, $I_1 = 17$mA. E_z is chosen as $10{\cdot}0$V ± 10 per cent: thus the maximum value of E_z is $11{\cdot}0$V. $E_{i(o)}$, the design centre input voltage, is chosen as $24{\cdot}0$V. Thus $E_{i(max)} = 28{\cdot}8$V; $E_{i(min)} = 19{\cdot}2$V. These voltages are no-load voltages, since source impedances are included in R. Allowing a $1{\cdot}0$V silicon rectifier drop, the d.c. component of secondary voltage should be $25{\cdot}0$V at design centre primary voltage. The no-load a.c. voltage at the secondary is therefore $1{\cdot}11\ (25{\cdot}0) = 27{\cdot}8$V across each half, for the centre-tapped full wave circuit.

A $0{\cdot}5$H choke is chosen for L. Equation (2) then gives

$$I_{z(min)} = \frac{19{\cdot}2}{942\ (0{\cdot}5)} = 41\text{mA}$$

R is found from equation (7):

$$R = \frac{19{\cdot}2 - 11{\cdot}0}{(160 + 41 + 17)\ 10^{-3}} = 37{\cdot}6\Omega$$

including transformer and choke resistance. Assuming, however, that R is actually adjusted to maintain regulation at -20 per cent input voltage, one may use equation (14). Using the high tolerance limit for E_z, $k = 24{\cdot}0/11{\cdot}0 = 2{\cdot}18$. Equation (15) or Fig. 3 gives $\phi = 1{\cdot}16$. Equation (14) then

TABLE 1
Design Equations

		P_T PASS ELEMENT DISSIPATION	P_Z ZENER DIODE DISSIPATION	P_R DROPPING RESISTOR DISSIPATION
Normal Operation		$(E_Z - E_O)\,I_L$	$E_Z\left(\frac{E_i - E_Z}{R} - I_1 - I_L\right)$	$\frac{(E - E_z)^2}{R}$
Maximum Dissipation Operation	Condition	$E_{i(max)}$, worst loading	$E_{i(max)}$, $I_{L(min)}$	$E_{i(max)}$
	Diss.	$(E_Z - E_O)\left(\frac{E_{i(max)} - E_Z}{R} - I_1\right)$	$E_Z\left(\frac{E_{i(max)} - E_Z}{R} - I_1 - I_{L(min)}\right)$	$\frac{(E_{i(max)} - E_z)^2}{R}$
	Condition	$E_{i(max)}$, worst loading, R as given by equation (7)	$E_{i(max)}$, $I_{L(min)}$, R as given by equation (7)	$E_{i(max)}$, short circuited output
	Diss.	$[E_Z - E_O]\,[(1+\phi)(I_{L(max)} + I_{Z(min)}) + \phi I_1]$	$E_Z[(1+\phi)(I_{L(max)} + I_{Z(min)}) + \phi I_1 - I_{L(min)}]$	$\frac{E_{i(max)}^2}{R}$

Notes: Equation (7): $R = \frac{E_{i(min)} - E_Z}{I_{L(max)} + I_{Z(min)} + I_1}$, $\phi = \frac{2\alpha k}{(1-\alpha)k - 1}$, $k = \frac{E_{i(o)}}{E_z}$, $E_{i(max)} = (1+\alpha)\,E_{(o)}$, $E_{(min)} = (1-\alpha)\,E_{i(o)}$

gives:

$$P_{T(max)} = [11{\cdot}0 - 7{\cdot}5]\,[2{\cdot}16\,(160 + 41) + 1{\cdot}16\,(17)] = 1590\text{mW} = 1{\cdot}59\text{W}$$

Dissipation at normal load is found using the formula in Table 1:

$P_T = (11{\cdot}0 - 7{\cdot}5)\,(160) = 560\text{mW} = 0{\cdot}56\text{W}$.

Maximum dissipation in the Zener diode is found from equation (17). In this case $I_{L(max)} = 160$mA, $I_{L(min)} = 160$mA, giving $P_{z(max)} = 3{\cdot}24$W. When the load is disconnected, $I_{L(min)} = 0$ and $P_{z(max)}$ rises to 5·0W.

With normal input voltage and normal load, P_z is found from the 'normal operation' formula of Table 1 to be 1·86W. Dissipation in R may be found using the formula of Table 1. Using the value of R found above, 37·6Ω, then for the following conditions:

Normal $E_i = E_{i(o)} = 24$V, low $E_z = 9{\cdot}0$V,

$$P_R = \frac{(24 - 9{\cdot}0)^2}{37{\cdot}6} \doteq 6{\cdot}0\text{W}$$

High $E_i = E_{i(max)} = 28{\cdot}8$V, low $E_z = 9{\cdot}0$V,

$$P_R = \frac{(28{\cdot}8 - 9{\cdot}0)^2}{37{\cdot}6} = 10{\cdot}4\text{W}$$

High $E_i = E_{i(max)} = 28{\cdot}8$V, short-circuit output,

$$P_R = \frac{(28{\cdot}8)^2}{37{\cdot}6} = 22{\cdot}0\text{W}.$$

A 15 to 20W resistor would be adequate, since the supply is not intended to work with a permanent short-circuit.

A Repetitive Electronic Switch for use in Isotope Research

By E van der S. Lotz*

In experiments with short-lived isotopes and for certain statistical tests, it is sometimes necessary to operate a nucleonic pulse counter many times in succession for relatively short periods. An electronic switch is described by means of which such counters can be operated repetitively for pre-determined periods of 1, 2, 4, 8 or 16sec, with 'off' periods of either 3 or 4sec between consecutive 'on' periods. The required timing sequence is generated by a 12-cathode glow-discharge tube, driven by pulses derived from rectified, unfiltered 50c/s sine waves. Both full- and half-wave rectification are used, with switch-over from one mode of rectification to another by means of a small gas tetrode.

(Voir page 847 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 854)

NUCLEONIC pulse counters are normally operated, i.e. made receptive to pulses, by manually throwing a switch which opens and/or closes two or more contacts. When this procedure has to be carried out many times in succession, appreciable errors could be introduced when stop-watches are used to time the 'on' periods, especially when the latter is of the order of a few seconds, as is the case when experimenting with short-lived isotopes, or when conducting certain statistical tests. More accurate results would obviously be obtained if the switching could be carried out electronically. The apparatus to be described has therefore been designed to open and/or close two or more contacts repetitively for 1, 2, 4, 8 or 16sec, with 'off' periods of either 3 or 4sec between consecutive 'on' periods. (It has been found that an 'off' period of 3 to 4sec is adequate for noting a reading and resetting the counter to zero).

Description

The circuit diagram of the electronic switch is shown in Fig. 1. Timing signals are generated by a glow-discharge tube, type GS12D, as follows: A multipole switch connects some of the 12 cathodes of the glow-discharge tube directly to earth, and others through a 100kΩ resistor to a potential of −20V (connexion X). The grids of a 6SN7 double triode are also connected to X. When the glow-discharge invests on a cathode connected to X, the potential at that point rises to approximately +15V, the 6SN7 becomes conducting, and relay A (type P.O. 3000) is energized. The relay contacts change condition, and the associated pulse counter is operated for a period determined by the number of consecutive cathodes connected to X, and the frequency of the GS12D drive pulses.

The fact that the GS12D glow-discharge tube has 12 cathodes implies that, to ensure repetitive operation, the sum of the 'on' and 'off' periods will have to be 2, 3, 4, 6 or 12 time units. Assuming drive pulses of one per second, cathode connexion 1, Fig. 2(a), could be used to yield periods of 1sec 'on' and 3sec 'off', connexion 2 for periods of 2sec 'on' and 4sec 'off', and connexion 3 for periods of 8sec 'on' and 4sec 'off'. Again using one second drive pulses, an 'on' period of 4sec would necessarily have to be followed by an 'off' period of either 2 or 8sec, the first being too short, and the second unnecessarily long. This problem was overcome by feeding the GS12D one pulse every two seconds during the 'on' periods, and using connexion 2. Connexion 3 is similarly used for 'on' periods of 16sec, with two-second drive pulses during the 'on' periods. Periods longer than 16sec can easily be obtained by simply shorting the relay contacts for the required number of cycles (where a single

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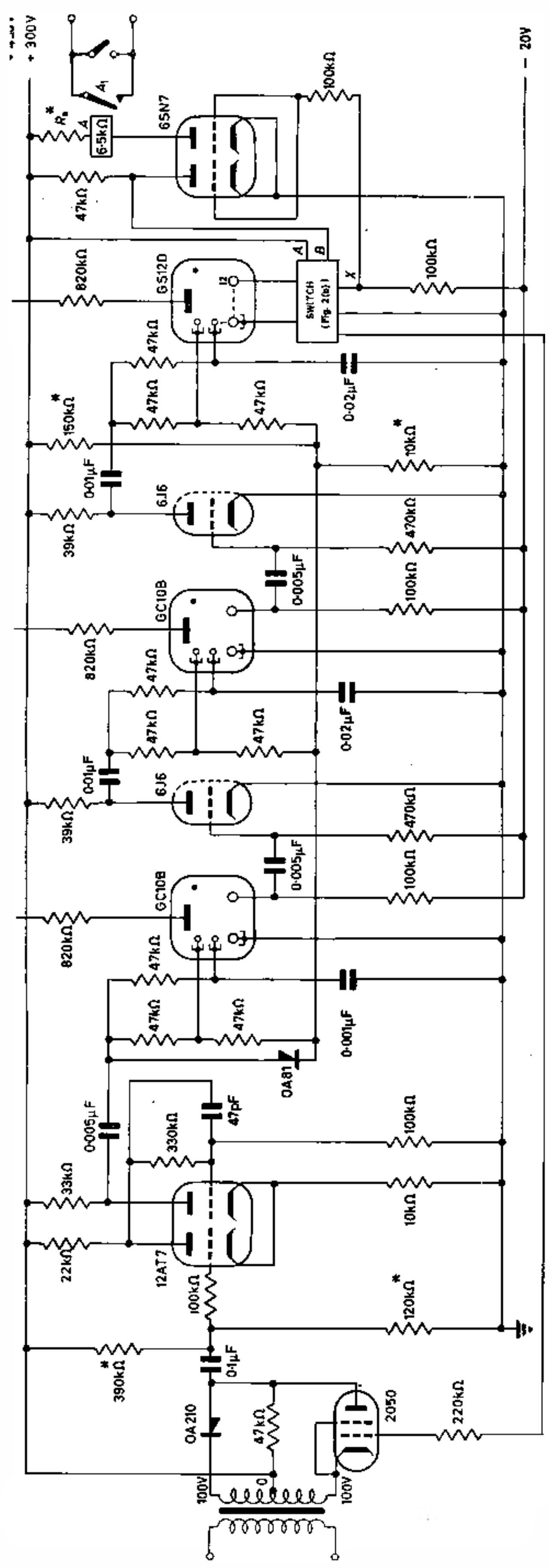


Fig. 1. Circuit diagram of repetitive electronic switch

pair of normally open contacts are used) during the 'off' periods.

Drive pulses of either one per second or one every two seconds can conveniently be derived from unfiltered full- or half-wave rectification of 50c/s sine waves, followed by a 100-fold division of pulse repetition frequency. It is obviously essential that, to ensure reliable operation, the change-over from half-wave rectification and *vice versa* should take place within a few milliseconds. In the apparatus described this fast change-over is achieved in the following manner: A full-wave rectifier circuit has in one arm a silicon diode, and in the other a small gas tetrode (thyratron), type 2050. The centre-tap of the transformer is at +300V. For 'on' periods of 1, 2 or 8sec, the control grid of the thyratron is connected through a 220kΩ resistor to +300V, the thyratron is consequently fully conducting, and full-wave rectification results. For 'on' periods of 4 or 16sec, however, the thyratron grid is connected to that anode of the 6SN7 having the resistive load. As the 6SN7 conducts during the 'on' periods, the anode potential drops to a low value, the thyratron is cut off, and half-wave rectification results.

The output from the rectifier circuit triggers a 12AT7 univibrator[1] which drives the first of two GC10B Dekatron counters, the function of which is to effect the required 100-fold division of pulse repetition frequency. It was found necessary to drive the first GC10B in this way, because the inherent slowness of sine wave drive would introduce an appreciable time lag in the circuit, and the thyratron would not be cut off fast enough, impairing the reproducibility of the 4 and 16sec 'on' periods. The grid of the thyratron could not, for similar reasons, be connected to that anode of the 6SN7 having the relay as load. In this case the drop in anode potential when the tube becomes conducting is relatively slow due to the inductance of the relay, again causing a delay in the cutting off of the thyratron. The drive circuit of the second GC10B, as well as that of the GS12D is conventional, using one half of a 6J6 double triode in each circuit.

An examination of the required GS12D cathode connexions, shown in Fig. 2(a), will show that cathode 1 can permanently be connected to *X*, cathodes 10, 11 and 12 to earth, and cathodes 3/4 and 7/8 to each other. The complete switch connexions are shown in Fig. 2(b).

It is of interest to note that, as cathodes 10, 11 and 12 are permanently earthed, the apparatus can be switched from one 'on' period to another without disturbing the continuity of the timing sequence, whenever the glow-discharge rests on any of these cathodes. This is very useful in the determination of the half-lives of short-lived isotopes, where, to avoid a decrease in counting accuracy, the period of measurement has to be increased progressively as the activity of the sample decreases, without affecting the total time as represented by the sum of the 'on' and 'off' periods. In the apparatus described, the position of these cathodes has therefore been marked on the front panel to facilitate switching at the correct moment.

Results and Discussion

The apparatus was tested as follows: A nucleonic pulse counter was fed pulses having a crystal-controlled repetition rate of 1000 per second, and controlled by a pair of contacts of relay *A*. The number of pulses registered during each 'on' period was therefore an indication of the duration of the period to the nearest millisecond. Both mains frequency and a 50c/s crystal-controlled oscillator were used for the generation of drive pulses. One hundred consecutive readings were taken for each 'on' period, and their mean values and standard errors calculated.

When using a crystal-controlled source of 50c/s sine waves, the deviations from the mean values were apparently less than 1msec, as readings of 1000, 2000, 3999, 8000, and 15999 were consistently recorded. The fact that the 4 and 16sec periods appeared to be 1msec too short may be of some academic interest, but is of no consequence for the purpose for which the apparatus was designed. The relatively high order of reproducibility was somewhat surprising, bearing in mind the fact that an ordinary Post Office type relay was used.

The results using mains frequency are summarized in Table 1.

Although the reproducibility of the 'on' periods is o more importance than the absolute duration thereof, i is of interest to examine the factors affecting the latte The main factor is naturally the electrical and mechanica characteristics of the relay. A test on four different Po Office type relays, all with coil resistances of approxi mately 6·5kΩ, and using identical series resistance yielded 'on' periods varying from 3·971 to 4·022se with the switch set to '4sec'. The value of the resistanc R_s, in series with the relay had, as could be expected, als an appreciable effect on the duration of the 'on' period With the relay used in this apparatus, and the switch agai set to '4-sec', a period of 4·020sec was recorded with th series resistor shorted, and 3·981sec with a series resisto of 68kΩ. Altering the value of the series resistor thu appears to be a convenient way of setting the duratio of the 'on' periods to an exact value.

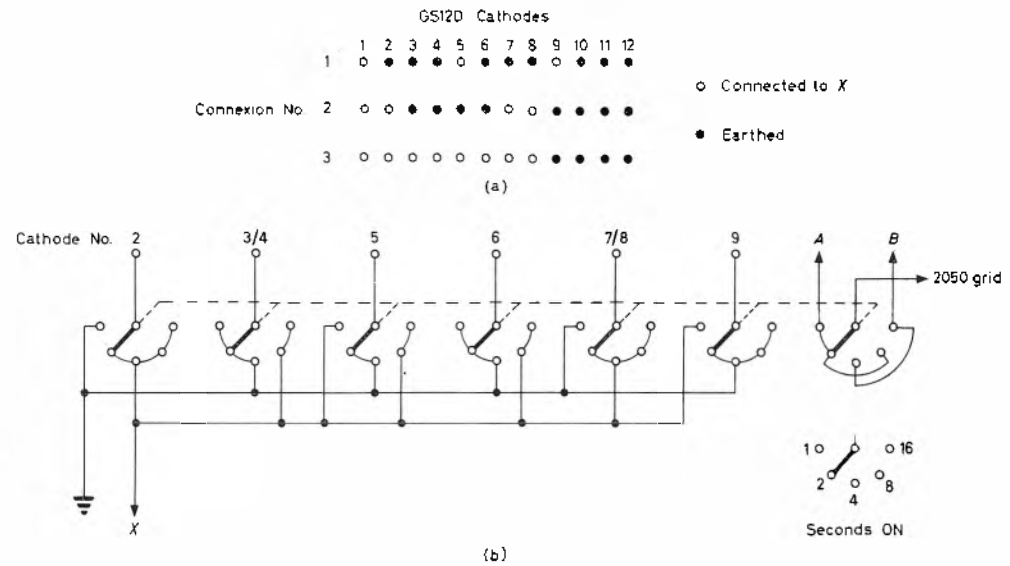


Fig. 2. GS12D cathode connexions

The results in Table 1 clearly illustrate the variations in mains frequency, and it is obvious that a crystal-controlled source of 50c/s sine waves would be essential where great accuracy is required. Mains frequency could however be used in routine experiments with short-lived isotopes, and would still yield far better results than could be realized with a stopwatch.

The maximum deviations from the mean values listed in Table 1 would conceivably vary during the day. It is interesting, however, that during the test period the maximum deviations of the mains frequency remained fairly constant at ±0·4 per cent, representing 0·2c/s.

TABLE 1

Mean number of pulses registered and standard error of 100 consecutive readings, using mains frequency

'ON' PERIOD (SEC)	MEAN NUMBER OF PULSES REGISTERED	STANDARD ERROR
1	999·46	±0·190
2	2000·25	±0·332
4	3996·51	±0·660
8	8007·50	±1·549
16	16011·64	±2·923

Of less importance were variations in the direct voltage used: It was found that a reduction of the anode suppl voltage from +300V to +290V resulted in a shortenin of the 'on' periods by approximately 2msec, while change in the negative voltage from −20V to −25V gav rise to a shortening of the 'on' periods by approximatel 3msec. It should be noted that none of the above change in the duration of the 'on' periods had any effect on th reproducibility of the readings.

It was furthermore verified experimentally that th duration of the 'off' periods was as reproducible as tha of the 'on' periods, and that the sum of the 'on' an 'off' periods were exactly 4, 6, 8, 12 and 20sec respec tively. These facts were used with advantage in th determination of the half-lives of short-lived isotopes where the sum of the 'on' and 'off' periods was take as the interval between consecutive readings.

This instrument has proved its reliability and accurac in a number of successful experiments, *inter alia* th determination, to a high degree of accuracy, of the half-life of barium-137m, and in resolving time determination of nucleonic apparatus.

REFERENCE

1. Ericsson Telephones Ltd. Cold Cathode Tubes Handbook, 49 (1954/55)

An R.F. Stimulus-Isolation Unit for Use in Electrophysiology

By W. T. Catton*, M.Sc. and R. G. Farrier*

The rectangular pulses from a conventional stimulator are used to screen-modulate an electron-coupled tetrode r.f. oscillator operating at 5Mc/s. The resulting r.f. pulses are fed through a low-loss coaxial cable to a tuned full-wave detector, whose d.c. output is isolated from earth. Good power-conversion efficiency is achieved and much greater power can be delivered than is possible with earlier types of isolation unit

(Voir page 847 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 854)

A WEAKNESS of earlier stimulus isolation units (e.g. Schmitt and Dubbert[1]), in which the stimulus pulse directly anode-modulates an r.f. oscillator, with radio-link coupling to a conventional detector, is the low power

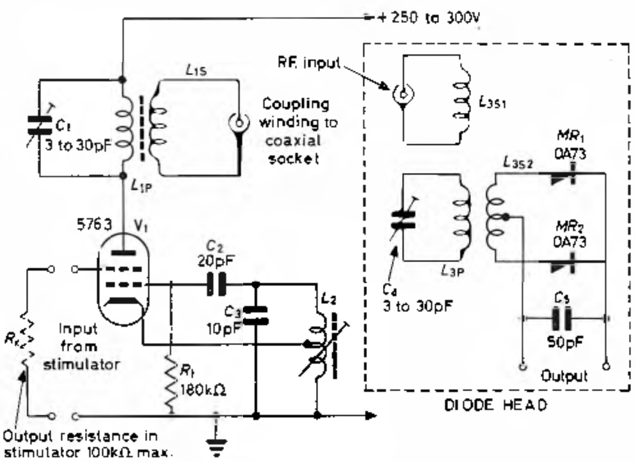


Fig. 1. R.F. stimulator link

Coils:

L_1 wound on 7/16in diameter former with 4mm dust core. L_{1P} = 1in close wound 36 s.w.g. L_{1S}=4 turns 30 s.w.g. L_2=former as L_1 75 turns 36 s.w.g. Tapped at 10 turns. L_3 = 1in o.d. Perspex former threaded 40 t/in by ·010in deep. L_{3P} = 1¼in winding of 30 s.w.g. L_{3S1} = 3 turns 30 s.w.g. L_{3S2} = 16 turns 30 s.w.g. Centre tapped. Insulated from primary by layer of tissue

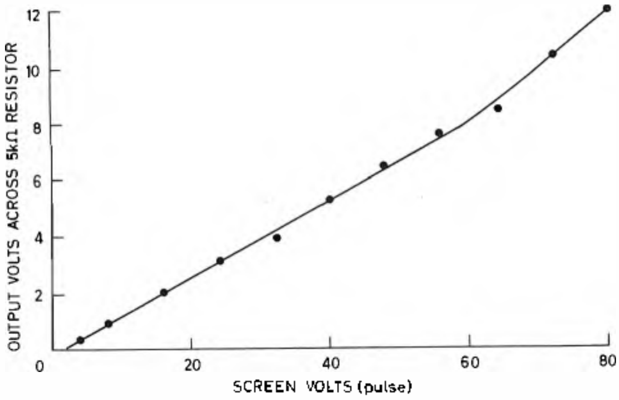


Fig. 3. Input/output linearity

available. The output voltage diminishes sharply when the load resistance falls below about 1kΩ, a not unusual situation in physiological experiments when tissue is being stimulated while immersed in saline. Insufficient stimulus current may be available in these circumstances and the pulse waveform is distorted.

In the present design (Fig. 1) the tetrode r.f. oscillator (5763) draws power from a conventional 250V h.t. supply, and is quiescent when the screen voltage is zero. The screen voltage is provided by the positive rectangular pulses delivered from the cathode-follower output of the main stimulator, connexion being direct. Oscillation commences when the screen voltage rises only one or two volts from zero, and the initial step effect is very small.

The use of a full-wave detector greatly increases power-conversion efficiency, but this was also found to be rather critically dependent on careful winding of the secondary (L3S2) of the air-cored r.f. transformer. The two half-windings should be exactly physically symmetrical. It is further essential to use r.f. type germanium diodes in the rectifier circuit. The screening can should be large compared to size of coil; an alloy can 4in by 2in diameter was found satisfactory. The capacitance to earth from the commoned output sockets was 0·02pF. The detector unit is placed as close as possible to the preparation, with short unscreened leads to the stimulation electrodes. The unit is tuned by feeding in a known pulse and peaking C_1 and C_4 with the output viewed on the oscilloscope.

The performance of a prototype unit is shown in Fig. 2, in which output voltage and power are plotted against load resistances ranging from 10Ω to 5kΩ. It will be seen that there is a sharp peak in output power at a load resistance close to 100Ω. Useful power was obtained even with a load of 10Ω, into which a current of 130mA was delivered. The output voltage was about one-eighth of the input and the relation was linear over a wide range (Fig. 3).

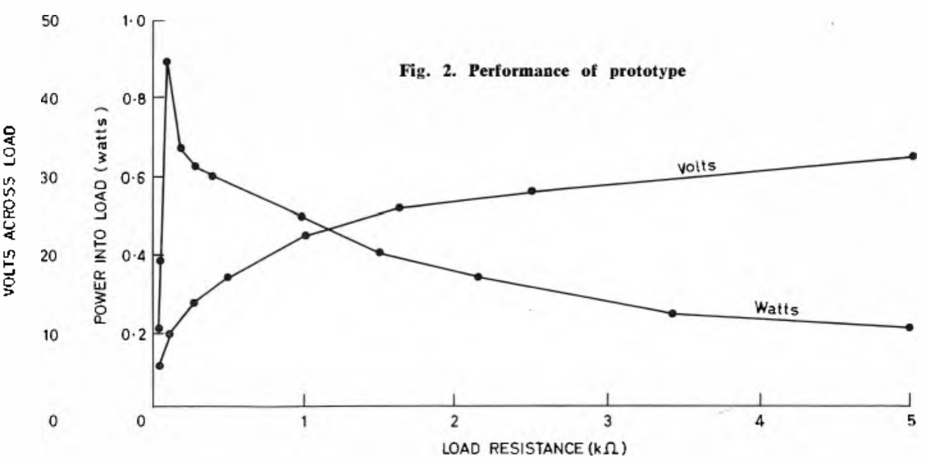


Fig. 2. Performance of prototype

REFERENCE

1. Schmitt, O. H., Dubbert, D. R. Tissue Stimulators Utilising Radio-frequency Coupling. *Rev. Sci. Instrum. 20*, 170 (1949).

* *University of Newcastle upon Tyne*

LETTERS TO THE EDITOR

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

Probability Density Function Measurement

DEAR SIR,—I am writing in reference to the article, "A Technique for Probability Density Function Measurement" by P. N. Nikiforuk and G. Squires in the May 1965 issue of ELECTRONIC ENGINEERING. A somewhat ambiguous, if not misleading, statement is made on page 317. While discussing the deviations of the experimental points from the theoretical curve, the authors suggest that "a better approximation to the actual probability density function could have been obtained by using a smaller value of $\Delta\nu_b$." As a matter of fact this is not the case at all unless the sampling time is increased by a greater proportion.

A reasonable measure of the deviation of the experimental points from the theoretical density is the variance of the density estimate. The estimate used is given on the same page as

$$p_{i(\max)} = \text{Estimate of } P(\nu_i)$$
$$= n_i/N\Delta\nu \qquad, (\Delta\nu \equiv \Delta\nu_b).$$

Denoting mathematical expectation by E, we have

$$E(p_{i(\max)}) = [F(\nu_i) - F(\nu_i - \Delta\nu)]/\Delta\nu$$

where

$$F(\nu) = \int_{-\infty}^{\nu} P(x)dx$$

Let

$$\Delta F_i = F(\nu_i) - F(\nu_i - \Delta\nu)$$

Then $y_i = n_i/N$ is a binomial variable with parameters ΔF_i and N and it is well known that its variance is given by

$$\text{Var}(y_i) = \Delta F_i(1 - \Delta F_i)/N$$

(It has been assumed that the sample points of the random process are independent from pulse period to pulse period.)

Thus

$$\text{Var}(p_{i(\max)}) = \Delta F_i(1 - \Delta F_i)/N(\Delta\nu)^2$$

For $\Delta\nu$ small enough

$$\Delta F_i \simeq \Delta V P(\nu_i)$$

so that

$$\text{Var}(p_{i(\max)}) \simeq P(\nu_i)[1 - \Delta\nu P(\nu_i)]/N\Delta\nu$$
$$\simeq P(\nu_i)/N\Delta\nu$$

Therefore it can be seen that decreasing $\Delta\nu$ will increase the variance of the estimate and will most certainly not improve the approximation. A better fit can be achieved only by increasing the product $N\Delta\nu$, i.e., by increasing the sampling time by a factor greater than that by which $\Delta\nu$ is decreased. The above formula also points out something else of interest, namely that the variance of $p_{i(\max)}$ is proportional to $P(\nu_i)$ for small values of $\Delta\nu$. This is borne out by noticing that in Fig. 5, page 317, the errors are greatest near the node and are negligible at the tails.

If the sample data points are not independent from pulse position to pulse position, i.e., if the half-bandwidth of the random signal (point A in Fig. 1, page 316) is less than the frequency of the oscillator (point F), then, assuming the spectrum of the random signal to be centred at zero frequency, an equivalent value of N is approximately

$$N_e = (B_s/2f_o)N \text{ for } B_s \leqslant 2f_o.$$

where $B_s/2$ is the half-bandwidth of the process and f_o is the oscillator output frequency. Thus for $B_s \leqslant 2f_o$

$$\text{Var}(p_{i(\max)}) \simeq P(\nu_i)(B_s/2f_o)/N_e\Delta\nu$$

Yours faithfully,

J. KEILIN

Bell Telephone Laboratories,
New Jersey, U.S.A.

The authors reply:

DEAR SIR,—Dr. Keilin presents an excellent clarification of a point in our article that may have been incorrectly interpreted by some readers. We wish to thank him for his comments and to let him know that we were aware of the need for increasing the sampling time.

Yours faithfully,

P. N. NIKIFORUK
G. SQUIRES
University of Saskatchewan,
Canada.

Pulse Modulated Amplifiers

DEAR SIR,—Messrs. Bell and Sergeant (August 1965) have provided a detailed analysis of the sideband structure of various types of width modulated pulse waveforms. It is apparent that sideband information which trespasses into the audible spectrum constitutes a component in the output of the amplifier not present at the input, and in this sense it is appropriate to classify it as distortion.

This is, however, quite a different kind of distortion to that normally experienced in audio amplifiers, and we are at once confronted with a difficulty and there is as yet no established limit to the magnitude of this distortion which may be excepted as tolerable.

It is noteworthy that in practice the distortion is only produced to any significant extent when the amplifier is called upon to deliver relatively large power at high frequencies, whereas speech and music material normally contains relatively little energy in the critical frequency range.

Many domestic and semi-professional tape recorders introduce a similar type of distortion which is very apparent if tests are made with an audio signal generator. There is some evidence to suggest that the subjective effect of the distortion on normal programme material is not unduly unpleasant, in that many critical listeners have failed to notice its presence in taped material.

The distortion becomes audible in a p.d.m. amplifier if the p.r.f. is deliberately made low, and in general has the character of background hiss accompanying sounds which have a large h.f. content. There is clearly a need for subjective listening tests aimed at determining an acceptable limit.

Sideband distortion has one feature in common with the more familiar distortion in that it is amenable to reduction by a.f. negative feedback.

In reference 3 of the article by Bell and Sergeant my name was incorrectly given as D. R. Britt.

Yours faithfully,

D. R. BIRT,
Oxted, Surrey.

The authors reply:

DEAR SIR,—We note Mr. Birt's comments on the distortion produced in pulse duration modulated amplifiers, and agree that with respect to audio amplifiers it is possible to establish an acceptable limit to the distortion. For amplifiers using conventional techniques, limits have been established which include both the harmonic distortion and intermodulation distortion produced in the amplifier.

Similar distortion products are produced in amplifiers using p.d.m. techniques due to the practical failings of the circuits and to these components the same suggested limits will apply.

In addition, however, a further distortion component is produced due to cross-modulation with the pulse repetition frequency. This distortion is of different form since it is cross-modulation between the varying signal frequency and a fixed amplitude, fixed frequency component. Since the pulse repetition frequency is chosen to be high, such cross-modulation distortion should not be as unpleasant as a similar amplitude of cross-modulation distortion between two frequency components whose difference frequency lies lower in the audible range.

Indeed, as Mr. Birt points out, distortion of this type often goes unnoticed in tape recording equipment.

However, we would agree that the results of subjective listening tests would be most interesting.

Yours faithfully,

E. C. BELL,
T. S. SERGEANT,
Bradford Institute of Technology.

Short News Items

The Fourth International Flight Test Instrumentation Symposium is to be held at the College of Aeronautics, Cranfield, on 21 to 24 March next year, under the new title of International Aerospace Instrumentation Symposium.

The meeting which is organized by the Department of Flight will be co-sponsored by the Aerospace Division of the Instrument Society of America.

Its purpose will be to bring together people concerned with the instrumentation and ground test support of aircraft, spacecraft, space-probes, satellites and other aerospace vehicles, in order to exchange information by formal presentation of papers and by discussion.

During the course of the meeting there will be an exhibition of equipment by invited companies.

Details concerning registration and programme are available on application to the Symposium Organizer M. A. Perry, College of Aeronautics, Cranfield, Bedford.

The Salon International des Composants Electroniques to be held in Paris on 3 to 8 February next year will now take place in the Halls Renan at the Parc des Expositions.

This exhibition hall has a floor area of some 350 000 ft² on two floors.

The Institute of Physics and The Physical Society propose to hold a conference on Solid State Physics in the Renold Building, Manchester College of Science and Technology, from 4 to 7 January 1966. This conference will follow the same general pattern as the meetings held in Bristol in preceding years.

Two copies of outlines of talks not longer than about 200 words, preferably typed on quarto paper in double spacing, should be sent to the Papers Secretary, Professor S. F. Edwards, Department of Physics, The University, Manchester 13. The presentation time for contributed papers will be not more than ten minutes. A leaflet (Reference ACB78) giving guidance on the preparation and submission of outlines of talks is available from the Institute and Society.

All enquiries regarding attendance and the accommodation which will be available in the University Halls of Residence should be addressed to the Meetings Officer, The Institute of Physics and The Physical Society, 47 Belgrave Square, London, S.W.1.

The Fifth International Television Symposium will be held at Montreux, Switzerland, on 22 to 26 May 1967, in the rebuilt Montreux Theatre.

The British Radio Valve Manufacturers' Association (BVA) and Electronic Valve and Semiconductor Manufacturers' Association (VASCA) announce the following sterling value of their members' sales during the three months ended 30 June, 1965:

Valves & Tubes	£10·2M
Semiconductor devices	£ 7·3M
Total	£17·5M

This brings the total value of sales for the first half of 1965 to £35·4 million.

A revision of B.S. 2131—'Fixed Capacitors for direct current using impregnated paper or paper/plastics film dielectric' has been published. It covers fixed capacitors designed essentially for d.c. use at not more than 6 300V and having a dielectric of paper or paper combined with plastics film. The electrodes are either thin metal foils or metal layers deposited on the dielectric.

The Standard gives uniform requirements for judging the electrical and mechanical properties and the effects of climate on this type of capacitor, and describes the appropriate test methods.

The revised Standard is identical in technical content and layout with IEC publication 80, with very few exceptions, and an indication of these differences is given in the Standard itself.

Copies of B.S. 2131 may be obtained from the BSI Sales Branch, 2 Park Street, London W.1. Price 15s. each. (Postage 1s. will be charged to non-subscribers.)

The China National Machinery Import-Export Corporation and the Compagnie Francaise Thomson-Houston of Paris have recently signed a contract under which the latter is to deliver to the Chinese Republic two short wave 350kW transmitters.

These transmitters, the most powerful yet supplied for short wave transmission, make use of the Vapotron technique developed by C.F.T.N.

Frequency changes can be made in less than 45 seconds.

The Eleventh International Exhibition of Measuring and Control Techniques and Automation (JUREMA) will be held at Zagreb, on 16 to 24 April next year.

Manufacturers in the United Kingdom who are proposing to exhibit at Zagreb should make application to Exhibition Consultants Ltd, 11 Manchester Square, London, W.1.

During the Exhibition a seminar will be held dealing with the mathematics and theory of control, regulation and cybernetics on which papers are invited.

Prospective authors are asked to apply for details to JUREMA P.O.B. 123, Zagreb, Yugoslavia.

The British Computer Society has published a list of Computer and Associated Courses to be held in the United Kingdom during the coming session.

Detailed in this publication are 224 academic, industrial and commercial courses run by Universities, Colleges of Advanced Technology, Technical Colleges, Professional Institutes, and Consultants. It does not include computer manufacturers' courses.

Copies are available from the British Computer Society, Finsbury Court, Finsbury Pavement, London, E.C.2. Price 4s.

The first two BBC-2 u.h.f. television relay stations for the London area have now been brought into service at Hertford and at Tunbridge Wells.

The Hertford relay station will transmit on Channel 64 with vertical polarization and will serve some 23 000 people in the Hertford and Ware areas.

The Tunbridge Wells relay station will transmit on Channel 44 with vertical polarization and will serve some 75 000 people in the Tonbridge and Tunbridge Wells areas.

The Hertford and Tunbridge Wells stations are two of the four relay stations being built to fill in the gaps in the coverage of the Crystal Palace BBC-2 transmitter where, due to the screening effect of intervening high ground, BBC-2 reception from the latter station is not satisfactory. A further two stations, to serve the Reigate and Guildford areas, will be brought into service later.

A major export order worth approximately £500 000 is announced for Marconi Doppler navigation systems and airborne computers, to be fitted in the entire Qantas Airways fleet of Boeing 707 aircraft.

Qantas aircraft operate around the world, with many long overwater sectors. On the Sydney-Tahiti-Mexico route to London, for example, where they have already been using Marconi Doppler for over six months, the overwater stages approach 4 000 nautical miles. In such circumstances, the accurate measurements of aircraft ground speed and drift angle provided by the Marconi Doppler navigation equipment, type AD560, have already proved to be of great importance for accurate navigation in the absence of ground aids.

The AD560 Doppler equipment has already been selected by British Overseas Airways Corporation, Air New Zealand, Ghana Airways, East African Airways, Iraqi Airways, Pakistan International Airlines, and a number of overseas government establishments, in addition to being ordered for the Anglo/French Concorde.

Telefunken A.G. of Berlin has produced an experimental radio traffic warning system which has been installed on the Berkhof section of the Hamburg-Hannover motorway.

This system which is designed to warn motorists of emergency conditions on the motorway, transmits the information via an induction loop buried about 1ft below the surface on either side of the motorway, thus restricting the coverage to the area enclosed by the loop.

Mounted at the side of the motorway is a 40W transmitter operating at 70kc/s and a multi-track tape recorder both remotely controlled.

To receive the traffic warning information the drivers' car is fitted with a specially designed receiver tuned to the appropriate transmitter frequency and which is automatically switched on by means of a coded signal.

Alternatively the normal car radio can be fitted with an adapter for the warning service frequency.

If this system is adopted for the German motorways which have a length of some 3 400km, some 800 transmitters will be required with induction loops laid in both directions of travel.

Photain Controls Ltd of Leatherhead Surrey has developed a thermocouple cartridge primarily designed for vehicle brake drum and disk temperature measurement.

The range of uses to which it can be applied is very wide, and includes bearing and shaft temperature measurement; moving wire rope; sheet steel in rolling mills and rotating drums.

The thermocouple junction itself is made of silver alloy brazed into an accurately-machined wear pad which is in turn insulated from the rest of the cartridge by a porcelain tube. A plunger and compression spring ensures that the pad maintains a uniform pressure of 15lb/in^2 on the measured surfaces, and the axial alignment is also kept constant. The entire cartridge fits into a carrier which is attached to a stationary member near the moving surface.

The useful life of a wear pad will depend, of course, upon the nature and condition of the surface, as well as the rubbing speed. Typical life in brake drum temperature measurement is 2 000 miles, but special long-life hard wear pads—giving life in excess of 5 000 miles—are available at very slightly above the cost of the normal pad. The junction is automatically severed when useful life is exhausted, ensuring that incorrect readings are not taken with a failing couple. The need for replacement is shown immediately, and a new cartridge can be inserted within a matter of minutes.

The carriers, plungers and wear pads are turned from brass to BS885, and three thermocouple materials are available — Nickel — Chromium / Nickel-Aluminium; Copper/Constantan; and Iron/Constantan. The Nickel-Chromium/Nickel-Aluminium has a temperature range of between 0 and 800°C; copper/constantan between 0 and 350°C and iron/constantan between 0 and 650°C.

Mullard Ltd has been awarded a contract to supply 3·1 million gold-bonded semiconductor diodes for the peripheral equipment of ICT's new 1900 series of computers.

It is thought to be the biggest single quantity of semiconductor devices ever ordered from a UK manufacturer for professional electronic equipment.

They will be made in Mullard's main diode production centre at Mitcham, Surrey, which has a capacity of over 50 million devices a year. Deliveries will be completed during 1966.

The matrix stacks for the 6μsec core store of the 1902's central processor are also being supplied by Mullard, as are other components including indicator tubes, ferrites, capacitors and other types of diode.

AEI Electronics Ltd has developed a power-off-memory device as one of the latest additions to its range of Logicon solid state static switching equipment. The power-off-memory enables a machine or plant controlled by Logicon switching equipment and stopped by a power failure, to re-start at the exact point reached in the operational sequence at the power breakdown.

This feature safeguards industrial plant where operational time schedules determine product quality; a power failure during a blending process could lead to improper mixing. The use of a power-off-memory device guards against this by ensuring that the process continues from the point of power failure. When the power is interrupted the memory coils switch the circuit to the original sequence reached before interruption.

The Society of Instrument Technology has now adopted new Articles of Association and Rules under which it becomes a qualifying body.

Essentially the revised Articles differ from those in force at present by instituting six fee-paying classes of membership (Member, Companion, Graduate, Associate, Student and Subscriber), and by providing explicitly for (a) examination of candidates for the various classes of membership and (b) admitting as non-Corporate Associates only such candidates as are advanced technicians at least.

The University of Hull announces the inauguration of a department of Electronic Engineering, to which undergraduate students can be admitted in October 1966. The course will take four years to the B.Sc. Special degree, most of the first year and much of the second being occupied with physics, mathematics and subsidiary studies. The Electronic Engineering part of the course will be in two stages:

(a) Electron devices (such as transistors, masers, lasers, etc.) and circuits.
(b) Electronic systems such as telecommunications and radar, industrial automatic-control systems ('automation'), digital computers.

One of the objects of the course is to establish a firm connexion between the fundamental physics on the one hand and the systems-design aspect of electronic engineering on the other hand.

Enquiries about the content of the course may be addressed to the Head of Department, Professor D. A. Bell, M.A., Ph.D., F.Inst.P., M.I.E.E.

Hirst Electronic Ltd, of Gatwick Road, Crawley, Sussex, announce that the name of the company has been changed to Hirst Electric Industries Limited. This change has been made following the Company's greatly increased activities in the electrical field in recent years.

BOOK REVIEWS

Data Smoothing and Prediction in Theory and Practice

By R. B. Blackman. 182 pp. Med. 8vo. Addison-Wesley Publishing Co. Inc. 1965. Price 87s.

MANY scientific disciplines rely on the sagacity of men who, for half a lifetime, quietly contribute to the verification and practical evaluation of new theories. Mr. Blackman is such a man. No light Sunday afternoon reading, this work is a somewhat concise mathematical treatment of the design of data processing schemes, exemplified by active analogue filters and special-purpose digital programmes, and which are intended to reduce the effects of noise or observational errors, with the minimum mutilation of the wanted signal. Systems dealt with are classified as fixed-continuous, variable—for analogue schemes, or fixed continuous, discrete or variable-discrete, for digital-type schemes. For the most part, the treatment deals only with linear systems, but with due regard to overload situations which sometimes arise in practice. In fact, it is this due regard to difficulties, both mathematical and practical, which gives this book its considerable value, based as it is on years of experience at Bell Telephone Laboratories. The author is always meticulous in explaining a point of theory, often showing both valid and invalid approaches to a problem, and taking care to label clearly in parenthesis, any false steps. This method of explanation is highly commendable in a field where intuitive solutions can often be wide of the truth. However, while rigour is upheld, properly-based approximation is encouraged where the theoretical optimum solution is shown to differ little from the more practical approach.

Chapter 1 outlines the basic principles required for the rest of the book, but, as the saying goes, 'it jumps in with both feet', making little concession to those not reasonably well versed in statistical communication theory. Knowledge of the characteristics of the wanted signal greatly assists the optimization of smoothing, and in chapter 2 this feature is discussed assuming a polynomial approximation for signal prediction. Chapters 3 and 4 deal with the optimization of continuous schemes with applications considering Markov noise and white noise, and the use of the Minimax criterion in interception problems. Practical analogue approximations to optimum schemes are described which, although obviously adequate, seem somewhat circumscribed in view of the current scope of active synthesis methods. Discrete schemes are similarly treated in chapters 6 to 8, and digital filter schemes are given in some detail, highlighting the so-called deadband effect, and its avoidance. As related examples, orbital calculations are cited based on the author's Telstar work. Chapter 9 covers methods of smoothing in time-variable continuous and discrete schemes, dealing with special types of non-stationary noise. Finally, chapter 10 is an excellent, if specialized study of the measurement of power spectra. Five appendices provide details needed in the main text, and well over a hundred pertinent references are listed.

This work must surely become a standard reference in its field, and is remarkable for the economy of words used to cover so much ground. Noise always appears however, even in the best systems, signified here perhaps by the solemn indexing of 'piggy-bank', seemingly as a synonym for buffer-store.

K. C. GARNER.

Michael Faraday

By L. P. Williams. 531 pp. Med. 8vo. Chapman & Hall Ltd. 1965. Price 70s.

LET it be said at once—this is a first-class book, an outstanding contribution to the history of science in the 19th century and a biography of Faraday which is likely to stand for all time as the standard work upon his life. Professor Williams devoted more than six years of painstaking research upon its preparation and if the resulting book of over 500 pages does not exactly constitute 'light reading', its author has unquestionably succeeded in his object of revealing for the first time the full dimensions of Faraday's genius.

In his own time, Faraday was widely regarded as an unreliable heretic whose views were not to be trusted. He was the son of a blacksmith, almost entirely self-taught and he had no mathematical ability whatever. 'How could such a man', asked many of his contemporaries, 'make original contributions in the realm of theoretical physics?' Later generations tended to regard Faraday as an intuitive empiricist—and perhaps one can scarcely blame them when even Clerk Maxwell was quite generally regarded with scepticism for many years. In our own times, we have been far too ready to regard Faraday merely as an experimentalist, as one who achieved his results by methods of trial and error —and it is a measure of the success of Professor Williams' work to say that such a view discounts completely Faraday's true brilliance as a scientist and theoretical philosopher of the first rank. Thanks to Prof. Williams, we can now see Faraday in his true place in the mainstream of the history of science.

Faraday was indeed a brilliant experimentalist but what is far too often forgotten is that he was first and foremost a daring theorist, for ever prepared to disagree with established beliefs if the circumstances seemed so to demand. It was indeed his readiness to conceive new theories which led to the stream of new physical facts and laws which flowed from him over a period of more than thirty years and it was his extraordinary experimental talent which enabled him to devise the sensitive laboratory tests by which alone he was able to confirm the particular effects which he was already expecting to find from theoretical considerations.

Faraday's discovery of electromagnetic induction is well known; what is less widely known is that this was no accidental discovery but was the outcome of almost ten years of intermittent theorizing and experiment into Oersted's discovery of electromagnetism in 1820. It is even less widely known that only six months later, Faraday was to compose a short document in which he declared his belief that electromagnetic effects were propagated "like the vibrations upon the surface of disturbed water or those of air in the phenomenon of sound, etc. "I am inclined to think that the vibratory theory will apply to these phenomena, as it does to sound and most probably to light."

Professor Williams quotes in full the fascinating document in which these views were set down but he seems not to have been aware of the romantic circumstances in which it came to light not many years ago. As the full text makes clear, Faraday was not then quite certain that his belief was well founded so he took the rather unusual step of sealing the document in which he expressed his veiws and depositing it with one of the Secretaries of the Royal Society. It remained, locked in the Society's safe, for more than 100 years —until it was opened by Sir William Bragg, the then President of the Royal Society, in the presence of the full Council in June 1937. There for all to see was the evidence that Faraday must henceforth be regarded as the original pioneer of radio-wave propagation!

Among the greatest of Faraday's contributions to theoretical physics must surely be his concept of electric and magnetic fields and Professor Williams shows clearly how this concept developed gradually in Faraday's mind over a period of almost thirty years. Conscious, however, of the mistrust with which his views were regarded, it was not until 1853 that he finally committed himself in his paper 'on the Physical Character of the Lines of Force'.

Clerk Maxwell was among the few who were prepared to accept his theories and by showing how Faraday's ideas were amenable to rigorous mathematical treatment he gave them a cloak of respectability which Faraday's voice alone had been unable to provide.

Maxwell's contribution to field theory is universally acknowledged. What is not so well recognized is the debt which Maxwell himself owed to Faraday. Maxwell freely acknowledged this debt and in a hitherto unpublished letter which he wrote to Faraday in November 1857 he said "As far as I know, you are the first person in whom the idea of bodies acting at a distance by throwing the surrounding medium into a state of constraint has arisen, as a principle to be actually believed in." The very structure, in fact, of Maxwell's "Dynamic Theory of the Electromagnetic Field" was built upon the foundations which had been laid by Faraday and in showing how these foundations were developed Professor Williams has made a notable contribution to the history of 19th century physics.

G. R. M. GARRATT.

Theory of Sampled Data Control Systems

By D. P. Lindorff. 305 pp. Med. 8vo. J. Wiley & Sons. 1965. Price 81s.

THIS book emphasizes the entirely new significance that sampled data systems have assumed over the past decade. One reason for this is that they are now in the mainstream of modern control theory where characterization of systems by matrix finite difference equations is often very convenient. In addition, s.d. control techniques will be increasingly needed as time shared digital computers are used for process control.

Professor Lindorff has tried to give a coherent account of classical theory and modern developments. He assumes his readers have a good background of linear control theory, a working knowledge of the complex variable including contour integration, and familiarity with matrices.

Roughly half the book deals with the pulse transfer function approach to system design and compensation. Many recent and stimuating ideas are concisely presented. Design in the frequency domain is accented. The problems set at the end of each chapter are imaginative and designed to give heuristic insight, though answers are not given. Two minor criticisms; figure 4.25 is missing, and figure 5.1 not really appropriate.

A chapter on statistical design shows how Wiener theory is extended to cover s.d. systems with quadratic costing of the input. Next, state variable ideas are introduced and used as a background for a simple geometric development of Lyapunov functions. Perhaps a comment on their practical limitations would have been useful here.

The last chapter surveys techniques for dealing with non-linear systems. Current developments suggest that Popov's method may soon deserve greater prominence.

The book can be recommended for engineers who wish to keep in touch with modern developments, and is a stimulating textbook for the postgraduate beginner.

R. W. WILDE.

Topology and Matrices in the Solution of Networks

By F. E. Rogers. 204 pp. Demy 8vo. Iliffe Ltd. 1965. Price 45s.

THIS book is suitable as a textbook on circuit theory for university degree and comparable courses in electrical and electronic engineering. Throughout the work the book is copiously illustrated with examples, many being past questions of London University, the I.E.E. and the I.E.R.E. examinations. The book is divided into four main chapters, the first and part of the second being concerned with an elementary account of the topology of electrical networks. These parts are well written and the explanation is very lucid. They alone justify its publication. The remainder of the second chapter and the whole of the third is devoted to the practical solution of network problems and to the introduction, justification and utility of the fundamental network theorems. The method of the Laplace transformation is introduced but the account is rather brief. Apparently, this was the deliberate intention of the Author. Also contained in the third chapter is an excellent development of network duality.

The fourth chapter is concerned with the theory of four-terminal networks and groups of networks. The work of this chapter is a necessary prerequisite to the study of transistor circuit theory and indeed examples from this latter topic are introduced. The book concludes with a set of problems. Brief comments on the solutions and numerical answers are given. It is likely that this excellent little book will become a standard text for many modern courses on network theory.

K. G. NICHOLS.

Electronics of Solids

By W. R. Beam. 633 pp. Med. 8vo. McGraw-Hill Book Co. 1965. Price $12.25

THIS book presents an up-to-date account of, and a wealth of information about, solid state electronics. The first two chapters on crystals and waves in crystals develop the necessary solid state fundamentals. The subsequent eight chapters deal in detail with electronic conduction, semiconductors, semiconductor junctions and devices, superconductivity, dielectrics, magnetism, domain magnetics and quantum electronics. The book finishes with three appendixes devoted to the elastic properties of solids, quantum theory and statistical physics. These appendixes thus provide a compact summary of the essentials of modern physics. It is, however, assumed that the reader has prior knowledge of differential equations vector analysis and basic electromagnetic theory.

The book also includes a discussion of various devices and the possible applications of some of them. It is well written with good cross-referencing. In the discussion good use is made of analogues and physical arguments. The book should prove useful to the final year undergraduate and, in particular, postgraduate students of electronic engineering and applied physics.

The practicing electronic engineer should also find a great deal of useful information in the book. This reviewer is in perfect sympathy with the author when, in the preface, he states that with the ever increasing importance of such solid state devices as transistors, ferrites, masers, superconductors, etc., it is becoming more necessary for the electronic engineer to have a sound knowledge of solid state electronics. This surely is true because it is only then that it is possible for the electronic engineer to have an appreciation of the potentialities and limitations of present day and future solid state devices.

S. S. HAKIM.

Digital Computer Applications to Process Control

Edited by W. E. Miller. 593 pp. Demy 4to. Plenum Press. 1965. Price $17.50

The first International Conference on Digital Computer Applications to Process Control was held at Stockholm in September 1964 at which 22 papers were presented and which are now contained, together with the discussion, in this volume.

The Conference was sponsored by the International Federation for Automatic Control (I.F.A.C.) and the International Federation for Information Processing (I.F.I.P.).

Batteries 2

By D. H. Collins. 543 pp. Med. 8vo. Pergamon Press. 1965. Price £8

This volume contains the proceedings of the Fourth International Symposium on Research and Development in Non-mechanical Electrical Power Sources held at Brighton in September 1964 at which 38 papers were presented.

The Symposium was sponsored by the Inter-Departmental Committee on Batteries.

Microwave Tubes

528 pp. Med. 8vo. Academic Press. 1965. Price 393s.

This volume contains the Proceedings of the Fifth International Congress on Microwave Tubes held at Paris in September 1964 at which some 140 papers were presented.

This Congress was organized jointly by the Société Française des Electroniciens et des Radioelectriciens and the Société Française des Ingénieurs et Techniciens du vide, and was sponsored by the Fédération Nationale des Industries Electroniques.

The first Congress was held in Paris in 1956, followed by the next Congress in London in 1958, then Munich (1960) and The Hague (1962).

Medical Electronics

It is regretted that in the review of this book, which appeared on page 691 of the October issue, the publishers name was not given. It is, DESOER, S. A., 21 Rue Sainte Véronique, Liège, Belgium.

ELECTRONIC EQUIPMENT

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

(Voir page 841 pour la traduction en français; Deutsche Übersetzung Seite 848)

PUSH-BUTTON SWITCH

EMI Sound Products Ltd, Hayes, Middlesex

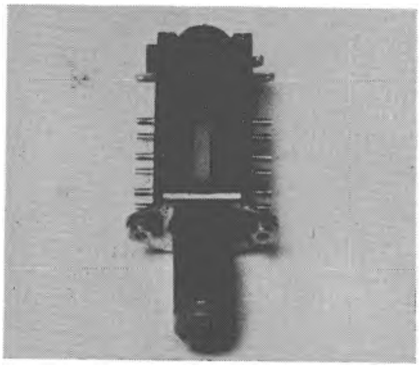
(Illustrated below)

A new illuminated push-button switch designed for use in equipment where space is limited is now available from the Components Division of EMI Sound Products Ltd.

The switch occupies a space 0·70in by 0·57in by 2·15in approximately. It fixes into panels up to $\frac{1}{4}$in thick by means of two 10 B.A. screws.

In spite of its small size it has a contact rating of $\frac{1}{2}$A at 30V d.c. with initial contact resistance 5mΩ which is not greater than 10mΩ after 20 000 operations.

The passivated cadmium plated steel



frame holds high grade s.r.b.p. contact panels. Fixed contacts are silver plated; the moving contact is made from standard silver and the switch has a four changeover break-before-make contact arrangement.

The lamp cap is in a thermo-plastic in a choice of colours (engraving to customers' requirements) and houses a LES bulb. The lamp contacts are independent of the switch contacts.

Two versions of the switch are available: push on-push off coded S 2750 and push-and-release coded S 3750.

EE 88 751 *for further details*

PHOTO-CHOPPERS

Kynmore Engineering Co. Ltd, 19 Buckingham Street, London, W.C.2

(Illustrated above right)

A photo-chopper consists of a light source and photocell operating together as a solid-state switch. In the Kynmore photo-choppers the light sources are neon lamps and the photocells are of the photo-resistive type. They are encapsulated together into complete units. The resultant device, in its single-pole single-throw configuration, can replace the mechanical chopper in a d.c. operational amplifier. It is, however, also suitable for use in pH meters, electrometers, servo-control, speed-sampling and similar systems where a low-noise non-mechani-

cal switch is needed. Two s.p.s.t. devices may also be used together to obtain single-pole double-throw operation; for example, to provide a self-oscillating circuit of established switching frequency for an amplifier with separate modulator and demodulator sections. Several double-pole double-throw models are also available, one of them being specially designed to provide a self-oscillating circuit.

On-resistance of the standard models covers a wide range, to enable the selected model to suit different applications. It may be as low as 50 to 1 000Ω or as high as 200 to 600kΩ. Off-resistance is 100MΩ (for low on-resistance models) or 1 000MΩ (for high on-resistance models). Both low-speed and high-speed models are available. Drive voltage for chopper applications is 120V d.c., drive power being 50mW maximum and switched power 100mW maximum in all cases. Conversion efficiency is 90 per cent at 25°C, 50c/s, with a 1MΩ load and zero source impedance.

Two physical variants of each device are available, designated types A and B. Type A is fitted with axial connecting leads for printed circuit board mounting, while type B has a shielded control lead. Both types are completely encapsulated and internally shielded.

EE 88 752 *for further details*

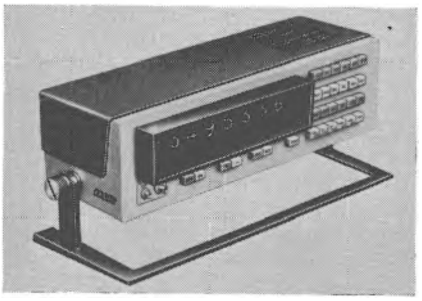
DIGITAL COUNTER

Racal Instruments Ltd, Dukes Ride, Crowthorne, Berkshire

(Illustrated below)

This fully microelectronic high speed digital counter measures only $3\frac{1}{4}$in by $4\frac{1}{2}$in by $12\frac{1}{2}$in and weighs less than 5 lb. It is powered by internal re-chargeable batteries. It has an eight-digit in-line numerical display and may be used for frequency measurements from 0 to 5Mc/s or time and period measurements from 0·1μsec.

Both thin-film and semiconductor com-



ponents have been used in this counter, which is fully microelectronic although naturally some conventional components must be incorporated. Each microelectronic component has been carefully chosen to produce maximum reliability with minimum production cost. The microelectronic design and the extensive use of miniature welded joints ensure that this counter is much more reliable than those of conventional pattern and construction.

Internal or external frequency standards may be used, thus providing any desired degree of accuracy and increasing the possible applications of the counter.

EE 88 753 *for further details*

LEVEL MEASURING SET

Standard Telephones & Cables Ltd, Testing Apparatus Division, Corporation Road, Newport, Monmouthshire

(Illustrated above)

The 74832 level measuring set, now available from STC's Testing Apparatus Division, is a portable general-purpose 75Ω input voltmeter for use between 50kc/s and 800Mc/s. Housed in a robust die-cast metal case the set is particularly suitable for field use, combining the advantages of small size and weight with the ability to operate for long periods from its internal dry cells. For greater convenience in use, no lid is fitted, but the meter is protected by metal flanges.

Three measuring ranges are provided, which are 0 to 60mV, 0 to 600mV and 0 to 3V (r.m.s.) and the scale on the 60mV range is also graduated from −10dB to 0dB in 1dB steps. The response/frequency characteristics is such that, with a constant input signal level from a 75Ω source, the meter reading will not change by more than 0·3dB between 1·0 and 30Mc/s, or by more than 0·6dB between 30 and 500Mc/s.

The set can be used in all climatic conditions, but changes of temperature affect the scale shape and the sensitivity. It is therefore necessary to calibrate at approximately the temperature of operation. However, the change in meter reading at full-scale deflexion with a 10°C change of temperature is only ±0·4dB in the worst case (60mV range); on the 600mV and 3V ranges the figures are ±0·3dB and ±0·2dB respectively.

A peak detector circuit is used for the three voltage ranges, the lowest range including a solid-state d.c. amplifier to give the required sensitivity.

Two additional features greatly extend the scope and versatility of the set. It will function as a beat-frequency indicator for checking the intermediate frequency on radio link systems, and also as a frequency deviation meter for measuring the f.m. deviation by the disappearing carrier method. For beat-frequency indication the second input is added to the main input in the detector circuit and the beat tone is passed through the amplifier to a 'phones jack. For f.m. deviation measurements an external transmission measuring set such as STC's 74156 T.M.E. becomes necessary.

There are two versions of the set, differing only in the type of input connectors (fitted). On both models the 'phones jacks are STC 4101A and the deviation output connectors are Post Office No. 1 plugs.

Only 2mA are taken from a 4·5V supply consisting of three dry cells housed in the case.

EE 88 754 *for further details*

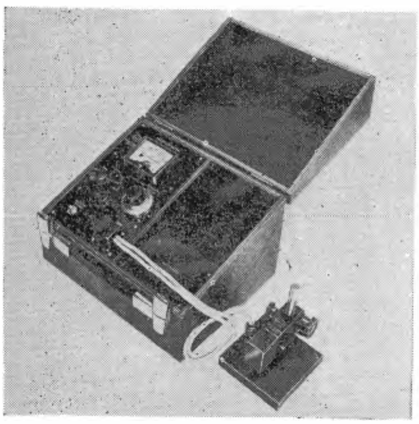
RESISTIVITY BRIDGE

J. A. Radley Research Institute, 220-222 Elgar Road, Reading, Berkshire

(Illustrated below)

This bridge, which gives direct readings of resistivity without preliminary calibration, has been designed primarily for the rapid measurement of the resistivity of semiconductor materials. The instrument can easily be adapted for relative readings of soil resistivity by connecting to an array of earth spikes.

A low frequency square wave is applied to the outer probe contacts, with the measuring potentiometer in series. A voltage is tapped off the potentiometer and applied to the primary of the transformer. The secondary of the transformer is placed in series with the inner probes and 180° out of phase with the primary. The unbalance voltage is amplified and rectified, then applied to a microammeter as a null indicator. By placing suitable range resistors across the potentiometer, direct readings of resistivity can be obtained from the dial when the output voltage is zero.



Series resistors are provided on each range to limit the current in the sample to 1mA, to avoid conductivity modulation.

In operation, the probes are brought into contact with the prepared surface of the sample, the 'on' button is depressed and the potentiometer dial is rotated until a null reading is observed on the meter. The setting of the potentiometer then indicates the resistivity of the sample directly. Very accurate readings are obtainable on all ranges by the use of a 10-turn helical potentiometer with a scale length of 1 000 divisions.

Four ranges are provided covering measurements from 0·01 to 1 000Ω.cm.

EE 88 755 *for further details*

WAVEFORM CONVERTOR

Dawe Instruments Ltd, Western Avenue, London, W.3

(Illustrated below)

Dawe Instruments Ltd has introduced a simple unit for sine wave to square wave conversion. Designated the type 438A Squareshaper, the unit consists of a two-stage transistor feedback circuit

producing a negative square wave of good characteristics when connected to a conventional electronic oscillator such as the Dawe type 440B. No power supply is required.

The Squareshaper is a torpedo-shaped probe measuring 1in diameter by 5in long; it weighs only 4 oz complete with leads. Twin leads are fitted to each end of the probe, each twin lead dividing into 4in single leads fitted with wander plugs on the input side and crocodile clips on the output side.

Amplitude and frequency of the Squareshaper output are governed by the oscillator controls, within the frequency range 1c/s to 1Mc/s for inputs up to 25V r.m.s. Output voltage for 25V r.m.s. input is approximately −30V d.c. on open-circuit, a good negative square wave with a rise time of 50nsec being produced when working into a 2kΩ load. Tilt is 5 per cent at 15c/s and is negligible from 100c/s to 1Mc/s. Overshoot is always less than 1 per cent. The output can be short-circuited without damage.

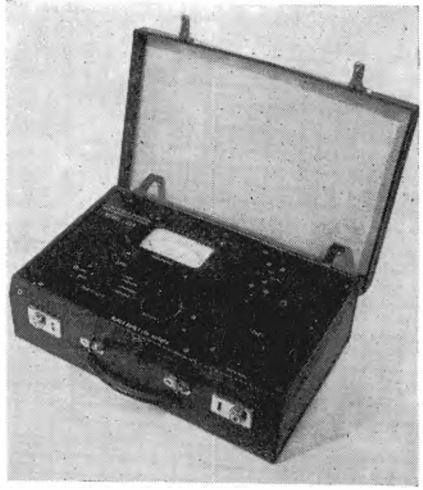
EE 88 756 *for further details*

ACOUSTIC TEST SET

Black Boxes Ltd, Dunsfold Green, Dunsfold, Surrey

(Illustrated above right)

An acoustic test set has been developed by Black Boxes Ltd. The test set comprises an audio frequency white noise



generator, a set of band-pass filters and a power amplifier, built into a compact leather case. It is intended as a noise source for measurements of the transmission, absorbtion and reflection of noise. Eight octave filters cover the audio frequency range. Twelve watts output power is available.

The circuit is fully transistorized and the instrument can be operated from the mains or from an internal battery. A variety of loudspeakers is available.

EE 88 757 *for further details*

PRINTED CIRCUIT TEST PANELS

Spaulding Ltd, Station Road, Edenbridge, Kent

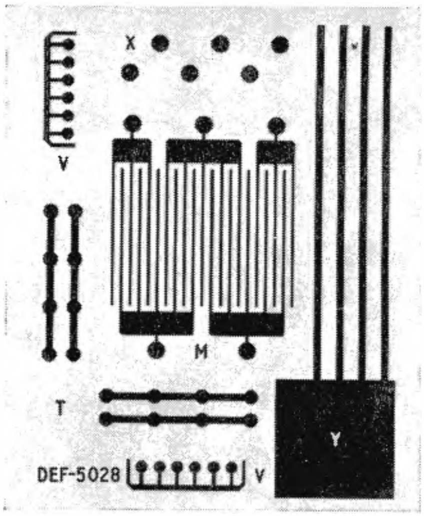
(Illustrated on page 836)

The British Standards Institution recently issued the first British Standard Specification for electric circuits printed on copper-clad plastic laminates (BS. 3888). The Air Ministry has also issued a specification, DEF.5028 Appendix 1, covering aspects of printed circuits.

Spaulding Ltd, the laminate manufacturer, has reproduced the essentials of both official test patterns etched in copper on Spauldite epoxy paper laminate EXXXP845. Both panels are supplied free to electronic engineers and designers who require them, thus saving manufacturers the cost and trouble of manufacturing their own one-off test panels.

The Spaulding test panels are etched on single-sided, one ounce copper 1/16in thickness flame-retardant Spauldite, a grade specially manufactured to meet the needs of advancing electronics technology. Firms needing one or other of the test panels should apply direct to Spaulding Ltd, and they will be despatched complete with a copy of the grade data. The patterns supplied accommodate the main B.S.I. requirements and the relevant requirements of the Air Ministry Specification.

EXXXP845 is an epoxy paper laminate used in the computer and instrument industries and was developed to provide a less costly alternative to epoxy glass laminates while retaining the flame resistance and other superior qualities of the more expensive glass material. It has excellent cold-piercing and general machining and electrical qualities.



The uses of the test patterns include standard peel, solderability and pull-off tests. A section on the DEF.5028 pattern serves as a control for the resistance to mass soldering test and is larger than any unbroken area of conductor normally allowed. It lines up with the $1in^2$ solder test for base materials.

The B.S.I. pattern covers, among other features, heat resistance, solvent resistance, and plating tests.

EE 88 758 *for further details*

MICROWAVE TUNNEL DIODES

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

(Illustrated below)

The first microwave tunnel diode to be manufactured in the U.K. has been announced by Mullard Ltd.

The diode is designed for use in low-noise amplifiers at frequencies up to S-band (2 to 4Gc/s) and samples are available with minimum cut-off frequencies of 6, 8 and 10Gc/s (type numbers AEY13, AEY15 and AEY16 respectively). Negative resistance is typically 50Ω. Peak-point current and peak-point voltage are 2mA and 50mV respectively and the typical series resistance is only 0·75Ω. The valley voltage is 300mV with typical peak/valley ratio of 8.

These characteristics can give an amplifier with a noise figure of 3·8dB compared with about 4·5dB for an amplifier using other tunnel diodes. A lower series resistance, obtained by using a lower etching current ratio, also contributes towards the improvement in noise performance; other factors include the use of doping techniques.

An additional feature of the Mullard

diode is its improved robustness, which is achieved by using a tin/arsenic alloying bead only 15 microns in diameter. This reduces the amount of electrolytic etching required to produce the small junction capacitance essential in a device of this type.

The diode housing measures only 1·5mm diameter by 1·4mm long. It has been designed to give a low inductance by using a thin ceramic annulus of small diameter, and by making short connexions of rectangular cross-section to connect the p-n junction to the housing. The use of rectangular cross-section foil, rather than circular cross-section wire, also increases the mechanical strength of the device.

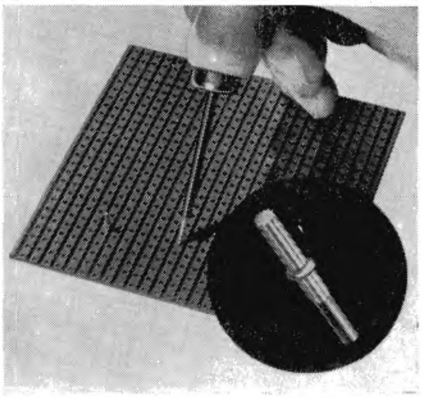
EE 88 759 *for further details*

TERMINAL PIN

Vero Electronics Ltd, South Mill Road, Regents Park, Southampton

(Illustrated below)

A new type of terminal pin, part number 2143, is now available from Vero Electronics Ltd. This pin is for holes of 0·052in diameter and has self-cutting



serrations to hold the pin firmly in the board during soldering operations. A collar prevents the pin from being pushed too far into the board. The Vero pin insertion tool makes the whole operation very simple and fast.

EE 88 760 *for further details*

FILTER-PHASE SHIFTER

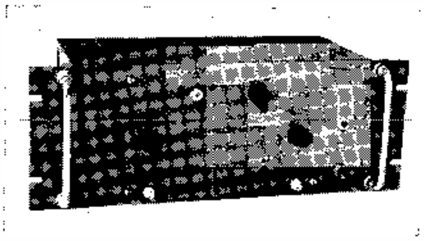
Brookdeal Electronics Ltd, Myron Place, Lewisham, London, S.E.13

(Illustrated above right)

The Brookdeal model FP316 which uses silicon transistor circuits throughout, incorporates a single frequency tuned amplifier and a phase shifter.

The function of the phase shifter section is to provide continuously variable phase control of a sinusoidal signal while giving less than 1dB output voltage variation over its whole phase range. A phase control which covers 160° and a 90° phase change switch give a total phase range of 250°. The unit can be supplied to operate at any frequency specified between 10c/s and 100kc/s and there is also a fine tuning adjustment on the front panel to allow precise alignment of the amplifier with the reference voltage frequency.

The tuned amplifier which precedes



the phase shifter provides filtering of the input signal so that a single frequency, variable phase output signal can be obtained from an input signal composed of several frequencies.

This instrument is designed especially for use with phase-sensitive detection systems where phase correction of a reference signal is required. Typical applications include phase correction of square wave reference voltage or phase correction of a reference voltage which has been derived from a mixer. In either case unwanted signals are rejected by the tuned amplifier before phase shifting the required signal.

The maximum gain is 30dB and the filter frequency can be between 10c/s and 100kc/s, as specified by the customer. The input impedance is 50kΩ and the output impedance less than 1kΩ, the maximum output voltage being 10V peak-to-peak and the phase range 250°.

EE 88 761 *for further details*

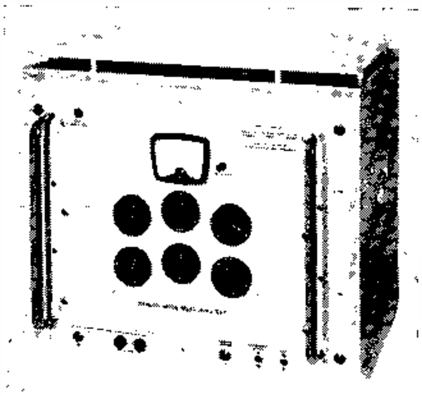
TELEVISION NOISE MEASURING SET

Gresham Lion Electronics Ltd, Twickenham Road, Hanworth, Middlesex

(Illustrated below)

A multi-standard random noise measuring set is now available from Gresham Lion Electronics Ltd. This set, based on a BBC design, has been designed for all 625, 525 and 405 line television networks and fulfils the need for the measurement of random noise in the presence of a television signal. The equipment provides a rapid accurate measurement eliminating human error factors involved in oscilloscope display or other techniques.

The set, known as the M.S.1, meets the latest CCITT/CCIR standards and is particularly suitable for measurements made at television camera channel outputs and video tape recorders during manufacture and operation. It will



simplify measurement techniques in equipment which requires synchronizing pulses. The M.S.1 measures noise in the presence of a standard level signal over the range 20 to 60dB signal/random noise ratio. No ancillary equipment is required.

The circuit is based on the time selection technique and the instrument is provided with an internal calibrating oscillator. Relatively large changes of supply voltage, temperature and valve characteristics cause less than 0·25dB change in power output. A special circuit provides a comparison between the random noise of each head on a video tape recorder. If required, the instrument can provide a signal which can be displayed on an oscilloscope. A difference of 1dB noise performance can be detected in this mode of operation.

EE 88 762 *for further details*

PUSH-BUTTON BRIDGE

The Wayne Kerr Laboratories Ltd, Sycamore Grove, New Malden, Surrey

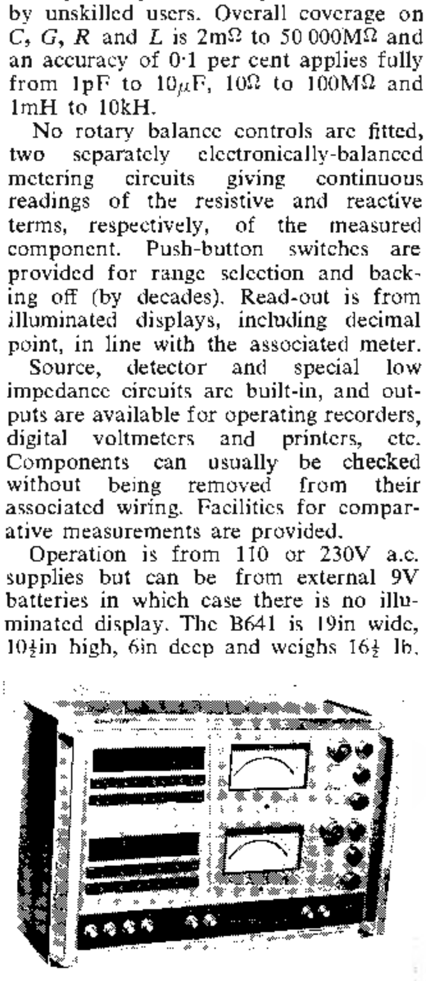
(Illustrated below)

The Autobalance universal bridge type B641 is a new a.f. bridge by Wayne Kerr for the accurate measurement of any component or impedance. A discrimination of 0·01 per cent of maximum on all eight ranges is readily obtained even by unskilled users. Overall coverage on C, G, R and L is 2mΩ to 50 000MΩ and an accuracy of 0·1 per cent applies fully from 1pF to 10µF, 10Ω to 100MΩ and 1mH to 10kH.

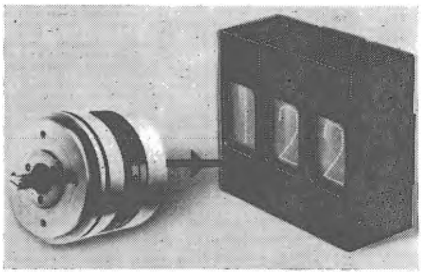
No rotary balance controls are fitted, two separately electronically-balanced metering circuits giving continuous readings of the resistive and reactive terms, respectively, of the measured component. Push-button switches are provided for range selection and backing off (by decades). Read-out is from illuminated displays, including decimal point, in line with the associated meter.

Source, detector and special low impedance circuits are built-in, and outputs are available for operating recorders, digital voltmeters and printers, etc. Components can usually be checked without being removed from their associated wiring. Facilities for comparative measurements are provided.

Operation is from 110 or 230V a.c. supplies but can be from external 9V batteries in which case there is no illuminated display. The B641 is 19in wide, 10½in high, 6in deep and weighs 16½ lb.



EE 88 763 *for further details*



REVERSIBLE DECIMAL DIGITIZERS

Moore Reed (Industrial) Ltd, Woodman Works, Durnsford Road, London, S.W.19

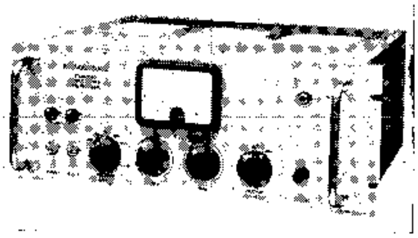
(Illustrated above)

Moore Reed (Industrial) Ltd, has now produced a reversible decimal digitizer for the conversion of mechanical motion to decimal electrical signals without the need for complex decoding equipment.

The digitizers are designed for use in systems displaying, controlling or recording parameters such as machine table or tool positions, flow, pressure, weight, temperature, etc, in process industries.

Produced in a standard size 23 housing, the digitizer employs the M.R. scan and brush technique and is supplied with flying leads.

EE 88 764 *for further details*



E.H.T. POWER SUPPLY

Brandenburg Ltd, 139 Sanderstead Road, South Croydon, Surrey

(Illustrated above)

Brandenburg Ltd is now in production with the first of a range of moderately priced high voltage photomultiplier supplies.

Type PM. 2 500/R provides an output of from 100V to 2·5kV d.c. in five switched ranges with a continuous coarse and fine control. The maximum current available is 5mA. The stability against input variations is 0·03 per cent and the residual ripple is less than 20mV.

A front panel switch with a centre 'off' position is provided for polarity reversal and a dual range output voltmeter is fitted.

The instrument is on a 19in Post Office panel 7in high, for bench or pack use. Standard input 200 to 250V, 50c/s.

EE 88 765 *for further details*

SILVERED MICA CAPACITORS

Johnson, Matthey & Co. Ltd, 73-83 Hatton Garden, London, E.C.1

Johnson, Matthey & Co. Ltd announce the addition of two new miniature types to the Silver Star range of high-stability silvered mica capacitors. The synthetic resin dipped type C5F is now the smallest capacitor in the range. It measures 0·275 × 0·275 × 0·15in (7 × 7 × 3·8mm) and has a capacitance range up to 1 000pF. Type C5E, measuring 0·3 × 0·3 × 0·1in (7·6 × 7·6 × 2·5mm), is the smallest synthetic resin encapsulated capacitor available, and has a maximum capacitance of 680pF. It has the high humidity classification (Class 21, B.S. 2132) associated with all encapsulated Silver Star capacitors.

Both the new types are produced in the E24 series of preferred capacitance values in accordance with B.S. 2488: 1954, and have tolerances down to ±2 per cent (minimum ±2pF). They have working voltages of 50, 200 and 350V, and temperature ranges of −55°C to +100°C. The radial leads are of solder-coated (electroplated) copper and are spaced 0·2in apart.

EE 88 766 *for further details*

CERAMIC DISK CAPACITORS

A. H. Hunt (Capacitors) Ltd, Bendon Valley, Garratt Lane, Wandsworth, London, S.W.18

A. H. Hunt (Capacitors) Ltd has now completed a Ministry of Aviation contract for the development of High 'K' ceramic disk capacitors to operate at temperatures up to 200°C.

The capacitors are of specially developed material to limit the capacitance change over a wide temperature range, −40°C to +200°C, and are protected by a cast resin housing. The capacitance range is 470 to 4 700pF with a tolerance (measured at 100kc/s) of ±20 per cent.

The capacitors have a thickness of 0·7cm; a 1 000pF capacitor has a diameter of 1·75cm and a 4 700pF capacitor a diameter of 3cm.

The insulation resistance at 500V d.c. after 1min is greater than 10^4MΩ at +20°C. The maximum power factor, tanδ, at +20°C is 250 × 10^{-4}.

EE 88 767 *for further details*

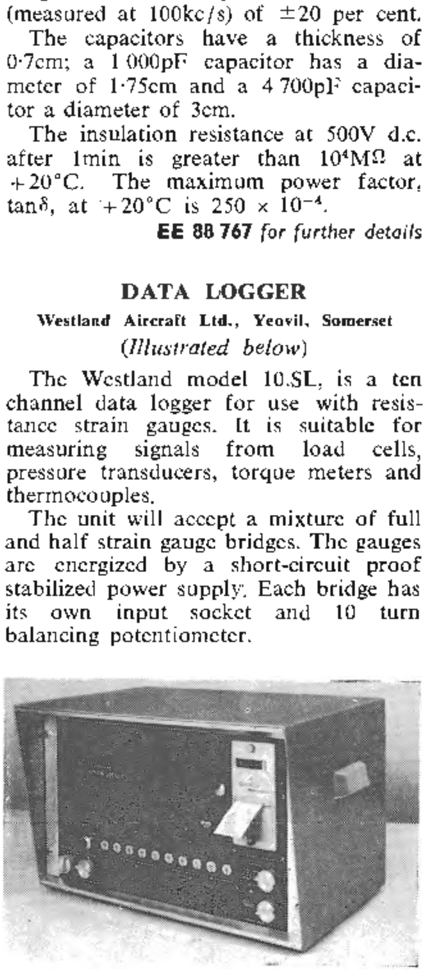
DATA LOGGER

Westland Aircraft Ltd., Yeovil, Somerset

(Illustrated below)

The Westland model 10.SL, is a ten channel data logger for use with resistance strain gauges. It is suitable for measuring signals from load cells, pressure transducers, torque meters and thermocouples.

The unit will accept a mixture of full and half strain gauge bridges. The gauges are energized by a short-circuit proof stabilized power supply. Each bridge has its own input socket and 10 turn balancing potentiometer.



The digital voltmeter has a gain control and an internal reference voltage allowing the digital output to be set directly in microstrain, or other convenient engineering units. A compact paper strip printer provides a permanent record of each measurement and also contains a digital display.

The 10.SL is fully automatic and provides a selection of control modes. It may be left unattended overnight or at weekends to gather and record data from long term experiments.

Other loggers in the new Westland range cater for larger numbers of input channels and different output recorders including paper tape punches.

EE 88 768 *for further details*

MICROSWITCHES

A. F. Bulgin & Co. Ltd, Bye Pass Road, Barking, Essex

(Illustrated below)

Now entering full production is a new reliable range of sub-miniature microswitches. The basic switch, List No. S.800, is enclosed in a moulded phenolic case to B.S. 771 and measures 25/32in long × 0·28in wide × 47/64in high, inclu-

ding button and tags. Fixing is by two transverse 0·089in diameter holes at ½in centres and the switches can be used singly or in groups. The internal mechanism has nickel-silver blades and fine silver contacts. The second high specification switch, List No. S.900 has a moulded case in Araldite X381/2224 to B.S. 771/2782 (high resistant) and gold plated terminals for ease of soldering. Switching in both types is single pole changeover, which can also be wired on-off.

The other models in this range are as follows: "S.800/L', single switch with leaf operator; 'S.800/2/L', two switches with dommon leaf operator, and 'S.801', two switches with a common push-button manual operator.

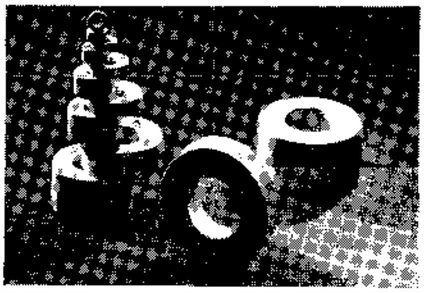
EE 88 769 *for further details*

FERRITE RING CORES

Standard Telephones & Cables Ltd, Magnetic Materials Division, Edinburgh Way, Harlow, Essex

(Illustrated above right)

The new STC range has been specially engineered to offer the designer great flexibility in core selection for any particular application. A logical progression of sizes provides a uniformly spaced range, with two heights of core available for each diameter, giving a 2:1 ratio of effective area. This gives a choice of two cores with the same specific induc-



tance (A_L) or the same effective area (A_e) over most of the range.

The ratio of internal diameter to external diameter of each core has been chosen to give the maximum specific inductance compatible with easy winding by hand or machine. This is based on experience gained over many years of manufacturing and using toroidal inductors, chokes, pulse and wideband transformers.

The new ring cores are available in two ferrite materials. These materials, SA503 and SA601, have already achieved success with the STC range of IEC pot cores and are now providing ring cores with higher permeability and lower losses than was possible previously.

Other features are: No sharp edges. Stove enamelling in a light colour. This makes it unnecessary to tape the core before winding and the white enamel makes it much easier to achieve correct winding distribution of the coil. (STC maintains a toroidal winding service for customers requiring ready-wound cores.)

EE 88 770 *for further details*

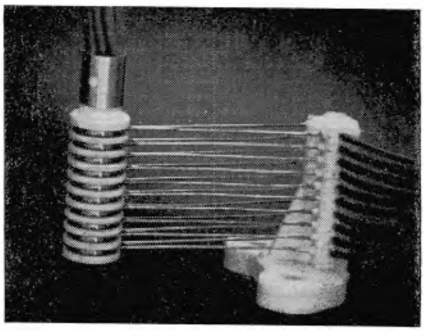
MINIATURE SLIP RINGS

S. Smith & Sons (England) Ltd, Industrial Division, Kelvin House, Wembley Park Drive, Wembley, Middlesex

(Illustrated below)

Developed originally for use in synchroscopes and gyroscopes, these miniature slip ring and brush assemblies are now being made available to the instrument industry generally, and are manufactured by Smiths Industrial Division in three standard ring diameters up to 22 'ways'.

Machined from copper-palladium alloy, the rings are assembled by injection moulding with nylon. This technique has proved much superior to the alternative method of electro-forming, in which the adhesion and homogeneity of the metal can never be assured and for which the choice of metals is limited



to those which may be satisfactorily electro-deposited.

Brush loading is pre-set by the form of the brushes and, as normally supplied, is 2 to 3g per brush, but where special applications require it, this figure can be altered. The slip ring assemblies are suitable for operation over a temperature range of −40°C to +140°C, while the materials used are inherently proof against corrosion and fungus growth.

Depending on the model chosen, the assemblies are free from electrical noise between 750 and 1 500 rev/min. The rating is 1A per way, and the mean contact resistance ·002Ω.

The average length of the assemblies is not greater than 1in. Other dimensions vary with the number of 'ways'.

EE 88 771 *for further details*

ELECTRONIC COUNTERS

Advance Controls Ltd, Imperial Lane, Cheltenham, Gloucestershire

(Illustrated below)

This new range of low-priced electronic counters has been designed to satisfy industrial requirements for totalizing at speeds up to 5 000 per second: it over-

comes the speed limitations of electromagnetic counters without the provision of unnecessary facilities.

Four models are in production having either four or five decades and either numerical (Digitron) or Dekatron display.

Starting and stopping the count and resetting the display to zero is carried out either by switches on the front panel or by remote contacts. A sensitivity control is provided and the input can be connected to front panel terminals or to a socket at the rear.

A 100c/s output, derived from the 50 c/s supply frequency, is provided and can be used for timing purposes.

EE 88 772 *for further details*

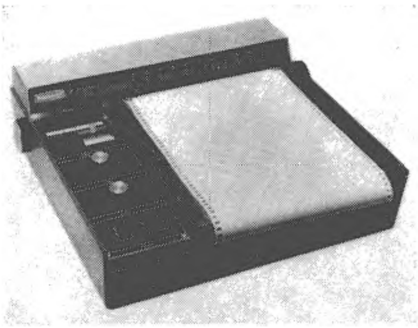
CHART RECORDER

Beckman Instruments Ltd, Glenrothes, Fife, Scotland

(Illustrated on page 839)

A new high-performance ten-inch recorder specifically designed for laboratory use has been introduced by the Scientific and Process Instruments Division of Beckman Instruments.

Combining high performance and convenient design, the new recorder is suitable for a wide range of laboratory applications: gas chromatography, ultra-



violet and infra-red spectrophotometry, pH, etc. Pen response is less than 0·5sec for the full 10in scale. It is accurate to ±0·25 per cent for linear recording, ±0·5 per cent for log recording.

The recorder has been designed for simplicity of operation. All major operating controls are grouped on an easily accessible control panel. Chart speed, range, mode of operation, zero adjustment and pen position can be instantly selected. The ten-inch chart is easy to read and can be advanced manually, inspected instantly, and marked easily.

In addition to the fast pen response and high accuracy, the recorder features; small dead band—less than 0·1 per cent full scale; high rejection of line frequency ripple and common mode signals; instant finger tip selection of seven chart drive speeds; three standard calibrated input spans—1, 10 and 100mV; full scale zero control; and linear or log recording.

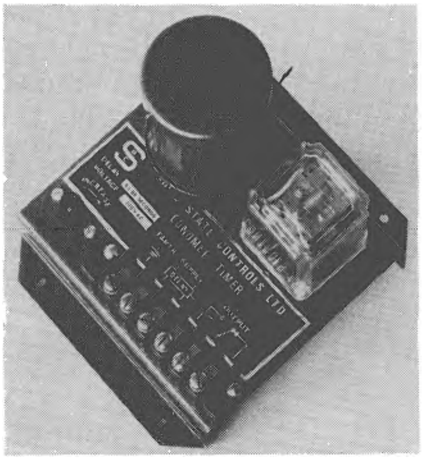
EE 88 773 *for further details*

INDUSTRIAL TIMER

Solid State Controls Ltd, 30-40 Dalling Road, London, N.6

(Illustrated below)

The 'CONOMEE' timer comprises a standard PT/900A fully transistorized plug-in timing module mounted on a chassis mounting bracket together with a reliable plug-in output relay. The simple assembly thus provides a completely integral timing unit which may be conveniently incorporated into any industrial application in the minimum of time. The two standard timing ranges available are 0·2 to 60sec and 1 to 4min, while adjustment over these ranges may



accurately and easily be made by means of a preset control mounted on the bracket.

The timer provides, as standard, one changeover output contact rated at 5A 250V, 50c/s, which operates at the end of the preselected timing period. The four standard operating voltages available are 24, 48, 115 and 250V 40 to 60c/s.

The unit is virtually unaffected by fluctuations in operaitng voltage of ±10 per cent and by temperature variations of −30°C to +60°C. A typical repetitive accuracy of 2 per cent may be expected, while protection is provided against high voltage transients of up to 2 000V for a limited period. The life of the timing unit is up to 100 million operations, if used according to specification.

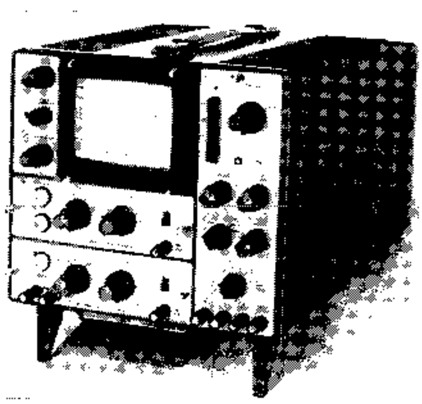
EE 88 774 *for further details*

DOUBLE BEAM OSCILLOSCOPE

Telequipment Ltd. 313 Chase Road, Southgate, London, N.14

(Illustrated below)

The type D53 oscilloscope uses a new double beam rectangular p.d.a. c.r.t.



(operating on the mesh principle) and provides integral delay facilities with automatic connexions for both X and Y amplifiers. Hybrid circuits are used which are 50 per cent solid state.

Main advantage of the new mesh c.r.t., apart from space saving, is that a higher p.d.a. ratio can be used, thus giving a brighter trace with higher sensitivity. The special tube has been manufactured by the M-O Valve Co. Ltd.

A feature of the D53 is a new X amplifier with two sweep delay ranges, up to 5msec and up to 50msec, each with variable uncalibrated delay control. The time-base has 22 calibrated speeds from 5sec/cm to 0·5μsec/cm in 1:2:5 ratio, and there is variable gain control for continuous coverage between ranges. Single shot facility with lock-out is also provided.

Bandwidth of the X amplifier is d.c. to 1Mc/s (−3dB), with sensitivity approximately 100mV/cm at maximum X gain. X expansion is continuously variable up to 10 screen widths, the trace expanding symmetrically about the centre.

Two new Y amplifiers, types CD (ultra-high gain) and HD (wideband) are also available with up to 0·25μsec signal delay. The new oscilloscope also accepts the complete range of interchangeable Y amplifiers already associated with the existing Telequipment S43 and D43 laboratory oscilloscopes.

The D53's rectangular face p.d.a. mesh c.r.t. gives a 6cm high × 10cm wide display for each beam, the tube operating at 8·5kV. Z modulation input is provided.

The graticule is illuminated and detachable, and the hood can be removed for camera fitting.

EE 88 775 *for further details*

HIGH SPEED STORAGE OSCILLOSCOPE

Tektronix UK Ltd, Beaverton House, Station Approach, Harpenden, Hertfordshire

(Illustrated below)

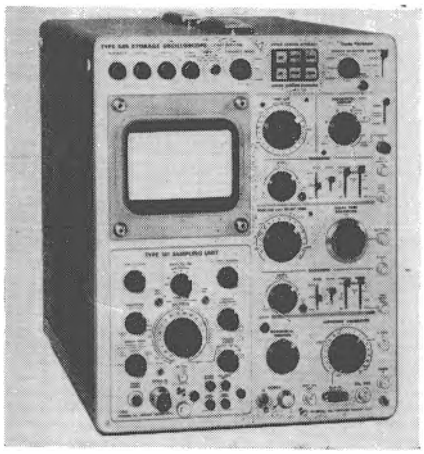
The type 549 oscilloscope is fitted with a fast-writing bistable storage c.r.t. The 6cm × 10cm display area is divided into two 3cm × 10cm independently controlled targets for split-screen applications plus a non-storing 'locate' zone at the left and a 'run-over' storage area at the right.

Independent target control of each half of the screen allows retaining one stored trace while obtaining conventional display, stored display, or enhanced display on the other screen half.

Another highly useful feature is the automatic-erase circuit which can be set to erase the display on either or both halves of the screen at preset intervals. The period between erasures can be varied between 0·5sec and 5sec. A manual erase button is also provided for each half of the screen, as well as provision for remote erase.

The type 549 uses any of the established Tektronix letter-series or 1-series plug-ins for either general-purpose or highly specialized applications in the d.c.- to 30Mc/s area. It also accepts new spectrum analyser units, making it a storage-type spectrum analyser.

Other features of the type 549 include calibrated sweep delay from 1μsec to 10sec, single-sweep, ×5 sweep magnifier, and complete full-passband triggering facilities.



EE 88 776 *for further details*

MEETINGS THIS MONTH

THE INSTITUTION OF ELECTRICAL ENGINEERS

All meetings will be held at Savoy Place, commencing at 5.30 p.m. unless otherwise stated.

Ordinary Meetings

Date: 2 December
Lecture: Appleton's Contribution to Radio Science
By: J. A. Ratcliffe

Control and Automation Division

Date: 8 December
Lecture: Simple New Power-Factor Meter Utilizing Dry Reed Relays
By: P. L. Moreton
Lecture: A Wattmetric Transducer
By: J. Dean and C. E. Lyth
Date: 10 December
Discussion on: The ONC: Better Opportunities for Failure?
Opened by: E. Toner
Date: 14 December
Lecture: Hot-Strip-Mill Adaptive Computer Control
By: D. J. Ray
Date: 16 December
Discussion: Recording Instruments
Opened by: F. R. Axworthy, R. L. Hales and T. Whiteley

Electronics Division

Date: 1 December
Discussion: Telemetry in Medicine
(Joint meeting with the I.E.R.E. at 2 p.m.)
Date: 6 December
Discussion: Acoustic-Wave Amplification
Opened by: E. G. S. Paige and C. P. Sandbank
Date: 8 December Time: 6 p.m.
Discussion: A Review of the Present Situation of Electronic Switching
Opened by: J. A. Lawrence and J. R. Pollard
Date: 9 December
Lecture: Cylindrical Arrays with Electronic Beam Scanning
By: D. E. N. Davies and B. S. McCartney
Lecture: Beam-Positioning Radar Systems Utilizing Continuous Scanning Techniques
By: D. E. N. Davies
Date: 14 December
Lecture: Recent Developments in Nuclear Instrumentation
By: E. H. Cooke-Yarborough
Date: 16 December
Colloquium: Storage: Future Possibilities
(Joint with the I.E.R.E. at 2.30 p.m. All wishing to attend must apply to the Secretary for a ticket. No fee for I.E.E. members: non members £1. 1s.)

Power Division

Date: 1 December
Lecture: Parallel Operation of Transductor-Controlled Silicon-Rectifier Equipment with Water-wheel driven D.C. Generators Supplying an Aluminium Potline
By: N. W. Notani, T. B. Burnett and J. H. Cheetham
Date: 9 December
Lecture: Two-speed Induction Motors Using Fractional-Slot Windings
By: G. H. Rawcliffe and W. Fong
Lecture: Wide-Ratio Two-speed Single-winding Induction Motors
By: W. Fong
Date: 15 December
Lecture: Further Developments in the Design and Performance of High-Voltage Terminal Boxes
By: K. K. Schwarz
Date: 16 December
Discussion: Recording Instruments
Opened by: F. R. Axworthy, R. L. Hales and T. Whiteley

THE INSTITUTION OF ELECTRONIC AND RADIO ENGINEERS

Electro-Acoustics Group

Held at: I.E.R.E. Lecture Room, 9 Bedford Square, W.C.1.
Date: 1 December Time: 6 p.m.
Lecture: Non-linear Effects in Acoustics, Radio and Optics.
By: V. G. Welsby

Joint I.E.E.-I.E.R.E. Medical Electronics Group

Held at: The Institution of Electrical Engineers, Savoy Place, London, W.C.2.
Date: 1 December Time: 2.30 p.m.
Discussion: Telemetry in Medicine.

Television Group

Held at: London School of Hygiene and Tropical Medicine, Keppel Street, London, W.C.1.
Date: 8 December Time: 6 p.m.
Symposium on: Semiconductor Scanning Circuits

Radar Group

Held at: I.E.R.E. Lecture Room, 9 Bedford Square, W.C.1.
Date: 15 December Time: 6 p.m.
Discussion: Reliability of Marine Radar Equipment
Opened by: F. J. Wylie

Joint I.E.E.-I.E.R.E. Computer Groups

Held at: Institution of Electrical Engineers, Savoy Place, London, W.C.2.
Date: 16 December Time: 2.30 p.m.
Colloquium on: Storage—Future Possibilities

Thames Valley Section

Held at: School of Electronic Engineering R.E.M.E., Arborfield, Berkshire
Date: 8 December Time: 7.15 p.m.
Lecture: Frequency Standards.
By: R. Bailey

South Western Section

Held at: University of Bristol, Engineering Laboratories
Date: 8 December Time: 7 p.m.
Lecture: Associative Storage in the Analysis of Multi-parameter Data.
By: I. N. Hooton

East Anglian Section

Held at: University Engineering Department, Trumpington Street, Cambridge.
Date: 9 December Time: 8 p.m.
Lecture: Infra-red Engineering
(Joint meeting with the Cambridge Electronics Section of the I.E.E.)
By: V. Roberts

South Wales Section

Held at: Welsh College of Advanced Technology, Cardiff.
Date: 8 December Time: 6.30 p.m.
Lecture: Modern Aspects of Integrated Circuits.
(Joint meeting with the Institution of Electrical Engineers)
By: M. S. Alderson.

Scottish Section

Held at: Department of Natural Philosophy, The University, Drummond Street, Edinburgh.
Date: 8 December Time: 7 p.m.
Lecture: Nanosecond Exposures with Image Convertor Tubes.
By: K. Coleman
Held at: Institution of Engineers and Shipbuilders, 39 Elmbank Crescent, Glasgow.
Date: 9 December Time: 7 p.m.
Lecture: Nanosecond Exposures with Image Convertor Tubes.
By: K. Coleman.

South Midland Section

Held at: The B.B.C. Club, High Street, Evesham.
Date: 7 December Time: 7 p.m.
Lecture: Some Thoughts, Experiments and Experiences with Transistor Circuits.
By: P. J. Baxandall.

Southern Section

Held at: Farnborough Technical College
Date: 9 December
Lecture: Cosmic Electric Field and Discharges, a New Universal Synthesis
By: C. E. R. Bruce

Yorkshire Section

Held at: Leeds University
Date: 2 December Time: 7 p.m.
Lecture: Machine Tool Control System.
By: K. J. Coppin

Merseyside Section

Held at: Walker Art Gallery, William Brown Street, Liverpool
Date: 15 December Time: 6.30 p.m.
Lecture: Industrial Applications of Electroluminescence.
By: D. Reaney

North Eastern Section

Held at: Institute of Mining and Mechanical Engineers, Neville Hall, Westgate Road, Newcastle
Date: 8 December Time: 6 p.m.
Lecture: Direct Digital Control
By: R. Feldman

THE TELEVISION SOCIETY

Meetings are held in the Conference Suite, I.T.A., 70 Brompton Road, London, S.W.3, and commence at 7 p.m.

Date: 3 December
Lecture: A Survey of Television Receiver Development
By: B. Eastwood, A. Landman, and P. L. Mothersole
Date: 10 December
Lecture: Colour Television Receiver Techniques for SECAM 111A (Followed by a demonstration)

PUBLICATIONS RECEIVED

TAPES FOR INDUSTRY and ELECTRICAL TAPING are two new brochures produced by Sellotape Products Ltd to help manufacturers identify and select the products which best suit their requirements. They give full details of the range of self-adhesive tapes, labels, and tape applying machinery.

'Tapes for Industry' is of a general nature while 'Electrical Taping' relates more closely to the highly specialized requirements of the Electrical Industry.

Both brochures give illustrated examples of typical applications and contain tables in which the characteristics, dimensions, and properties of the tapes are set out.

These brochures are available from Sellotape Products Ltd, Sellotape House, 54-58 High Street, Edgware, Middlesex.

MAGNETIC RECORDING HEADS is a new series of A4 size data sheets describing some of the most up-to-date developments in magnetic recording heads for computers, data recorders, simulators and magnetic drum information storage.

Eleven types of head for digital and analogue applications with up to 33 tracks/in are included in the series. The range includes 'off the shelf' heads for record, playback and write/read applications for use with all types of computer peripheral equipment including IBM hardware.

Further information and binders containing the series of data sheets can be obtained from the Magnetic Recording Head Department, Gresham Lion Electronics Ltd, Lion Works, Hanworth Trading Estate, Feltham, Middlesex.

SERIES 901 IMLOK MANUAL has recently been published by Imhofs and is a 56-page manual on the Series 901 Imlok construction system. Series 901 Imlok is a much improved version of the system previously marketed under the name of Miniature Imlok. Many entirely new parts have now been designed into the system and, in the main, these enable much larger and stronger units to be made, although there is no loss of the facilities previously offered and units down to a 4½in cube can still be produced.

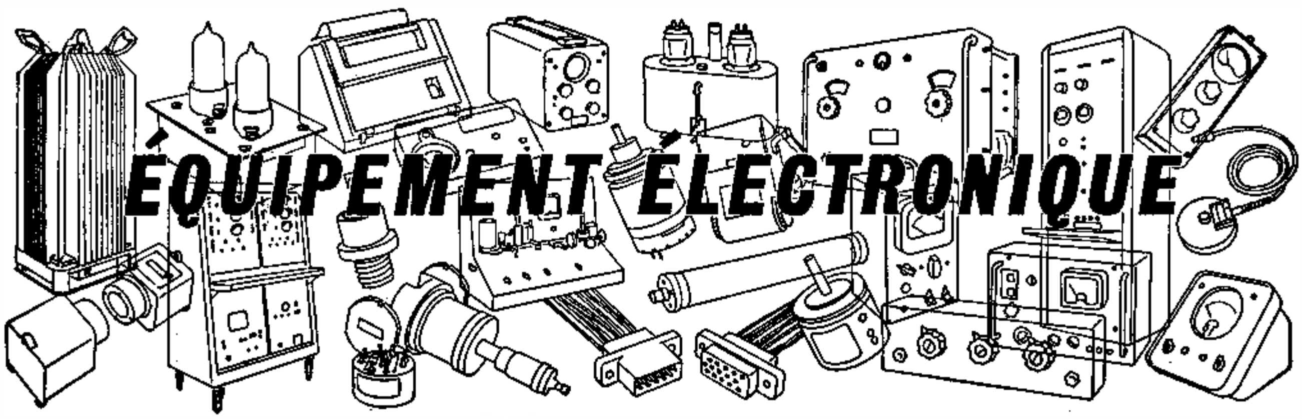
The Series 901 Imlok manual is a comprehensive guide to the system with full dimension information on all the component parts, together with application drawings which clearly show how the various parts are used. There is a guide to designing Series 901 Imlok structures and a guide to using the system which contains easy to follow illustrations showing a typical case being made.

Copies of this manual are available from Alfred Imhofs Ltd, Ashley Works, Cowley Mill Road, Uxbridge, Middlesex.

CAREERS WITH INSTRUMENTS gives an outline of the various careers in instrumentation and details the various educational channels which a student may pursue. Copies of this booklet are available, price 2s. 6d. post free from The Secretary, The Society of Instrument Technology, 20, Peel Street, London, W.8.

ACHIEVING BALANCE BETWEEN CAPACITY AND SALES, a booklet containing the papers given at a Conference on this theme at Solihull, Birmingham, in May last, is now available. The Conference was organized by the British Productivity Council in co-operation with the British Institute of Management and the Centre for Interfirm Comparison in co-operation with Birmingham and Coventry Productivity Associations.

Copies are available, price 15s. post free, from British Productivity Council, Vintry House, Queen Street Place, London, E.C.4.



ÉQUIPEMENT ÉLECTRONIQUE

Une description basée sur des renseignements fournis par les fabricants de nouveaux organes, accessoires et instruments d'essai

Traduction des pages 834 à 839

COMMUTATEUR À BOUTON-POUSSOIR

EMI Sound Products Ltd, Hayes, Middlesex

(Illustration à la page 834)

Un nouveau commutateur à bouton-poussoir éclairé, étudié pour l'emploi dans un matériel exigeant un espace restreint, vient d'être mis au point par la division des composants de la société EMI Sound Products Ltd.

Le commutateur occupe un espace d'environ 0,70 pouce × 0,57 pouce × sur 2,15 pouce. Il peut être fixé dans des panneaux d'une épaisseur de 0,63 cm au moyen de deux vis appropriées.

En dépit de son encombrement réduit, la puissance de contact nominal du commutateur est de ½ A à une tension continue de 30 V, avec une résistance de contact initiale de 5 mΩ qui ne dépasse pas 10 mΩ après 20 000 opérations.

Le cadre d'acier cadmié porte des panneaux de contact en papier imprégné de résine synthétique de haute qualité. Les contacts fixes sont argentés; le contact mobile est en argent ordinaire et le commutateur comprend un dispositif de quatre contacts de permutation de mise hors circuit avant la mise en circuit.

Le capuchon de la lampe est en matière plastique thermique est il est offert dans un choix de différentes couleurs (portant les inscriptions voulues par le client) et loge une ampoule LES. Les contacts de la lampe sont indépendants des contacts de commutation.

Il existe deux versions du commutateur: le modèle code S 2750 à pression marche-arrêt et le modèle code S 3750 à pression marche-déclenchement.

EE 88 751 *pour plus amples renseignements*

RUPTEURS PHOTOÉLECTRIQUES

Kynmore Engineering Co. Ltd, 19 Buckingham Street, London, W.C.2

(Illustration à la page 834)

Le rupteur photoélectrique consiste en une source lumineuse et une cellule photoélectrique fonctionnant en liaison avec la source, sous forme de commutateurs constitués de corps solides. Dans les rupteurs photoélectriques Kynmore, les sources de lumière sont des lampes au néon et les cellules photoélectriques sont du type à résistance photoélectrique. Il sont encapsulés dans des éléments complets. L'instrument qui en résulte est un dispositif unipolaire à une direction pouvant remplacer le rupteur-obturateur mécanique dans un amplificateur opérationnel à tension continue. Il peut cependant, être utilisé également dans les enregistreurs pH, les électromètres, le matériel de contrôle d'asservissement, l'échantillonnage de vitesse et les systèmes analogues exigeant un commutateur non mécanique à faible bruit. Deux dispositifs unipolaires à une direction peuvent être utilisés conjointement pour obtenir le fonctionnement unipolaire à deux directions, par exemple pour réaliser un circuit à oscillation automatique d'une fréquence de commutation déterminée pour un amplificateur avec modulateur et démodulateur indépendants. Il existe également plusieurs modèles unipolaires à deux directions, dont l'un a été spécialement étudié pour fournir un circuit oscillateur automatique.

La résistance des modèles standard couvre une gamme étendue, pour permettre au modèle choisi de s'adapter à différentes applications. La résistance minima peut aller de 50 à 1000 Ω ou au maximum de 200 à 600 kΩ. La résistance hors-circuit est de 100 MΩ (pour les modèles à faible résistance) ou 1000 MΩ (pour les modèles à résistance élevée). Il existe des modèles à grande vitesse aussi bien que des modèles à vitesse réduite. La tension d'entraînement pour les applications de ruption-obturation est de 120 V c.c., la puissance d'entraînement maxima étant de 50 mW et la puissance de commutation de 100 mW au maximum dans tous les cas. L'efficacité de conversion est de 90% à 25° C, 50 Hz, avec une charge de 1 MΩ et une impédance de source nulle.

Deux variantes de chacun de ces modèles sont prévues: les types A et B. Le type A est muni de câbles axiaux de connexion pour montage sur circuit imprimé, tandis que le type B comprend un câble de commande blindé. Les deux types sont entièrement encapsulés et à blindage interne.

EE 88 752 *pour plus amples renseignements*

COMPTEUR NUMÉRIQUE

Racal Instruments Ltd, Dukes Ride, Crowthorne, Berkshire

(Illustration à la page 834)

Ce compteur numérique microélectronique à action rapide ne mesure que 8,25 cm × 11,43 cm × 31,75 cm et pèse moins de 2,26 kg. Il est actionné par batteries rechargeables internes. Il fournit un affichage numérique en ligne de huit chiffres et il peut être utilisé pour la mesure de fréquence de 0 à 5 MHz ou pour la mesure de temps et de périodes à partir de de 0,1 μsec.

Il comprend des composants semi-conducteurs et à pellicule mince et il est entièrement électronique bien que quelques composants classiques y soient naturellement incorporés. Chacun des composants microélectroniques a été choisi avec soin pour produire une fiabilité maxima à un coût de production minimum. La construction microélectronique et l'emploi étendu de joints soudés miniature donnent à ce compteur une sûreté d'emploi beaucoup plus grande que celle des compteurs de type classique.

Des étalons de fréquence internes ou externes peuvent être utilisés, ce qui donne le degré de précision voulu pour n'importe quelle application et augmente d'ailleurs le nombre de ces dernières.

EE 88 753 *pour plus amples renseignements*

INSTRUMENT DE MESURE DU NIVEAU

Standard Telephones & Cables Ltd, Testing Apparatus Division, Corporation Road, Newport, Monmouthshire

(Illustration à la page 834)

L'instrument de mesure du niveau 74832 que vient de réaliser la division des appareils de contrôle de la société Standard Telephones & Cables Ltd, est un voltmètre portatif universel à entrée de 75Ω pouvant être utilisé entre 50 kHz et 800 MHz. Logé dans un robuste coffret métallique moulé en coquille sous pression, le nouvel instrument de mesure convient particulièrement pour les travaux pratiques car il allie les avantages d'un encombrement et d'un poids réduit à la faculté de pouvoir fonctionner pendant de longues périodes de temps à partir de ces cellules sèches internes. Afin de faciliter son emploi, il n'est muni d'aucun couvercle mais l'instrument de mesure peut être protégé par des flasques métalliques.

Il comporte trois gammes de mesure, à savoir 0 à 60 mV, 0 à 600 mV et 0 à 3 V (efficaces) et l'échelle de la gamme de 60 mV est également graduée de −10 dB à 0 dB en degrés de 1 dB. Les caractéristiques réponse/fréquence sont telles que, avec un niveau de signaux d'entrée constants d'une source de 75Ω, l'indication de l'appareil ne changera pas de plus de 0,3 dB entre 1 et 30 MHz ou de plus de 0,6 dB entre 30 et 500 MHz.

L'appareil peut être utilisé dans toutes les conditions climatériques mais les changements de température affectent la forme de l'échelle et la sensibilité. Il est donc nécessaire de l'étalonner à environ la température de fonctionnement. Cependant, le changement d'indication de l'instrument pour une déviation sur la totalité de l'échelle avec un changement de température de 10° C n'est que de ±0,4 dB dans le pire des cas (gamme de 60 mV); dans les gammes de 600 mV et 3 V ces chiffres sont de ±0,3 dB et ±0,2 dB respectivement.

Un circuit détecteur de pointes est utilisé pour les trois gammes de tension, la gamme la plus basse comprenant un amplificateur à courant continu constitué de corps solides qui assure la sensibilité voulue.

Deux caractéristiques supplémentaires étendent considérablement l'utilité et la souplesse d'emploi de l'appareil. Il peut être utilisé comme indicateur de fréquence de battements pour le contrôle de la fréquence intermédiaire sur les systèmes à câbles hertziens, ainsi que comme instrument de mesure de la déviation pour mesurer la déviation de modulations de fréquence par la méthode à porteuse. Pour l'indication de la fréquence de battements, la deuxième entrée est ajoutée à l'entrée principale dans le circuit détecteur et la tonalité de battement est transmise à travers l'amplificateur à un jack téléphonique. Pour la mesure de déviation de la modulation de fréquence, il est nécessaire d'employer un instrument de mesure à transmission externe tel que l'appareil STC 74156 T.M.E.

L'appareil comporte deux versions, qui ne diffèrent que par le type de connecteurs d'entrée dont elles sont munies. Le jack téléphonique des deux modèles est le STC 4101A et les connecteurs de sortie de déviation sont des fiches du type No. 1 de l'administration des téléphones.

La consommation électrique n'est que de 2 mA. Elle est fournie par un bloc d'alimentation de 4,5 V se composant de 3 cellules sèches logées dans le coffret.

EE 88 754 *pour plus amples renseignements*

PONT DE RÉSISTIVITÉ

J. A. Radley Research Institute, 220-222 Elgar Road, Reading, Berkshire

(Illustration à la page 835)

Ce pont, qui fournit des indications directes de résistivité sans étalonnage préalable, a été principalement conçu pour la mesure rapide de la résistivité de matériaux semi-conducteurs. Il peut être adapté aisément pour la lecture relative de la résistivité du sol en le reliant à un réseau de prises de terre.

Une onde carrée à basse fréquence est appliquée aux contacts de sonde extérieurs lorsque le potentiomètre de mesure est en série. Une tension est prise du potentiomètre et appliquée au primaire du transformateur. Le secondaire du transformateur est placé en série par rapport aux sondes intérieures et à 180° de déphasage par rapport au primaire. La tension de déséquilibre est amplifiée et redressée, puis appliquée à un microampèremètre comme indicateur de zéro. En plaçant des résistances de gamme appropriées aux bornes du potentiomètre on peut obtenir des indications directes de résistivité du cadran, la tension de sortie étant nulle.

Des résistances série sont prévues sur chaque gamme pour limiter le courant dans l'échantillon à 1 mA, afin d'éviter la modulation de conductivité.

Durant le fonctionnement, les sondes sont mises en contact avec la surface traitée de l'échantillon, le bouton "marche" est pressé et on fait tourner le cadran du potentiomètre jusqu'à ce que l'on observe une indication de zéro sur l'instrument de mesure. Le réglage du potentiomètre indiquera ensuite directement la résistivité del'échantillon. On peut obtenir des lectures très précises sur toutes les gammes par l'emploi d'un potentiomètre hélicoïdal à 10 tours et à longueur d'échelle de 1 000 divisions.

Quatre gammes sont prévues pour des mesures de 0,01 à 1 000 Ω/cm.

EE 88 755 *pour plus amples renseignements*

CONVERTISSEUR DE FORME D'ONDE

Dawe Instruments Ltd, Western Avenue, London, W.3

(Illustration à la page 835)

La société Dawe Instruments Ltd a réalisé un élément simple pour la conversion des ondes sinusoïdales en ondes carrées. Le nouvel instrument qui porte le nom de "Squareshaper", type 438A, se compose d'un circuit de réaction à transistor et à deux étages produisant une onde carrée négative ayant de bonnes caractéristiques lorsqu'il est relié à un oscillateur électronique classique tel que l'oscillateur Dawe type 440 B. Nulle alimentation n'est exigée.

Le Squareshaper est une sonde en forme de torpille mesurant 2,5 cm de diamètre sur 12,5 cm de long; sont poids, câbles compris, est de 113,2 grammes. Des câbles jumelés sont fixés à chacune des extrémités de la sonde, chaque câble jumelé étant divisé en câbles individuels de 10 cm munis de fiches sur le côté d'entrée et de pinces crocodile sur le côté sortie.

L'amplitude et la fréquence de sortie du Squareshaper sont régies par les commandes d'oscillateur, dans la gamme de fréquence de 1 Hz à 1 MHz pour des entrées maxima de 25 V efficaces. La tension de sortie pour une entrée de 25 V efficace est d'environ −30 V c.c. sur circuit ouvert, une bonne onde carrée négative avec temps de montée de 50 nsec étant produite lorsque on travaille dans une charge de 2 kΩ. L'inclinaison est de 5 pour cent à 15 Hz et elle est négligeable de 100 Hz à 1 MHz. La surmodulation est toujours inférieure à 1%. La sortie peut être court-circuitée sans dommage.

EE 88 756 *pour plus amples renseignements*

CONTRÔLEUR ACOUSTIQUE

Black Boxes Ltd, Dunsfold Green, Dunsfold, Surrey

(Illustration à la page 835)

La société Black Boxes Ltd a réalisé un contrôleur acoustique comprenant un générateur de bruit blanc BF, un jeu de filtres passe-bande et un amplificateur de puissance. Tous ces éléments sont enfermés dans un coffret en cuir compact. Le contrôleur a été conçu comme source de bruit pour la mesure de la transmission, de l'absorption et de la réflexion du bruit. Huit filtres à octave couvrent la gamme de basse fréquence. La puissance de sortie est de douze watts.

Le circuit est entièrement transistorisé et l'appareil peut être utilisé sur courant secteur ou sur batterie intérieure. Une variété de haut-parleurs est prévue.

EE 88 757 *pour plus amples renseignements*

PANNEAUX DE CONTRÔLE À CIRCUIT IMPRIMÉ

Spaulding Ltd, Station Road, Edenbridge, Kent

(Illustration à la page 836)

La British Standards Institution a publié récemment la première norme britannique pour les circuits électriques imprimés sur lamelles en matière plastique à revêtement de cuivre (norme britannique 3888). Le Ministère de l'air a également publié une spécification (DEF.5028, annexe 1) sur les circuits imprimés.

La société Spaulding Ltd qui fabrique des produits laminés, a reproduit l'essentiel des deux diagrammes de contrôle officiel, gravés sur cuivre sur leurs produits laminés en papier d'époxyde Spauldite, EXXXP845.

Les deux panneaux sont fournis gratuitement aux ingénieurs électroniciens et aux réalisateurs qui en auraient besoin, épargnant ainsi aux fabricants les frais et l'inconvénient de devoir fabriquer eux même leurs panneaux de contrôle.

Les panneaux de contrôle Spaulding sont gravés sur un des côtés de la plaque ignifugée Spauldite dont l'épaisseur est de 0,15 cm et qui contient 31,1 g de cuivre. Il s'agit en effet d'un matériau spécialement construit pour répondre au besoin de la technologie de l'électronique, en progression constante. Les sociétés qui pourraient avoir besoin de ces panneaux de contrôle peuvent s'adresser directement à la société Spaulding Ltd et recevront des panneaux complets avec un exemplaire des données de type. Les modèles fournis sont conformes aux conditions de la British Standards Institution, ainsi qu'à la spécification du Ministère de l'air.

Le produit laminé EXXXP845 est un matériau en papier d'époxyde utilisé dans l'industrie des calculatrices et il a été mis au point pour fournir une variante moins coûteuse aux produits laminés en verre d'époxyde, tout en gardant les propriétés ignifuges ainsi que les autres qualités des matériaux en verre plus coûteux. Il a d'excellentes qualités électriques et d'usinage.

Les applications des modèles de contrôle comprennent le contrôle standard de soudabilité et de suspension. Une section du modèle DEF.5028 sert au contrôle de résistance au soudage en masse et elle est plus étendue que n'importe quelle surface ininterrompue de conducteur normalement permise. Elle s'ajoute au contrôle de soudage de 6,45 cm^2 pour les matériaux de base.

La norme B.S.I. couvre entre autre les essais de résistance à la chaleur, de résistance au dissolvant et de galvanoplastie.

EE 88 758 *pour plus amples renseignements*

DIODES TUNNEL MICROONDES

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

(Illustration à la page 836)

La première diode tunnel microondes fabriquée dans le Royaume-Uni a été réalisée par la société Mullard Ltd.

Cette diode a été prévue pour l'emploi dans les amplificateurs à faible bruit à des fréquences allant jusqu'à la bande S (2 à 4 GHz) et il existe des échantillons à fréquence de coupure minima de 6, 8 et 10 GHz (numéros de type AEY13, AEY15 et AEY16 respectivement). La résistance négative typique est de 50 Ω. Le courant de pointe et la tension de pointe sont de 2 mA et 50 mV respectivement et la résistance série caractéristique n'est que de 0,75 Ω. La tension minimum est de 300 mV avec un rapport caractéristique de crête/minimum de 8.

Ces caractéristiques peuvent donner un amplificateur à index de bruit de 3,8 dB, par rapport à 4,5 dB pour un amplificateur utilisant d'autres diodes tunnel. Une résistance série inférieure, obtenue par l'emploi d'un rapport de courant inférieur, contribue également à l'amélioration du bruit. D'autres facteurs comprennent l'emploi des méthodes de dopage.

Un avantage supplémentaire de la diode Mullard est sa grande robustesse, obtenue par l'emploi d'une perle d'alliage d'étain et d'arsenic d'un diamètre de 15 microns seulement. On a pu réduire ainsi la quantité de gravures électrolytiques nécessaires pour produire la faible capacitance de jonction qui est essentielle dans un dispositif de ce type.

Le logement de la diode ne mesure que 1,5 mm de diamètre × 1,4 mm de long. Il a été étudié pour assurer une faible inductance en utilisant un anneau mince en céramique d'un diamètre réduit et en effectuant de petites connexions à section rectangulaire pour relier la jonction p-n au boîtier. L'emploi d'une feuille à section rectangulaire plutôt qu'un fil à section circulaire augmente également la force mécanique du dispositif.

EE 88 759 *pour plus amples renseignements*

BROCHE DE BORNE

Vero Electronics Ltd, South Mill Road, Regents Park, Southampton

(Illustration à la page 836)

Un nouveau type de broche de borne, portant le numéro 2143, est maintenant fourni par la société Vero Electronics Ltd. Cette borne a été étudiée pour des trous de 0,052 pouce de diamètre et elle comporte des cannelures à auto-coupage pour maintenir la broche fermement en place durant les opérations de soudure. Un collier empêche la broche d'être enfoncée trop profondémment dans la plaquette. L'outil d'enfonçage pour broche Vero rend l'ensemble de l'opération très simple et rapide.

EE 88 760 *pour plus amples renseignements*

DÉPHASEUR À FILTRE

Brookdeal Electronics Ltd, Myron Place, Lewisham, London, S.E.13

(Illustration à la page 836)

L'appareil Brookdeal modèle FP316 qui est entièrement muni de circuits à transistors au silicium, comprend un amplificateur accordé à une seule fréquence et un déphaseur.

La fonction du déphaseur est d'assurer le contrôle de phase à variation continue d'un signal sinusoïdal, tout en fournissant une variation de tension de sortie inférieure à 1dB sur toute sa gamme de phases. Une commande le phase couvrant 160° et un commutateur de changement de phase de 90° donnent une gamme de phases totale de 250°. L'appareil peut être fourni pour fonctionner à n'importe quelle fréquence spécifiée entre 10 Hz et 100 kHz et il est également muni sur le panneau frontal d'une commande d'accord pour permettre l'alignement précis de l'amplificateur sur la fréquence de tension de référence.

L'amplificateur accordé qui précède le déphaseur assure le filtrage du signal d'entrée de sorte qu'un signal de sortie à phase variable et à fréquence unique peut être obtenu à partir d'un signal d'entrée composé de plusieur fréquences.

L'instrument a été spécialement étudié pour l'emploi en liaison avec des systèmes de détection sensibles aux phases et exigeant la correction de phase d'un signal de référence. Les applications caractéristiques comprennent la correction de phase d'une tension de référence à onde carrée ou la correction de phase d'une tension de référence provenant d'un mélangeur. Dans les deux cas, les signaux indésirables sont rejetés par l'amplificateur accordé avant le déphasage du signal voulu.

Le gain maximum est de 30 dB et la fréquence de filtrage peut varier entre 10 Hz et 100 kHz, selon la spécification du client. L'impédance d'entrée est de 50 kΩ et l'impédance de sortie inférieure à 1 kΩ, la tension de sortie maxima étant de 10 V de crête à crête et la gamme de phase de 250°.

EE 88 761 *pour plus amples renseignements*

APPAREIL DE MESURE DU BRUIT DE TÉLÉVISION

Gresham Lion Electronics Ltd, Twickenham Road, Hanworth, Middlesex

(Illustration à la page 836)

La société Gresham Lion Electronics Ltd fournit maintenant un instrument de mesure du bruit aléatoire à plusieurs normes. Cet appareil qui est basé sur une réalisation de la BBC, a été prévu pour tous les réseaux de télévision de 625, 525 et 405 lignes et permet de mesurer le bruit aléatoire face à un signal de télévision. L'appareil fournit une mesure rapide et précise et élimine les facteurs d'erreur humaine que comporte l'affichage oscilloscopique ou d'autres méthodes.

Le nouveau contrôleur qui porte la référence M.S.1 répond aux plus récentes normes CCITT/CCIR et il est particulièrement indiqué pour les mesures effectuées aux sorties de canaux de caméra de télévision et sur les magnétophones vidéo en cours de fabrication ou de fonctionnement. Il simplifie les méthodes de mesure dans le matériel exigeant des impulsions de synchronisation. Le M.S.1 mesure le bruit par rapport à un signal de niveau standard dans la gamme de rapport signal/bruit aléatoire de 20 à 60 dB. Nul équipement auxiliaire n'est nécessaire.

Le circuit est basé sur le principe de la sélection de temps et l'instrument est muni d'un oscillateur d'étalonnage interne. Des changements relativement importants de tension d'alimentation, de température et de caractéristique des tubes causent un changement inférieur à 0,25 dB dans la sortie de puissance. Un circuit spécial fournit des éléments de comparaison entre les bruits aléatoires de chacune des têtes sur un magnétophone vidéo. Le cas échéant, l'instrument peut fournir un signal qui peut être affiché sur un oscilloscope. Une différence de bruit de 1 dB peut être décelée par ce mode de fonctionnement.

EE 88 762 *pour plus amples renseignements*

PONT À BOUTON-POUSSOIR

The Wayne Kerr Laboratories Ltd, Sycamore Grove, New Malden, Surrey

(Illustration à la page 837)

Le pont universel Autobalance, type B641, est un nouveau pont BF conç par Wayne Kerr pour la mesure précise de n'importe quel composant ou impédance. Une discrimination de 0,01% au maximum sur toutes les huit gammes est obtenue facilement même par des utilisateurs inexpérimentés. La plage d'utilisation totale sur C, G, R et L est de 2 mΩ à 50 000 MΩ et une précision de 0,1% est pleinement obtenue de 1pF à 10 μF, 10 Ω à 100 MΩ et 1 mH à 10 kH.

L'appareil n'est muni d'aucune commande d'équilibre rotative, deux circuits de mesure à équilibrage électronique fournissant des indications continues des conditions réactives et résistives du composant soumis au contrôle. Des commutateurs à bouton-poussoir sont prévus pour la sélection de gamme et le dépouillement (par décades). L'indication se fait par affichage éclairé comprenant la décimale, en ligne avec l'instrument de mesure connexe.

Les circuits de source, de détection et

ı impédance réduite spéciale sont ncorporés, et des sorties sont prévues ›our l'utilisation d'enregistreurs, de voltnètres numériques, d'imprimeurs, etc. ꞁes composant peuvent habituellement ꞉tre vérifiés sans qu'il soit nécessaire de es détacher de leur bobinage. On peut ꞁgalement effecteur des essais de mesure ꞉omparatifs.

L'appareil est prévu pour une alinentation en tension alternative de 110 ›u de 230 V mais il peut être utilisé sur ›atterie extérieure de 9 V; ans ce cas, ꞉ependant, sans affichage éclairé. L'ap›areil B641 mesure 48,26 cm de large × 26,67 cm de haut × 15,24 cm de profondeur et son poids est de 7,5 kg.

EE 88 763 *pour plus amples renseignements*

CONVERTISSEURS NUMÉRIQUES RÉVERSIBLES

Moore Reed (Industrial) Ltd, Woodman Works, Durnsford Road, London, S.W.19

(Illustration à la page 837)

La société Moore Reed (Industrial) Ltd vient de mettre au point un convertisseur numérique décimal réversible pour la conversion du mouvement mécanique en signaux électriques décimaux, sans qu'il soit nécessaire d'utiliser un matériel de décodage complexe.

Ces convertisseurs numériques ont été prévus pour l'emploi dans les systèmes d'affichage, de contrôle ou d'enregistrement de paramètres tels que la position d'un outil ou d'un banc de machine, le flux, la pression, le poids, la température, etc. dans les industries de processus.

Présenté dans un coffret de format standard numéro 23, le convertisseur numérique utilise la méthode Moore Reed de balayage et d'analyse et il est fourni avec conducteurs mobiles.

EE 88 764 *pour plus amples renseignements*

BLOC D'ALIMENTATION E.H.T.

Brandenburg Ltd, 139 Sanderstead Road, South Croydon, Surrey

(Illustration à la page 837)

La société Brandenburg Ltd produit actuellement les premiers blocs d'alimentation d'une gamme de multiplicateurs photoélectriques à haute tension et à prix modéré.

Le type PM.2500/R fournit une tension continue de 100 V à 2,5 kV en cinq gammes à commutation avec contrôle de réglage continu approximatif et précis. Le courant maximum disponible est de 5 mA. La stabilité par rapport aux variations d'entrée est de 0,03% et l'ondulation résiduelle est inférieure à 20 mV.

Un commutateur de panneau frontal avec position "arrêt" centrale est prévu pour l'inversion de polarité. L'appareil comprend, en outre, un voltmètre de sortie à double gamme.

L'instrument est prévu pour panneau du type de l'administration postale de 48,26 cm × 17,5 cm de haut, pour montage sur banc d'essai ou sur bâti. La tension d'entrée normale est de 200 à 250 V, 50 Hz.

EE 88 765 *pour plus amples renseignements*

CONDENSATEURS À MICA ARGENTÉ

Johnson, Matthey & Co. Ltd, 73-83 Hatton Garden, London, E.C.1

La société Johnson Matthey & Co. Ltd vient d'ajouter deux nouveaux modèles miniature à la gamme Silver Star de capacités au mica argenté à haute stabilité. Le type C5F à revêtement de résine synthétique est actuellement le plus petit modèle de cette gamme de capacités. Il mesure 7 mm × 7 mm × 3,8 mm et sa gamme de capacité s'élève jusqu'à 1 000 pF. Le type C5E qui mesure 7,6 mm × 7,6 mm × 2,5 mm est le plus petit condensateur à enrobage de résine synthétique existant actuellement et sa capacité maxima est de 680 pF. Il a la classification d'humidité élevée (classe 21, B.S. 2132) propre à tous les condensateurs encapsulés Silver Star.

Les deux nouveaux types sont produits dans la série E24 de valeurs préférées de capacité et conformément à la norme britannique 2488 : 1954. Ils ont des tolérances allant jusqu'à ±2 pour cent (minimum ±2 pF). Leurs régimes de travail sont de 50, 200 et 350 V et leurs gammes de températures vont de −55°C à +100°C. Les câbles radio sont en cuivre revêtu de soudure (électroplastie) et il sont espacés de 0,2 pouce l'un de l'autre.

EE 88 766 *pour plus amples renseignements*

CONDENSATEURS À DISQUE EN CÉRAMIQUE

A. H. Hunt (Capacitors) Ltd, Bendon Valley, Garratt Lane, Wandsworth, London, S.W.18

La société A. H. Hunt (Capacitors) Ltd vient de remplir un contrat de fourniture du Ministère de l'aviation pour la production de condensateurs à disque de céramique à "K" élevé, pouvant être utilisés à des températures allant jusqu'à 200° C.

Ces condensateurs sont en un matériau spécialement conçu pour limiter le changement de capacité dans une gamme de température étendue, soit −40° C à +200° C, et ils sont protégés par un revêtement en résine moulée. La gamme de capacités s'étend de 470 à 4 700 pt avec une tolérance (mesurée à 100 kHz) de ±20 %.

L'épaisseur des condensateurs est de 0,7 cm; un condensateur de 100 pt a un diamètre de 1,75 cm et un condensateur de 4 700 pt à un diamètre de 3 cm.

La résistance à l'isolement à une tension continue de 500 V après une minute est supérieure à 10^4 MΩ à +20° C. Le facteur de puissance maximum, c'est à dire tan δ, à +20° C est de 250×10^{-4}.

EE 88 767 *pour plus amples renseignements*

APPAREIL DE CONCENTRATION DE DONNÉES

Westland Aircraft Ltd., Yeovil, Somerset

(Illustration à la page 837)

L'appareil de concentration de données Westland, modèle 10.SL, a été prévu pour l'emploi avec les extensomètres de résistance. Il peut mesurer des signaux de cellules de charge de transducteurs de pression de torsiomètres et de thermocouples.

L'appareil peut recevoir un mélange de ponts d'extensomètres à mi-contrainte et à pleine contrainte. Les extensomètres sont excités par une alimentation stabilisée à l'épreuve des court-circuits. Chaque pont a sa propre douille d'entrée et un potentiomètre d'équilibrage à 10 tours.

Le voltmètre numérique a une commande de gain et une tension de référence interne permettant de régler la sortie numérique directement en microcontraintes ou en tout autre élément technique pratique. Un imprimeur compact à diagramme de papier fournit un enregistrement permanent de chaque mesure et comporte également un affichage numérique.

L'appareil de concentration de données 10.SL est entièrement automatique et fournit un choix de modes de commande. Il peut être laissé sans surveillance pendant la nuit ou durant les week-ends, périodes pendant lesquelles il réunit et enregistre des données d'expériences à longe terme.

D'autres appareils d'enregistrement de données dans la gamme Westland comportent un plus grand nombre de canaux d'entrée et divers enregistreurs de sortie comprenant, entre autres, les perforateurs de bande de papier.

EE 88 768 *pour plus amples renseignements*

MICRORUPTEURS

A. F. Bulgin & Co. Ltd, Bye Pass Road, Barking, Essex

(Illustration à la page 838)

Une nouvelle gamme de microrupteurs subminiature est actuellement en cours de production. L'interrupteur de base, qui porte la référence S800, est enfermé dans un boîtier en matière phénolique moulée conforme à la norme britannique 771. Il mesure 1,98 cm de long × 0,70 cm de large × 1,86 cm de haut, bouton et cosses compris. Le montage se fait sur deux ouvertures transversales de 0,089 pouce de diamètre à entre-axes de 1,25 cm et les interrupteurs peuvent être utilisés en groupe ou individuellement. Le mécanisme interne comporte des lames de nickel-argent et des contacts en argent fin. Le deuxième rupteur, qui porte la référence S900, est enfermé dans un boîtier moulé en araldite, X381/2224, selon la norme britannique 771/2782 (haute résistance) et il est muni de bornes dorées qui facilitent le soudage. La commutation des deux types se fait par permutation unipolaire pouvant être également bobinée sur "arrêtmarche".

Cette gamme comprend en outre les modèles suicants: S.800/L, rupteur unique avec opérateur à feuille; S.800/2/L, deux rupteurs avec opérateur à feuille commun, et S.800 deux rupteurs avec opérateur manuel commun à bouton-poussoir.

EE 88 769 *pour plus amples renseignements*

NOYAUX D'ANNEAUX EN FERRITE

Standard Telephones & Cables Ltd, Magnetic Materials Division, Edinburgh Way, Harlow, Essex

(Illustration à la page 838)

Cette nouvelle gamme STC a été spécialement conçue pour offrir aux réalisateurs une grande variété de choix de noyaux pour n'importe quelle application. Une progression logique de dimensions fournit une gamme à intervalles uniformes avec deux longueurs de noyaux pour chaque diamètre, donnant ainsi un rapport de zone effective de 2:1. On obtient ainsi un choix de deux noyaux avec la même inductance spécifique (A_L) ou la même surface effective (A_e) sur la plus grande partie de la gamme.

Le rapport du diamètre intérieur au diamètre extérieur de chaque noyau a été étudié pour assurer l'inductance spécifique maxima compatible avec la facilité du bobinage à la main ou à la machine. Cette méthode est basée sur l'expérience acquise au cours de nombreuses années de construction et d'emploi d'inducteurs toroïdaux, de bobines d'arrêt, de transformateurs à impulsions et à large bande.

Les nouveaux noyaux d'anneaux sont fournis en deux matériaux de ferrite. Ces matériaux, le SA503 et le SA601, ont déjà été employés avec succès dans la gamme STC des noyaux IEC et fournissent maintenant des noyaux d'anneaux d'une perméabilité plus élevée et à pertes inférieures qu'auparavant.

Les nouveaux noyaux ne comportent pas de bords à arêtes vives. Ils sont émaillés en une couleur claire. L'émail blanc facilite en effet le bobinage correct de la bobine. (La société STC offre un service de bobinage toroïdal pour les clients exigeant des noyaux bobinés d'avance.)

EE 88 770 *pour plus amples renseignements*

BAGUES COLLECTRICES MINIATURE

S. Smith & Sons (England) Ltd, Industrial Division, Kelvin House, Wembley Park Drive, Wembley, Middlesex

(Illustration à la page 838)

Conçus à l'origine pour l'emploi dans les synchroscopes et les gyroscopes, ces assemblages miniature de bagues collectrices et de balais sont maintenant offerts à l'ensemble de l'industrie des instruments et sont fabriqués par la Division Industrielle de la société S. Smith & Sons (England) Ltd en trois diamètres standard de bagues jusqu'à 22 directions.

Usinées en alliage de cuivre et de palladium, ces bagues sont assemblées par moulage à injection avec le nylon. Cette méthode s'est avérée nettement supérieure à la méthode d'électroformage, qui ne permet jamais de s'assurer de l'adhésion et de l'homogénéité du métal et qui limite le choix des métaux à ceux pouvant être déposés par voie galvanique de manière satisfaisante.

La charge des balais est déterminée par la forme de ces derniers. Normalement, elle est de 2 à 3 g par balai mais en cas d'application spéciale, cette charge peut être modifiée. Les assemblages de bagues collectrices conviennent pour l'utilisation dans une gamme de température de $-40°$ C-$+140°$ C, tandis que les matériaux utilisés sont à l'épreuve de la rouille et de la moisissure.

Les assemblages sont exempts de bruit électrique entre 750 et 1 500 tours/min, selon le modèle choisi. L'indice nominal est de 1 A par direction et la résistance de contact moyenne est de 0,002 Ω.

La longueur moyenne des assemblages ne dépasse pas 2,5 cm. D'autres dimensions varient suivant le nombre de directions.

EE 88 771 *pour plus amples renseignements*

COMPTEUR ÉLECTRONIQUE

Advance Controls Ltd, Imperial Lane, Cheltenham, Gloucestershire

(Illustration à la page 838)

Cette nouvelle gamme de compteurs électroniques à bas prix a été conçue pour répondre à la demande de compteurs industriels pouvant totaliser à des vitesses de 5000 coups par seconde. Il obvie aux limitation de vitesse des compteurs électromagnétiques sans qu'il soit besoin de le munir de dispositifs supplémentaires.

Quatre modèles sont actuellement en cours de production. Ils sont soit à quatre soit à cinq décades et avec affichage numérique (Digitron) ou Dékatron.

La mise en action ou l'arrêt du comptage ainsi que le réenclenchement de l'affichage à zéro s'effectue soit au moyen de commutateurs sur le panneau frontal soit par contact à distance. Une commande de sensibilité est en outre prévue et l'entrée peut être reliée à des bornes du panneau frontal ou à une douille à l'arrière.

Une sortie de 100 Hz, provenant de la fréquence d'alimentation de 50 Hz, peut être fournie. Elle peut être reliée à l'entrée de sorte que les compteurs puissent être utilisés pour les opérations de minutage.

EE 88 772 *pour plus amples renseignements*

ENREGISTREUR À DIAGRAMME

Beckman Instruments Ltd, Glenrothes, Fife, Scotland

(Illustration à la page 839)

La Division des Instruments Scientifiques et de Processus de la société Beckman Instruments Ltd vient de créer un nouvel enregistreur de 254 mm à haute performance, spécialement conçu à l'intention des laboratoires.

Alliant une conception pratique à un rendement élevé, le nouvel enregistreur peut être utilisé pour un grand nombre d'applications de laboratoire, telles que la chromatographie des gaz, la spectrophotométrie aux rayons ultra-violets et infra-rouges, l'enregistrement pH, etc. La réponse de plum est inférieure à 0,5 sec pour la totalité de l'échelle de 254 mm. Sa précision est de $\pm 0{,}25$ % pour l'enregistrement linéaire et de $\pm 0{,}5$ % pour l'enregistrement logarithmique.

Le nouvel enregistreur a été conçu de manière fort pratique et son fonctionnement s'en trouve considérablement simplifié. Toutes les commandes principales sont groupées sur un panneau de contrôle d'un accès facile. La vitesse d'avance du diagramme, la distance, le mode de fonctionnement, le réglage d[e] zéro et la position de la plume peuve[nt] être choisis instantanément. Le di[a]gramme de 254 mm est d'une lectu[re] aisée et on peut le faire avancer à [la] main. Il peut être contrôlé instantan[é]ment et marqué facilement.

En plus de la réponse de plum[e] rapide et de sa grande précision, l'e[n]registreur se caractérise par une band[e] "morte" réduite (inférieure à 0,1 % d[e] la totalité de l'échelle), un rejet élev[é] de la fréquence d'ondulation de lign[e] et des signaux de mode commun, [le] choix instantané par le bout du doi[gt] de sept vitesses d'avance du diagramm[e,] trois portées d'entrée normales étalon[n]ées: 1, 10 et 100 mV, contrôle du zér[o] sur la totalité de l'échelle et enregistr[e]ment linéaire ou logarithmique.

EE 88 773 *pour plus amples renseignemen[ts]*

MINUTERIE INDUSTRIELLE

Solid State Controls Ltd, 30-40 Dalling Roa[d,] London, N.6

(Illustration à la page 839)

La minuterie "CONOMEE" compren[d] un module de minutage standard à fiche[s] et entièrement transistorisé, PT/900A monté sur un support de châssis avec u[n] relais de sortie également à fiches. C[et] assemblage d'une grande simplicit[é] fournit un élément de minutage faisan[t] corps avec le reste de l'appareil e[t] pouvant être incorporé facilement [à] n'importe quelle application industriell[e] dans un minimum de temps. Les deu[x] gammes de minutage standard vont d[e] 0,2 à 67 et de 1 à 4 min, cependant qu[e] le réglage dans ces deux gammes peu[t] être effectué rapidement et avec préci[]sion au moyen d'une commande prérégl[ée] montée sur le support.

La minuterie comprend, en temps qu[e] composant normal, un contact de sortie [à] permutation d'une puissance nominale [de] à 5 A, de 250 V, 50 Hz, qui entre e[n] action au terme de la période de minu[]tage préselectionnée. Les quatre tension[s] de fonctionnement standard sont: 24, 48 115 et 250 V, 40 à 60 Hz.

L'appareil est pratiquement insensibl[e] aux variations de tension de régime d[e] ± 10 pour cent et aux variations de température de $-30°$C à $+60°$C. Une précision à répétition caractéristique que de 2% peut être obtenue, tandis que la protection est assurée contre les phénomènes transitoires de haute tension jusqu'à 2000 V pendant une période de temps limitée. La durée de l'appareil de minutage est de 100 millions d'opérations, s'il est employé suivant les spécifications.

EE 88 774 *pour plus amples renseignements*

OSCILLOSCOPE À DOUBLE FAISCEAU

Telequipment Ltd, 313 Chase Road, Southgate, London, N.14

(Illustration à la page 839)

L'oscilloscope type D53 comporte un tube cathodique de post-accélération rectangulaire à double faisceau de conception nouvelle. Il fonctionne selon le principe du montage polygonal et comprend un dispositif à retard avec connexions automatiques pour amplificateurs X et Y. Il est muni de circuits hybrides constitués à 50 % de corps solides.

Le principal avantage du nouveau tube cathodique à mailles, en dehors de l'économie d'espace qu'il assure, est de permettre l'emploi d'un plus grand rapport de post-accélération, donnant ainsi une trace plus brillante et une plus grande sensibilité. Le tube spécial a été conçu et réalisé par la société M-O Valve Co. Ltd.

L'oscilloscope D53 de distingue également par son nouvel amplificateur X à deux gammes de balayage à retard de 5 msec et 50 msec respectivement, pourvues toutes deux d'une commande de retard variable non-étalonné. La base de temps comprend 22 vitesses étalonnées de 5 sec/cm à 0,5 µsec/cm d'un rapport de 1:2:5. Il y a, en outre, une commande de gain variable pour le contrôle continu entre les gammes. Enfin, l'appareil est muni d'un dispositif monocourse à verrouillage.

La largeur de bande de l'amplificateur X s'étend du courant continu à 1 MHz (−3 dB) et la sensibilité est d'environ 100 mV/cm pour un gain maximum X. L'étalement X est à variation continue jusqu'à 10 largeurs d'écran, la trace s'étalant de manière symétrique autour du centre.

Il existe également deux nouveaux amplificateurs Y, types CD (à gain ultra-élevé) et HD (à large bande) avec retard de signal jusqu'à 0,25 µsec. Le nouvel oscilloscope peut recevoir la gamme complète d'amplificateurs Y interchangeables, ces derniers étant déjà prévus pour les oscilloscopes Telequipment S43 et D43 pour laboratoires.

Le tube cathodique à mailles de post-accélération à face rectangulaire de l'oscilloscope D53 donne une image de 6 cm de haut et 10 cm de large pour chaque faisceau; il fonctionne sur 8,5 kV. Une entrée de modulation Z est prévue par ailleurs.

La mire de points est éclairée et détachable, le capot pouvant être enlevé pour le montage d'une caméra.

EE 88 775 *pour plus amples renseignements*

OSCILLOSCOPE D'EMMAGASINAGE À ACTION RAPIDE

Tektronix UK Ltd, Beaverton House, Station Approach, Harpenden, Hertfordshire

(*Illustration à la page 839*)

L'oscilloscope type 549 est muni d'un tube cathodique à emmagasinage bistable et à écriture rapide. La surface d'affichage de 6 cm sur 10 cm est divisée en deux cibles de 3 cm × 10 cm à commande indépendante pour les applications à écran divisé et elle comprend une zone de non-emmagasinage à gauche et une zone d'emmagasinage de révision à droite.

La commande de cible indépendante d'une moitié de l'écran permet de conserver une trace emmagisinée tout en obtenant une image classique, une image emmagasinée ou une image rehaussée sur l'autre moitié de l'écran.

L'oscilloscope comprend une autre particularité des plus utiles consistant en un circuit d'effacement automatique pouvant être réglé de maniere à effacer l'image sur l'une ou l'autre ou sur les deux moitiés de l'écran à intervalles préréglés. La période de temps entre deux "effacements" peut être variée entre 0,5 sec et 5 sec. L'appareil est également pourvu d'un bouton d'effacement manuel pour chaque moitié de l'écran, ainsi que d'un dispositif d'effacement à distance.

L'oscilloscope type 549 utilise les fiches Tektronix à série de lettres ou à série L pour les applications générales ou pour les applications particulières dans la gamme allant du courant continu à 30 MHz. Il peut, d'autre part, être muni des nouveaux analyseurs de spectres qui en font un analyseur de spectres à emmagasinage.

L'appareil se distingue, par ailleurs, par son dispositif de retard de balayage étalonné de 1 µsec à 10 sec, par son amplificateur de balayage ×5 à balayage unique et par son déclencheur complet de passe-bande totale.

EE 88 776 *pour plus amples renseignements*

Résumés des Principaux Articles

Un compteur à trois voies et à vitesse réduite pour la mesure des particules alpha par P. B. Roberts

Résumé de l'article aux pages 784 à 787

L'auteur décrit un instrument à trois voies pour la mesure des matières radioactives émettant des particules alpha dont l'activité s'élève à quelques centaines de pico-curies. Cet appareil est entièrement transistorisé et il peut être utilisé sur batterie ou sur courant secteur. Les circuits de comptage peuvent être utilisés pour l'enregistrement d'événements produits à des vitesses réduites.

Stabilisation de tension par tube de déclenchement à cathode froide par E. P. T. Atkinson et C. F. Hill

Résumé de l'article aux pages 788 à 792

Cet article nous documente sur l'emploi des tubes de déclenchement à cathode froide pour la stabilisation des petits blocs d'alimentation redressée à sortie variable. Le comportement et les limites du circuit sont analysés ainsi que les facteurs régissant la conception des circuits à cathode froide. Il est fait mention du Z806W dont les caractéristiques d'allumage sont stables et à tolérence serrée. L'article conclut par une illustration graphique des valeurs de circuit pour les tensions de sortie de la gamme de 200V à 360V.

Un stimulateur à transistors propre à de nombreuses applications par W. J. Bannister et R. H. Kay

Résumé de l'article aux pages 793 à 795

Il s'agit ici d'un stimulateur à transistors d'une grande souplesse d'emploi conçu pour la recherche neurophysiologique. L'appareil se compose de trois éléments à part:

(a) L'élément à retard, actionné par la forme d'onde de la base de temps de l'affichage oscilloscopique final, fournissant une impulsion ou une paire d'impulsions aux éléments de déclenchement (b) et (c). Ces déclencheurs, fixés à la base de temps, peuvent être "retardés" indépendamment, c'est à dire calés de manière stable à n'importe quelle position le long de l'axe X du tube cathodique. Les déclencheurs restent à la même position spatiale le long de la base de temps, quelle que soit la vitesse de la base de temps.

(b) L'élément "Tetanus", fournissant des trains d'impulsions périodiques au déclencheur (c), lorsqu'il est amorçé par une impulsion de déclenchement de l'élément (a). Les trains d'impulsions peuvent également être mis en oeuvre indépendamment de l'élément (a) par bouton-poussoir manuel.

(c) L'élément de sortie du stimulateur, fournissant une impulsion rectangulaire d'amplitude, de durée et de polarité controlées lorsqu'elle est déclenchée par les éléments (a) ou (b) et également lorsqu'elle est déclenchée à la main par bouton-poussoir. Les sorties d'impulsions sont normalement isolées par un transformateur à double blindage de conception appropriée qui retient la forme d'impulsions rectangulaires pour des impulsions d'une durée maxima de 10msec. Une sortie de tension continue est également prévue pour ces impulsions ainsi que pour des impulsions d'une plus longue durée.

Un changeur de fréquence pour compteurs à grande vitesse par P. Adby et E. A. Edis

Résumé de l'article aux pages 796 à 802

Le changeur de fréquence dont il est question dans cet article porte la limite de fréquence d'u compteur-type de 100MHz à 900MHz, tout en conservant la précision de mesure fondamentale a compteur. Sa conception rend superflue la commande d'accord continu que l'on trouve normalemer dans les instruments analogues, de sorte qu'on peut mesurer rapidement une série de fréquences en n choisissant que l'une des six gammes de fréquence à commutation. Le contrôle du niveau de sorti et d'entrée permet de mesurer un signal inconnu sans ambiguité aucune. A l'exception d'un tub d'entrée intermédiaire, le circuit se compose entièrement de semi-conducteurs et comporte diverse caractéristiques intéressantes qui sont décrites en détail.

Transistors à effet de champ pour amplificateur biologique par R. E. Webb

Résumé de l'article aux pages 803 à 805

Après un court exposé sur les conditions requises pour l'amplification de potentiels bioélectrique: l'auteur donne quelques indications sur l'emploi des transistors à effet de champ servant à ce but. . effectue des comparaisons entre les tubes électrométriques à vide et les transistors à effet de champ e donne des détails d'un amplificateur à impédance élevée. Le bruit de circuit et la manière de l réduire sont également étudiés brièvement.

Un émetteur électrocardiographique portatif par A. R. Atkins et E. B. Lock

Résumé de l'article aux pages 806 à 808

Cet article décrit un émetteur électrocardiographique portatif utilisé pour les recherches physio logiques dans des conditions de température et d'humidité élevées, telles que dans la chambr climatérique ou souterraine dans les mines. Cet appareil est utilisé pour détecter et transmettre . un récepteur éloigné des signaux d'impulsions électrocardiaques d'un sujet sous observation. C'es un instrument de petit format fixé à une ceinture portée par le patient.

On a également conçu un transducteur pour la détection d'impulsions électrocardiaques de patient en transpiration.

Le récepteur a été réalisé de manière à ce qu'on puisse choisir un émetteur (d'un choix de quatre émet teurs) permettant à l'observateur de compter le ta uxdes battements de coeur de patients en écoutan les signaux au moyen d'écouteurs ou en reliant le récepteur à un électrocardiographe et en effectuan un enregistrement des signaux reçus.

Réseaux à varistors utilisés comme éléments de calcul analogique par W. G. P. Lamb et M. L. Stones

Résumé de l'article aux pages 809 à 811

Les auteurs décrivent des réseaux simples de résistance linéaire et non linéaire, conçus pour produir un élement qui se conforme avec précision à un rapport courant/potentiel de $i = Kv^n$ dans lequel n s trouve entre un et trois. Une application se rapportant à la diffusion de ions d'hydrogène dans le fe est également indiquée.

Multivibrateurs astables à transistors pour éléments logiques universels par G. H. Stearman

Résumé de l'article aux pages 812 à 817

L'auteur analyse les caractéristiques que devrait avoir un circuit multivibrateur astable utilis comme générateur à minuterie avec des éléments logiques universels. Il signale les défauts de circui classique, de plusieurs variantes de ce dernier ainsi que ceux de certains circuits peu conventionnels. Il présente enfin quelques nouvelles variantes de montage qui semblent répondre davantage aux conditions exigées que les circuits existant précédemment.

Amplificateur de puissance BF à réaction de recouvrement par A. R. Bailey

Résumé de l'article aux pages 818 à 821

La conception des amplificateurs de puissance classiques BF pour la reproduction à haute fidélite est impropre à l'usage de nombreux laboratoires. En particulier la distorsion est souvent excessive sur une partie du spectre de fréquence, et la stabilité n'est pas toujours maintenue sur les charges capacitives. Cet article décrit la conception et le comportement d'un amplificateur de laboratoire dont la spécification est nettement meilleure que celle habituellement acceptée.

On obtient ainsi une puissance nominale totale sur toute la gamme de fréquence acoustique à des niveaux de distorsion inférieurs à 0,1%. De plus, l'amplificateur garde une stabilité à toute épreuve pour toutes les valeurs de charge et sa faculté de rétablissement après une surcharge est parfaitement satisfaisante.

Le circuit de filtrage à diode Zéner-bobine d'arrêt par J. M. Diamond

Résumé de l'article aux pages 822 à 825

Les avantages du circuit à diode Zéner-bobine d'arrêt sont examinés dans cet article. Ces avantages comprennent la protection contre les court-circuits, la régulation et la réduction du bourdonnement. Lorsqu'il est suivi d'un régulateur dégénératif classique, le circuit à diode Zéner-bobine d'arrêt assure une forme de protection simple et à toute épreuve contre les court-circuits et rend possible des valeurs extrêmement basses de régulation de ligne et de sortie de bourdonnement. Le circuit est analysé en ce qui concerne la dissipation des divers éléments, y compris l'élément passe-bande du régulateur dégénératif dans les conditions de charge et de ligne les plus défavorables.

Un commutateur électronique à répétition pour la recherche des isotopes par E. van der S. Lotz

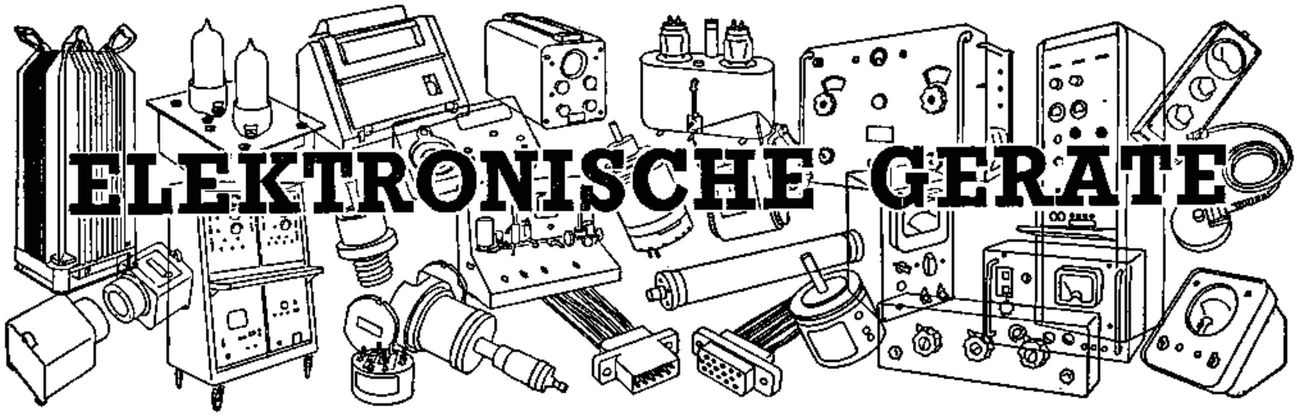
Résumé de l'article aux pages 825 à 827

Pour les expériences effectuées avec des isotopes de courte durée et pour certains contrôles statistiques, il est parfois nécessaire d'utiliser un compteur d'impulsions nucléoniques à plusieurs reprises successives et pendant des périodes de temps relativement courtes. L'auteur décrit un commutateur électronique au moyen duquel ces compteurs peuvent être utilisés de manière répétée pendant des périodes prédéterminées de 1, 2, 4, 8 ou 16 secondes. La séquence de minutage exigée est produite par un tube à décharge luminescente de 12 cathodes, actionné par des impulsions provenant d'ondes sinusoïdales redressées et non filtrées de 50cps. On peut utiliser le redressement à pleine onde et à demi-onde, avec commutation d'un mode de redressement à un autre au moyen d'une petite tétrode à gaz.

Un isolateur de stimulus HF à haute efficacité pour usages électrophysiologiques par W. T. Catton et R. G. Farrier

Résumé de l'article àla page 828

Les impulsions rectangulaires d'un stimulateur classique ont été utilisées pour la modulation par écran d'un oscillateur HF à tétrode et à couplage électronique fonctionnant à 5MHz. Les impulsions HF qui en résultérent furent transmises, à travers un câble coaxial à faible pertes, à un détecteur toutes-ondes accordé dont la sortie de tension continue était isolée de la masse. D'excellents résultats furent enregistrés en ce qui concerne la conversion de puissance et on a pu constater qu'une puissance beaucoup plus élevée peut être fournie par rapport aux isolateurs des modèles entérieurs.



Beschreibung neuer Bauelemente, Zubehörteile und Prüfgeräte auf Grund der von Herstellern gemachten Angaben

Übersetzung der Seiten 834 bis 839

Drucktastenschalter

EMI Sound Products Ltd, Hayes, Middlesex

(*Abbildung Seite 834*)

Der Geschäftsbereich Bauelemente der EMI Sound Products Ltd hat einen Leuchtdrucktastenschalter für Verwendung in Geräten, in denen nur begrenzter Raum zur Verfügung steht, herausgebracht.

Der Schalter nimmt etwa 17,8 × 14,5 × 54 mm Raum ein, und Befestigung auf bis zu 6,35 mm dicken Platten erfolgt mittels zweier Schrauben 10BA.

Trotz der kleinen Abmessungen kann der Schalter mit 0,5 A bei 30 V— belastet werden; er hat einen ursprünglichen Kontaktwiderstand von 5 mΩ, der nach 20 000 Vorgängen nicht höher als 10 mΩ ist.

Der passivierte kadmierte Stahlrahmen enthält die Kontaktplatten aus hochgradigem Hartpapier. Die Festkontakte sind versilbert, die beweglichen Kontakte bestehen aus Standardsilber, und der Schalter hat vier Umschalter-Folgekontakte.

Es gibt zwei Ausführungen des Schalters: Drucktasten mit Arretierung S2750 und Drucktasten ohne Arretierung S3750.

EE 88 751 *für weitere Einzelheiten*

Lichtelektrischer Zerhacker

Kynmore Engineering Co. Ltd, 19 Buckingham Street, London, W.C.2

(*Abbildung Seite 834*)

Ein lichtelektrischer Zerhacker besteht aus einer Lichtquelle und einer Fotoquelle, die als Festkörperschalter zusammenarbeiten. In den Kynmore-Zerhakkern dienen Neonlampen als Lichtquellen, und Fotowiderstände werden als Zellen eingesetzt. Sie werden zusammen in komplette Bausteine verkapselt und können in der einpoligen Ausschalterausführung den mechanischen Zerhacker in einem Gleichstrom-Funktionsverstärker ersetzen. Sie sind aber auch für Verwendung mit PH-Messern, Elektrometern, Servoreglern, Geschwindigkeits-Sampling und ähnlichen Systemen geeignet, in denen geräuscharme, nichtmechanische Schalter benötigt werden. Zwei der einpoligen Ausschalter können auch zusammen als einpoliger Umschalter Verwendung finden, z.B. in einem selbsterregten Oszillator mit feststehender Schaltfrequenz für einen Verstärker mit getrennten Modulator- und Demodulatorteilen. Es gibt auch verschiedene zweipolige Umschaltermodelle, von denen eins besonders für eine Schaltung mit Selbsterregung entwickelt wurde.

Widerstand im Ein-Zustand gibt es in grosser Auswahl, um den verschiedenen Anwendungsmöglichkeiten gerecht zu werden. Er kann so niedrig wie 50...1000 Ω oder so hoch wie 200...600 kΩ sein. Der Widerstand im Aus-Zustand ist 100 MΩ (für Modelle mit niedrigem Widerstand im Ein-Zustand) und 1 GΩ (für im Ein-Zustand hochohmige Modelle). Die Modelle sind mit niedriger sowie hoher Geschwindigkeit zu haben. Die Treibspannung bei Verwendung als Zerhacker ist 120 V—, die Treibleistung höchstens 50 mW und die Schaltleistung in allen Fällen maximal 100 mW. Der Wirkungsgrad der Umsetzung ist bei 25°C, 50 Hz, mit einem 1-MΩ-Abschluss und Quellenimpedanz von Null etwa 90 Prozent.

Jeder der Bausteine kommt mechanisch in zwei Ausführungen, als Typ A und B. Typ A hat Axialanschlüsse für gedruckte Schaltungen, Typ B hat eine abgeschirmte Steuerleitung. Beide Typen sind völlig gekapselt und intern abgeschirmt.

EE 88 752 *für weitere Einzelheiten*

Digitalzähler

Racal Instruments Ltd, Dukes Ride, Crowthorne, Berkshire

(*Abbildung Seite 834*)

Der vollmikroelektronische Digital-Schnellzähler wiegt bei Abmessungen von etwa 82,5 × 114 × 317,5 mm weniger als 2,25 kg und wird mit internen aufladbaren Batterien betrieben. Er hat eine 8stellige Einzeilen-Ziffernanzeige und kann für Frequenzmessungen von 0...5 MHz oder Zeit- und Periodenmessungen von 0,1 μs ab eingesetzt werden.

In diesem Zähler finden sowohl Dünnfilm- wie Halbleiterelemente Verwendung, und er ist vollmikroelektronisch, obwohl natürlich einige konventionelle Bauelemente eingebaut werden. Jedes der mikroelektronischen Bauelemente wurde sorgfältig auf höchste Zuverlässigkeit mit niedrigsten Fertigungskosten hin ausgesucht. Der mikroelektronische Aufbau und die weitgehende Anwendung von geschweissten Miniaturverbindungen gewährleisten, dass dieser Zähler viel zuverlässiger ist als diejenigen herkömmlicher Anordnung und Konstruktion.

Man kann interne oder externe Frequenznormale verwenden, um dem Zähler jede gewünschte Genauigkeitsklasse zu geben und dadurch seine Anwendungsmöglichkeiten noch zu erweitern.

EE 88 753 *für weitere Einzelheiten*

Pegelmessgerät

Standard Telephones & Cables Ltd, Testing Apparatus Division, Corporation Road, Newport, Monmouthshire

(*Abbildung Seite 834*)

Der STC-Geschäftsbereich Prüfgeräte kann jetzt das Pegelmessgerät 74832, ein tragbares 75-Ω-Eingangsvoltmeter für Allgemeinverwendung im Frequenzbereich 50 kHz...800 MHz, liefern. Es ist in einem robusten Metallspritzgussgehäuse untergebracht und besonders für den Aussendienst geeignet. Ausser kleinen Abmessungen und niedrigem Gewicht hat das Gerät den Vorteil, dass es für lange Betriebszeiten aus den internen Trockenelementen gespeist werden kann. Zur Erleichterung der Benutzung ist das Messinstrument ohne Deckel gebaut, dagegen aber durch Metallflansche geschützt.

Das Gerät hat drei Messbereiche von 0...60 mV, 0...600 mV und 0...3 V (eff), und die Skala für den 60-mV-Bereich ist auch mit einer Teilung von −10 dB... 0 dB in 1-dB-Intervallen versehen. Bei konstantem Eingangssignal von einer 75-Ω-Quelle ändert sich die Anzeige auf Grund des Frequenzganges zwischen 1,0 und 30 MHz um nicht mehr als 0,3 dB und zwischen 30 und 500 MHz um nicht mehr als 0,6 dB.

Das Gerät kann zwar unter allen klimatischen Bedingungen arbeiten, jedoch beeinflussen Temperaturänderungen Skalenform und Empfindlichkeit. Die Eichung muss daher ungefähr bei der Betriebstemperatur erfolgen. Die Anzeige ändert sich jedoch bei Temperaturänderungen von 10°C und Vollausschlag im schlimmsten Fall (60-mV-Bereich) nur um ±0,4 dB. Für die 600-mV- und 3-V-Bereiche sind die entsprechenden Werte ±0,3 dB bzw. 0,2 dB.

Für die drei Spannungsbereiche wird eine Scheitelwertdetektorschaltung mit einem Festkörper-Gleichstromverstärker zur Erzielung der für den niedrigsten Bereich erforderlichen Empfindlichkeit benutzt.

Anwendungsbereich und Vielseitigkeit des Gerätes werden durch zwei weitere Eigenschaften erweitert. Es kann als Schwebungsfrequenzanzeiger für die Überprüfung der Zwischenfrequenz in Richtfunkstrecken sowie als Frequenzhubanzeiger zum Messen des FM-Hubes nach dem Verfahren mit verschwindendem Träger funktionieren. Für die Schwebungsfrequenzanzeige wird das zweite Eingangssignal dem Haupteingang in der Detektorschaltung zugefügt und der Schwebungston durch den Verstärker zu einer Kopfhörerklinke gespeist. Für die FM-Hubmessung muss ein externer Pegelmessatz wie z.B. der STC 74156 T.M.E. zur Verfügung stehen.

Das Gerät ist in zwei Ausführungen lieferbar, die sich nur durch ihre Eingangsanschlüsse unterscheiden. Die Kopfhörerklinken an beiden Modellen sind STC-Typ 4101A, und für das Hubausgangssignal finden Stecker Nr. 1 der britischen Postbehörde Verwendung.

Der aus drei Trockenelementen im Gehäuse bestehenden 4,5-V-Stromversorgung werden nur 2 mA entnommen.

EE 88 754 *für weitere Einzelheiten*

Messbrücke für spezifischen Widerstand

J. A. Radley Research Institute, 220-222 Elgar Road, Reading, Berkshire

(*Abbildung Seite 835*)

Diese Messbrücke gestattet Direktablesung des spezifischen Widerstandes ohne Voreichung und ist in erster Linie für das schnelle Messen des spezifischen Widerstandes von Halbleitermaterialien bestimmt. Das Gerät kann ohne Schwierigkeiten durch Anschluss an ein Erdsondenfeld für relative Messungen des spezifischen Bodenwiderstandes eingesetzt werden.

Den mit dem Messpotentiometer in Serie liegenden Aussensondenkontakten wird eine niederfrequente Rechteckwelle zugeführt. Eine vom Potentiometer abgegriffene Spannung wird an die Primärwicklung des Übertragers gelegt, deren Sekundärwicklung mit den Innensonden in Serie geschaltet ist und gegenüber der Primärwicklung eine um 180° verschobene Phase hat. Die Diagonalspannung wird verstärkt und gleichgerichtet an einen Mikrostrommesser als Nullanzeige gelegt. Durch Parallelschalten geeigneter Bereichwiderstände zum Potentiometer kann man direkte Ablesungen des spezifischen Widerstandes von der Potentiometerskala erhalten, wenn die Ausgangsspannung Null ist.

In jedem Bereich sind zur Vermeidung veränderlicher Leitfähigkeit Serienwiderstände vorgesehen, die den Strom in der Probe auf 1 mA begrenzen.

In Betrieb werden die Sonden mit der vorbereiteten Oberfläche der Probe in Berührung gebracht, die "Ein"-Taste gedrückt und die Potentiometerskala so lange gedreht, bis das Messgerät Null anzeigt. Die Potentiometereinstellung erlaubt dann direktes Ablesen des spezifischen Widerstandes der Probe. Durch Anwendung eines 10gängigen Wendelpotentiometers mit 1000 Teilungen der Skala werden sehr genaue Anzeigen erzielt.

Die vier Bereiche geben einen Messumfang von 0,01...1000 Ω.cm.

EE 88 755 *für weitere Einzelheiten*

Wellenformwandler

Dawe Instruments Ltd, Western Avenue, London, W.3

(*Abbildung Seite 835*)

Dawe Instruments Ltd hat ein einfaches Gerät für die Umwandlung von Sinuswellen in Rechteckwellen herausgebracht. Der Squareshaper (Rechteckformer) 438A besteht aus einer zweistufigen Transistor-Gegenkopplungsschaltung, die bei Anschluss an einen herkömmlichen elektronischen Oszillator, z.B. den Dawe-Typ 440B, eine negative Rechteckwelle mit guten Kenndaten erzeugt. Es ist keine Stromversorgung erforderlich.

Der Squareshaper ist eine torpedoförmige Sonde von 25,4 mm Durchmesser und 127 mm Länge, die mit Zuleitungen nur etwa 114 g wiegt. Jedes Ende der Probe hat zweiadrige Zuleitungen, die über die letzten 10 cm in Einzelschnüre mit Wandersteckern am Eingang und Krokodilklemmen an der Ausgangsseite unterteilt sind.

Amplitude und Frequenz des Squareshapers werden im Frequenzbereich 1 Hz...1 MHz und bis zu 25 V_{eff} durch die Regelorgane des Oszillators bestimmt. Für ein Eingangssignal von 25 V_{eff} ist die Ausgangsspannung im Leerlauf etwa −30 V—; eine gute negative Rechteckwelle mit 50 ns Anstiegszeit wird bei Abschluss mit 2 kΩ erzeugt. Die Dachschräge ist bei 15 Hz 5 Prozent und zwischen 100 Hz und 1 MHz vernachlässigbar klein; das Überschwingen ist in allen Fällen unter 1 Prozent. Der Ausgang kann ohne Schaden kurzgeschlossen werden.

EE 88 756 *für weitere Einzelheiten*

Tonfrequenz-Testgerät

Black Boxes Ltd, Dunsfold Green, Dunsfold Surrey

(*Abbildung Seite 835*)

Ein von Black Boxes Ltd entwickelte Schalltestgerät besteht aus einem Tonfrequenzgenerator für weisses Rauschen einem Satz von Bandfiltern und einem Endverstärker, die in einer kompakten Ledertasche untergebracht sind. Das Gerät ist als Rauschquelle für das Messen der Übertragung, Schluckung und Reflexion von Geräuschen gedacht. Seine acht Oktavenfilter überstreichen den Tonfrequenzbereich. Am Ausgang stehen 12 W Leistung zur Verfügung.

Die Schaltung ist volltransistorisiert und das Gerät kann entweder vom Netz oder internen Batterien gespeist werden. Eine Auswahl von Lautsprechern ist lieferbar.

EE 88 757 *für weitere Einzelheiten*

Testleiterplatten

Spaulding Ltd, Station Road, Edenbridge, Kent

(*Abbildung Seite 836*)

Der britische Normenausschuss (British Standards Institution) hat vor kurzem die erste Norm für auf kupferkaschierten Plastikschichtstoffen (BS.3888) gedruckte Schaltkreise veröffentlicht. Das Luftfahrtministerium behandelt auch im Pflichtenblatt DEF 5028 Anhang 1 Druckschaltungsgesichtspunkte.

Als Schichtstoffhersteller stellt Spaulding Ltd Platten ihres Spauldite Epoxyd-Papierschichtstoffes EXXXP845 her, in deren Kupferbelag die wesentlichen Züge der amtlichen Testanordnungen eingeätzt sind.

Beide Leiterplatten werden an Elektroniker und Konstrukteure, die sie benötigen, ohne Berechnung abgegeben und sparen somit Fabrikanten die mit der Erstellung von Einzeltestplatten verbundenen Kosten und Mühen.

Die Spaulding-Testleiterplatten werden auf einseitig kaschiertes, 1,6 mm starkes, flammenverzögerndes Spauldite gätzt, das in dieser Qualität besonders entsprechend den Anforderungen der fortschrittlichen Elektronik-Technologie hergestellt wird. Firmen, die die eine oder andere Testplatte benötigen, erhalten sie auf Anforderung von Spaulding Ltd zusammen mit einem Datenblatt der Spauldite-Qualität. Die Leiterzüge dieser Platten entsprechen den Hauptanforderungen des Normenausschusses und den infragekommenden Anforderungen des Pflichtenblattes des Luftfahrtministeriums.

EXXXP845 ist ein Epoxyd-Papierschichtstoff, der in der Datenverarbeitung und in Messgeräten Anwendung findet; er wurde als preisgünstigere Alternative zu den Epoxyd-Glasfaserschichtstoffen entwickelt, behält jedoch den Flammenwiderstand und die anderen überlegenen Eigenschaften des teureren Glasfasermaterials. Es lässt sich ausgezeichnet kalt lochen sowie anderweitig bearbeiten und hat hervorragende elektrische Eigenschaften.

Vorgesehen sind u.a. Leiterzüge für Schäl- und Abreisstests sowie Lötbarkeit. Ein Abschnitt der Leiterzüge nach DEF.5028 dient der Kontrolle der Festigkeit gegen Massenlötungen und ist grösser als normalerweise für ununterbrochene Leiterflächen erlaubt ist. Das entspricht dem Quadratzoll-Löttest für Trägerstoffe.

Auf der BSI-Platte vorgesehen sind u.a. Tests auf Wärmewiderstand und Lösungsmittelwiderstand sowie ein Plattierungstest.

EE 88 758 *für weitere Einzelheiten*

Mikrowellen-Tunneldioden

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

(Abbildung Seite 836)

Die erste in Grossbritannien hergestellte Mikrowellen-Tunneldiode wurde von Mullard Ltd angekündigt.

Die Diode ist für Anwendung in rauscharmen Verstärkern bei Frequenzen bis zum S-Band (2...4 GHz) bestimmt, und Muster stehen mit einer Mindestgrenzfrequenz von 6,8 und 10 GHz (Typ AEY13, AEY15 bzw. AEY16) zur Verfügung. Ein typischer negativer Widerstand ist 50 Ω. Spitzenpunktstrom und -spannung sind 2 mA bzw. 50 mV, und ein typischer Serienwiderstand ist nur 0,75 Ω. Die Talspannung ist 300 mV bei einem beispielsweisen Spitzen-Talverhältnis von 8.

Mit diesen Kenndaten kann man einen Verstärker mit einem Rauschwert von 3,8 dB erzielen, der in einem Verstärker mit anderen Tunneldioden 4,5 dB betragen würde. Ein durch ein niedrigeres Ätzstromverhältnis erreichter niedriger Serienwiderstand trägt zur Verbesserung der Rauscheigenschaften bei; unter anderen Faktoren ist die Anwendung von Dotierungsverfahren zu erwähnen.

Ausserdem wurde die robuste Konstruktion der Mullard-Tunneldiode weiterhin durch Verwendung einer Zinn-Arsen-Legierungsperle von nur 15 μm Durchmesser verbessert. Dadurch wird das zur Herstellung der in diesen Elementen erforderlichen niedrigen Trennschichtkapazität notwendige elektrolytische Ätzen verringert.

Das Diodengehäuse ist bei 1,5 mm Durchmesser nur 1,4 mm lang. Durch Verwendung eines dünnen Keramik-Kreisringes kleinen Durchmessers und kurze Verbindungen mit rechteckigem Querschnitt zwischen der p-n-Verbindung und dem Gehäuse wird die Induktivität niedrig gehalten. Verwendung von Folie mit rechteckigem Querschnitt anstelle von Draht mit kreisrundem Querschnitt erhöht ausserdem die mechanische Festigkeit des Elementes.

EE 88 759 *für weitere Einzelheiten*

Anschlussstift

Vero Electronics Ltd, South Mill Road, Regents Park, Southampton

(Abbildung Seite 836)

Vero Electronics Ltd kann jetzt einen neuen Anschlussstift—Teilnummer 2143—liefern, der für Lochdurchmesser von 1,3 mm bestimmt ist und eine selbstschneidende Riffelverzahnung hat, um den Stift während des Lötvorganges in der Platte festzuhalten. Durch einen Bund wird verhindert, dass der Stift zu weit in die Platte hineingeschoben wird. Das Vero-Stifteinsetzwerkzeug bewältigt den ganzen Arbeitsgang einfach und schnell.

EE 88 760 *für weitere Einzelheiten*

Filter-Phasenschieber

Brookdeal Electronics Ltd, Myron Place, Lewisham, London, S.E.13

(Abbildung Seite 836)

Das durchweg mit Silizium-Transistoren bestückte Brookdeal-Modell FP316 besteht aus einem auf eine Frequenz abgestimmten Verstärker und einem Phasenschieber.

Der Phasenschieberteil gestattet stetige Phasenregelung eines sinusförmigen Signals bei weniger als 1 dB Änderung der Ausgangsspannung über den gesamten Phasenbereich. Ein Phasenregler überstreicht 160° und gibt zusammen mit einem 90°-Phasenschieber-Schalter einen Regelbereich von 250°. Das Gerät kann auf jede beliebige Frequenz zwischen 10 Hz und 100 kHz abgestimmt geliefert werden; ausserdem ist auf der Frontplatte ein Feinregler angeordnet, der genauen Abgleich auf die Frequenz der Bezugsspannung erlaubt.

Der dem Phasenschieber vorgeschaltete abgestimmte Verstärker filtert das Eingangssignal, so dass aus einem aus mehreren Frequenzen zusammengesetzten Eingangssignal ein Einzelfrequenz-Ausgangssignal mit regelbarer Phase entsteht.

Das Gerät wurde besonders für Verwendung mit phasenempfindlichen Nachweissystemen entwickelt, für die Phasenkorrektur eines Bezugssignals erforderlich ist. Anwendungsbeispiele sind u.a. Phasenkorrektur von Rechteckwellen-Bezugsspannungen oder Phasenkorrektur einer einem Mischer entnommenen Bezugsspannung. In jedem Fall wird das ungewünschte Signal durch den abgestimmten Verstärker unterdrückt, bevor die Phase des gewünschten Signals verschoben wird.

Die Höchstverstärkung ist 30 dB, und die Filterfrequenz kann vom Kunden zwischen 10 Hz und 100 kHz bestellt werden. Die Eingangsimpedanz ist 50 kΩ und die Ausgangsimpedanz niedriger als 1 kΩ; die Höchstausgangsspannung ist 10 V_{SS} und der Phasenbereich 250°.

EE 88 761 *für weitere Einzelheiten*

FS-Rauschmessgerät

Gresham Lion Electronics Ltd, Twickenham Road, Hanworth, Middlesex

(Abbildung Seite 836)

Ein Mehrfachnorm-Messgerät für statistisches Rauschen ist nunmehr von Gresham Lion Electronics Ltd zu beziehen. Das auf einem BBC-Entwurf beruhende Instrument wurde für alle 625-, 525- und 405-Zeilen-Fernsehnetzwerke entwickelt, um den Bedarf nach einem Gerät zum Messen statistischen Rauschens in Gegenwart von Fernsehsignalen zu erfüllen. Das Gerät gestattet schnelles, genaues Messen und beseitigt die der Oszillografendarstellung und anderen Verfahren eigenen menschlichen Fehlerquellen.

Das mit M.S.1 bezeichnete Gerät genügt den neusten CCITT/CCIR-Normen und ist vor allem für Messungen am Ausgang von Fernseh-Kamerazügen und Videobandgeräten während der Fertigung und in Betrieb geeignet. Besonders wo Synchronisierimpulse erforderlich sind, wird das Instrument die Messtechnik vereinfachen. Modell M.S.1 misst Rauschen in Gegenwart eines Standardpegelsignals und hat für das Verhältnis Signal : statistischem Rauschen einen Messumfang von 20...60 dB. Zusatzeinrichtungen sind nicht erforderlich.

Die Schaltung beruht auf einer Zeitselektionstechnik; das Instrument hat einen eingebauten Eichoszillator. Verhältnismässig grosse Schwankungen in Netzspannung, Temperatur oder Röhrenkennlinien rufen weniger als 0,25 dB Änderung der Ausgangsleistung hervor. Eine Spezialschaltung gestattet einen Vergleich des statistischen Rauschens in jedem Kopf eines Video-Magnetbandgerätes. Auf Wunsch kann das Gerät auch ein Signal für Oszillografen-Darstellung abgeben. In dieser Betriebsart lässt sich ein Unterschied von 1 dB in den Rauscheigenschaften nachweisen.

EE 88 762 *für weitere Einzelheiten*

Drucktasten-Messbrücke

The Wayne Kerr Laboratories Ltd, Sycamore Grove, New Malden, Surrey

(Abbildung Seite 837)

Die Universal-Messbrücke Autobalance B641 von Wayne Kerr ist eine neue Tonfrequenz-Messbrücke für das genaue Messen beliebiger Bauelemente oder Impedanzen. Auf allen acht Bereichen können selbst ungelernte Bediener 0,01 Prozent des Höchstwertes unterscheiden. Der Gesamtmessumfang für C, G, R und L ist 2 mΩ...50 GΩ und die Messunsicherheit zwischen 1 pF und 10 μF, 10 Ω und 100 MΩ sowie 1 mH und 10 kH 0,1 Prozent.

Drehregler sind für den Abgleich nicht vorgesehen, da zwei getrennte, elektronisch abgleichende Messschaltungen kontinuierlich die ohmsche und Blindkomponente des gemessenen Bauelementes anzeigen. Druckknopfschalter sind für Bereichwahl und Rückstellen (in Dekaden) vorgesehen. Die Anzeige ist beleuchtet, einzeilig, mit Komma und zugehörigem Messgerät.

Quelle, Detektor und niederohmige Spezialschaltungen sind eingebaut, und Schreiber, Digitalvoltmeter, Drucker usw. können direkt angeschlossen werden. Bauelemente kann man üblicherweise ohne Ausbau aus der Schaltung prüfen. Einrichtungen für vergleichsmässiges Messen sind vorhanden.

Das Gerät arbeitet mit Netzanschluss für 110 oder 230 V~, kann aber auch aus externen 9-V-Batterien gespeist werden, allerdings ohne beleuchtete Anzeige. Die Brücke B641 ist 19" breit, 267 mm hoch, 152 mm tief und wiegt etwa 7,5 kg.

EE 88 763 *für weitere Einzelheiten*

Umkehrbare Dezimal-Digitaldarstellung

Moore Reed (Industrial) Ltd, Woodman Works, Durnsford Road, London, S.W.19

(*Abbildung Seite 837*)

Moore Reed (Industrial) Ltd hat einen umkehrbaren Dezimal-Digitaldarsteller für die Umsetzung mechanischer Bewegung in elektrische Dezimalsignale herausgebracht, der keine komplexen Entschlüsselungseinrichtungen benötigt.

Die Digitaldarsteller wurden für Systeme entwickelt, in denen Parameter wie z.B. die Position eines Maschinentisches oder Werkzeuges, Durchströmung, Druck, Gewicht, Temperatur usw. in der Verfahrenstechnik dargestellt, geregelt oder registriert werden.

Der in einem Standardgehäuse Grösse 23 untergebrachte Digitaldarsteller arbeitet nach der M.R.-Abtast- und Bürstentechnik und wird mit freien Zuleitungen geliefert.

EE 88 764 *für weitere Einzelheiten*

Hochspannungsstromversorgung

Brandenburg Ltd, 139 Sanderstead Road, South Croydon, Surrey

(*Abbildung Seite 837*)

Bei Brandenburg Ltd ist die Fertigung des ersten Types einer Serie preisgünstiger Hochspannungsstromversorgungen für Photovervielfacher angelaufen. Typ PM.2500/R gibt 100 V—...2,5 kV— in fünf umschaltbaren Bereichen mit kontinuierlicher Grob- und Feinregelung ab. Der Höchststrom ist 5 mA, die Konstanz gegen Eingangsschwankungen 0,03 Prozent und die Restwelligkeit niedriger als 20 mV.

Für Polaritätsumkehr ist auf der Frontplatte ein Schalter mit der Mittelposition "Aus" vorhanden, und ein Ausgangsvoltmeter mit Doppelbereich ist eingebaut.

Das Gerät ist auf eine 19"-Frontplatte der britischen Postbehördennorm montiert, die 178 mm hoch ist, und für Tischeinsatz oder Einbau geeignet. Die Eingangsnennspannung ist 200...250 V, 50 Hz.

EE 88 765 *für weitere Einzelheiten*

Silber-Glimmerkondensatoren

Johnson, Matthey & Co. Ltd, 73-83 Hatton Garden, London, E.C.1

Johnson, Matthey & Co Ltd kündigt die Ergänzung ihres Silver-Star-Programmes hochkonstanter Kondensatoren mit versilberten Glimmerblättchen durch zwei neue Miniaturtypen an. Der in Kunstharz getauchte Typ C5F ist der kleinste Kondensator dieser Serie mit Abmessungen von 7 mm × 7 mm × 3,8 mm und einem Kapazitätsbereich bis zu 1000 pF. Typ C5E mit 680 pF Höchstkapazität und 7,6 mm × 7,6 mm × 2,5 mm Abmessungen ist der kleinste Kondensator mit Kunststoffumhüllung. Er hat die allen umhüllten Silver-Star-Kondensatoren eigene hohe Feuchtigkeitsbeständigkeit (Klasse 21, nach BS 2132).

Beide Typen werden in den bevorzugten Kapazitätswerten der Serie E24 BS 2488:1954 hergestellt und haben Toleranzen bis zu ±2 Prozent herunter (Minimum ±2 pF). Sie sind für Betriebsspannungen von 50, 200 und 350 V sowie Temperaturbereiche von −55°C...+100°C lieferbar. Die radialen Anschlüsse bestehen aus versilbertem Kupfer und haben 5,1 mm Abstand voneinander.

EE 88 766 *für weitere Einzelheiten*

Keramische Scheibenkondensatoren

A. H. Hunt (Capacitors) Ltd, Bendon Valley, Garratt Lane, Wandsworth, London, S.W.18

A. H. Hunt (Capacitors) Ltd hat soeben einen Entwicklungsauftrag des Luftfahrtministeriums für die Entwicklung von keramischen Scheibenkondensatoren mit hohem "K", die Temperaturen bis zu 200°C aushalten können, erfüllt.

Für die Kondensatoren wurde zur Begrenzung der Kapazitätsänderungen über den grossen Temperaturbereich −40°C ...+200°C ein Spezialwerkstoff entwickelt; sie werden durch Giessharzumhüllung geschützt. Der Kapazitätsbereich ist 470 bis zu 4700 pF mit einer bei 100 kHz gemessenen Toleranz von ±20 Prozent.

Die Kondensatoren sind 7 mm dick; ein 1000-pF-Kondensator hat einen Durchmesser von 17,5 mm, ein 4700-pF-Kondensator einen von 30 mm.

Der Isolationswiderstand ist bei 500 V— nach einer Minute grösser als 10^4 MΩ bei +20°C. Der höchste Verlustfaktor—tan δ—bei +20° ist 250×10^{-4}.

EE 88 767 *für weitere Einzelheiten*

Datenerfassung

Westland Aircraft Ltd., Yeovil, Somerset

(*Abbildung Seite 837*)

Das Westland-Modell 10.SL ist ein Zehnkanal-Datenerfassungsgerät für Einsatz mit Dehnungsmessstreifen; es ist für das Messen der von Kraftdosen, Druckgebern, Drehmomentmessern und Thermoelementen abgegebenen Signale geeignet.

Das Gerät kann ein Gemisch von vollständigen und halben Dehnstreifen-Brückenschaltungen akzeptieren. Die Streifen werden durch eine kurzschlusssichere Konstantstromversorgung erregt. Jede Brückenschaltung hat ihre eigne Eingangsbuchse und ein 10gängiges Abgleichpotentiometer.

Das Digitalvoltmeter hat einen Verstärkungsregler und eine interne Bezugsspannung, die Einstellen des digitalen Ausgangs direkt in Mikrodehnung oder anderen bequemen technischen Einheiten gestatten. Ein kompakter Streifendrucker gibt ein Protokoll jeder Messung und enthält ausserdem eine Digitalanzeige.

Modell 10.SL arbeitet vollautomatisch und bietet eine Auswahl von Arbeitsmethoden. Es kann über Nacht oder das Wochenende unbemannt für die Erfassung und Registrierung der Daten von Langzeitversuchen arbeiten.

Andere Erfassungsgeräte des neuen Westland-Programmes akzeptieren eine grössere Anzahl von Eingangskanälen und verschiedene Ausgangs-Registriergeräte einschliesslich Lochstreifenstanzen.

EE 88 768 *für weitere Einzelheiten*

Mikroschalter

A. F. Bulgin & Co. Ltd, Bye Pass Road, Barking, Essex

(*Abbildung Seite 838*)

Kleinstmikroschalter eines neuen, zuverlässigen Baumusters sind jetzt in voller Produktion. Der Grundschalter—Katalog-Nr. S.800—hat ein aus Phenolharz gepresstes Gehäuse nach BS.771; die Abmessungen sind einschliesslich Knopf und Lötösen 18,3 mm lang, 7,1 mm breit und 18,7 mm hoch. Befestigung erfolgt mittels zweier Querlöcher von 2,3 mm Durchmesser mit 12,7 mm Mittellinienabstand; die Schalter können entweder einzeln oder in Gruppen Verwendung finden. Der interne Mechanismus hat Neusilbermesser und Feinsilberkontakte. Der zweite Schalter—Nr. S.900—hat ein anspruchsvolleres Pflichtenblatt und ein in Araldite nach BS.771/2782 gepresstes, hochwiderstandsfähiges Gehäuse; die Anschlüsse sind zur Erleichterung des Lötens vergoldet. Beide Typen sind einpolige Umschalter, können aber auch als Ein-Ausschalter verdrahtet werden.

Die folgenden sind weitere Typen dieses Baumusters: S.800/L ist ein Einzelschalter mit Blattfederbetätigung; S.800/2/L besteht aus zwei Schaltern mit gemeinsamer Blattfederbetätigung; S.801 besteht aus zwei Schaltern mit gemeinsamer manueller Druckknopfbetätigung.

EE 88 769 *für weitere Einzelheiten*

Ferrit-Ringkerne

Standard Telephones & Cables Ltd, Magnetic Materials Division, Edinburgh Way, Harlow, Essex

(*Abbildung Seite 838*)

Diese neue STC-Baureihe wurde besonders darauf ausgelegt, dem Konstrukteur grösste Vielseitigkeit in der Kernwahl für beliebige Anwendungsaufgaben zu bieten. Ein logischer Grössenverlauf gibt eine Baureihe mit regelmässigen Abständen und zwei Kernhöhen für jeden Durchmesser, wodurch sich ein Verhältnis der wirksamen Fläche von 2:1 ergibt. Fast über die ganze Baureihe besteht daher die Wahl zwischen zwei Kernen mit der gleichen spezifischen Induktivität (A_L) oder derselben wirksamen Fläche (A_e).

Das Verhältnis des Innen- zum Aussendurchmesser jedes Kernes wurde so gewählt, dass sich die höchste mit leichtem Wickeln von Hand oder Maschine zu vereinbarende spezifische Induktivität ergibt. Grundlage dafür waren die über viele Jahre als Hersteller und Verwender ringförmiger Induktoren, Drosseln, Impuls- und Breitbandübertrager gesammelten Erfahrungen.

Die neuen Ringkerne gibt es in zwei Ferrit-Werkstoffen—SA503 und SA601—die im STC-Programm für IEC-Topfkerne bereits Erfolge verzeichnen konnten und jetzt Ringkerne mit höherer Permeabilität und niedrigeren Verlusten als früher möglich war geben.

Andere Eigenschaften sind, dass sie keine scharfen Kanten haben und in einer hellen Farbe einbrennlackiert sind. Das macht Bewickeln mit Band vor dem Wickeln überflüssig, und durch den weissen Einbrennlack wird die korrekte Wicklungsverteilung der Spule leichter erreicht. (STC hat einen Wickelservice für Kunden, die bewickelte Kerne beziehen wollen).

EE 88 770 *für weitere Einzelheiten*

Miniaturschleifringe

S. Smith & Sons (England) Ltd, Industrial Division, Kelvin House, Wembley Park Drive, Wembley, Middlesex

(*Abbildung Seite 838*)

Diese ursprünglich für Anwendung in Synkroskopen und Gyroskopen entwickelten Miniaturschleifring- und Bürstengruppen werden jetzt der allgemeinen Geräteindustrie geliefert; sie werden vom Fachbereich Industrie in drei Standardringdurchmessern und bis zu 22polig hergestellt.

Die aus einer Kupfer-Palladiumlegierung maschinenbearbeiteten Schleifringe werden in einem Nylon-Spritzvorgang zusammengebaut. Diese Technik hat sich der alternativen der Galvanoplastik gegenüber als überlegen bewiesen, da im letzteren Verfahren die Haftung und Homogenität des Metalls niemals gewährleistet werden konnte und die Wahl der Materialien auf diejenigen begrenzt war, die zufriedenstellend elektrolytisch niedergeschlagen werden können.

Die Bürstenbelastung ist durch die Bürstenform vorbestimmt, normalerweise 2 bis 3 g je Bürste. Auf Wunsch kann dieser Wert im Sonderauftrag geändert werden. Die Schleifringgruppe ist für Betrieb im Temperaturbereich −40°C... +140°C; die benutzten Werkstoffe sind von Natur aus gegen Korrosion und Pilzwuchs beständig.

Je nach gewähltem Modell sind die Gruppen zwischen 750 und 1500 UPM von regellosen Schwankungen frei. Sie sind für 1 A je Pol bemessen, und der mittlere Kontaktwiderstand ist 0,002 Ω.

Die durchschnittliche Länge der Gruppen ist nicht grösser als 25,4 mm. Andere Abmessungen schwanken mit der Anzahl der Pole.

EE 88 771 *für weitere Einzelheiten*

Elektronische Zähler

Advance Controls Ltd, Imperial Lane, Cheltenham, Gloucestershire

(*Abbildung Seite 838*)

Eine neue Serie preiswerter elektronischer Zähler wurde entwickelt, um industriellen Anforderungen, mit Geschwindigkeiten bis zu 5000 je Sekunde zusammenzuzählen, zu genügen; die Geschwindigkeitsgrenzen der elektromagnetischen Zähler werden bewältigt, ohne unnötige Einrichtungen vorzusehen.

Vier Modelle mit entweder vier oder fünf Dekaden, mit Ziffernzählröhren oder Dekatron-Anzeige sind in Produktion.

Start und Stopp der Zählung und Rückstellung der Anzeige auf Null werden entweder durch Schalter auf der Frontplatte oder Fernsteuerungskontakte ausgelöst. Ein Empfindlichkeitsregler ist vorhanden, und das Eingangssignal kann entweder an Klemmen auf der Frontplatte oder Buchsen auf der Rückseite gelegt werden.

Ein von der 50-Hz-Netzfrequenz abgeleitetes 100-Hz-Ausgangssignal kann in den Eingang gespeist werden, so dass man das Gerät als Zeitgeber einsetzen kann.

EE 88 772 *für weitere Einzelheiten*

Streifenschreiber

Beckman Instruments Ltd, Glenrothes, Fife, Scotland

(*Abbildung Seite 839*)

Ein vom Fachbereich Wissenschaftliche und Verfahrensinstrumentierung der Beckman Instruments angekündigter neuer 254-mm-Hochleistungsschreiber wurde speziell für Laboraufgaben entwickelt.

Der neue, hohe Leistung und zweckdienliche Konstruktion vereinende Schreiber hat breite Anwendungsmöglichkeiten in Laboratorien, z.B. in der Gaschromatografie, ultravioletten und infraroten Spektrophotometrie, pH-Messungen usw. Der Schreibstift spricht in unter 0,5 s über die ganze Skalenbreite von 254 mm an. Die Genauigkeit ist bei linearem Registrieren ±0,25 Prozent und bei logarithmischem Registrieren ±0,5 Prozent.

Bei der Konstruktion wurde einfache Bedienung stark in den Vordergrund gestellt. Alle Hauptbedienelemente sind in einem leicht zugänglichen Kontrollfeld zusammengruppiert. Streifenvorschub, Bereich, Arbeitsweise, Nulljustierung und Schreibstiftposition können sofort eingestellt werden. Man kann den 254 mm breiten Streifen leicht lesen, von Hand vorwärtsbewegen, sofort besichtigen und leicht markieren.

Ausser dem schnellen Ansprechen des Schreibstiftes und hoher Genauigkeit hat der Schreiber folgende Eigenschaften: eine kleine tote Zone—weniger als 0,1 Prozent der vollen Skala; starke Unterdrückung der Netzrestwelligkeit und Gleichtaktsignale; sofortige Fingerspitzenwahl von sieben Streifenvorschüben; drei geeichte Standard-Eingangsbereiche: 1, 10 und 100 mV; volle Nullpunktkontrolle; lineare oder logarithmische Aufzeichnung.

EE 88 773 *für weitere Einzelheiten*

Industrieller Zeitschalter

Solid State Controls Ltd, 30-40 Dalling Road, London, N.6

(*Abbildung Seite 839*)

Der "COHOMEE"-Zeitschalter besteht aus einem volltransistorisierten Standard-Zeitschaltungs-Einsteckmodul PT/900A, der mit einem zuverlässigen Einsteck-Ausgangsrelais auf einen Chassis-Befestigungswinkel montiert ist. Diese einfache Baugruppe bildet daher einen völlig integralen Zeitschalter, der in kürzester Zeit in jede industrielle Ausrüstung eingebaut werden kann. Es sind zwei Zeitbereiche von 0,2...60 s und 1...4 min lieferbar, innerhalb derer der Schaltzeitpunkt mittels eines Reglers am Befestigungswinkel vorgewählt werden kann.

In der Standardausführung ist der Zeitschalter mit einem für 5 A, 250 V~ bemessenen Umschalt-Ausgangskontakt ausgerüstet, der am Ende des vorgewählten Zeitintervalls betätigt wird. Lieferbar für folgende vier Betriebsspannungen: 24, 48, 115 und 250 V, 40...60 Hz.

Der Baustein wird durch Schwankungen der Betriebsspannung um ±10 Prozent und der Temperatur zwischen −30°C und +60°C praktisch nicht beeinflusst. Eine charakteristische Wiederholgenauigkeit von 2 Prozent kann erwartet werden, und Schutz gegen Hochspannungseinschwingvorgänge ist für begrenzte Zeit vorhanden. Die Lebensdauer des Zeitschalters ist bei Verwendung nach Pflichtenblatt bis zu 100 Millionen Schaltvorgänge.

EE 88 774 *für weitere Einzelheiten*

Zweistrahl-Oszillograf

Telequipment Ltd, 313 Chase Road, Southgate, London, N.14

(*Abbildung Seite 839*)

Der Oszillograf D53 ist mit einer neuen Zweistrahl-Elektronenröhre mit Nachbeschleunigung und rechteckigem Bildschirm, die nach dem Maschenprinzip arbeitet, bestückt. Fünf Verzögerungseinrichtungen mit automatischer Verbindung zu X- sowie Y-Verstärkern sind vorhanden. Die angewendeten Hybrid-Schaltungen sind zu 50 Prozent mit Festkörperelementen bestückt.

Ganz abgesehen von den Raumersparnissen hat die neue Maschen-Elektronenstrahlröhre den Vorteil, dass man ein höheres Nachbeschleunigungsverhältnis benutzen kann, woraus sich eine hellere Spur und grössere Empfindlichkeit ergibt. Diese Spezialröhre wird von der M-O Valve Co Ltd hergestellt.

Der neue X-Verstärker des D53 hat zwei Ablenkverzögerungsbereiche bis zu 5 ms und bis zu 50 ms; beide haben ungeeichte Verzögerungsregler. Die Ablenkschaltung hat 22 geeichte Ablenkgeschwindigkeiten von 5 s/cm bis zu 0,5 µs/cm im Verhältnis 1:2:5 und einen Verstärkungsregler für kontinuierliches Überstreichen zwischen Bereichen. Einmal-Ablenkung mit Ausblenden ist auch vorhanden.

Der X-Verstärker hat eine Bandbreite von 0...1 MHz (−3 dB) und bei Höchstverstärkung einen Ablenkfaktor von etwa 100 mV/cm. Die X-Dehnung ist bis etwa 10 Schirmbreiten stufenlos regelbar, wobei die Spur symmetrisch um die Mitte gedehnt wird.

Ausserdem gibt es zwei neue Y-Verstärker mit Signalverzögerung bis zu 0,25 μs, und zwar Typ CD (ultrahohe Verstärkung) und Typ HD (Breitband). Ausserdem können alle bereits mit den bestehenden Telequipment-Typen S43 und D43 lieferbaren auswechselbaren Y-Verstärker in diesem Oszillografen Verwendung finden.

Die Maschen-Bildröhre mit rechteckigem Schirm und Nachbeschleunigung im D53 gibt für jeden Strahl eine 6 cm hohe und 10 cm breite Darstellungsfläche. Die Röhre wird mit 8,5 kV betrieben, Z-Modulationseingang ist vorhanden.

Die Rasterscheibe ist beleuchtet und abnehmbar, und die Blende lässt sich zum Aufsetzen einer Kamera entfernen.

EE 88 775 *für weitere Einzelheiten*

Breitband-Speicheroszillograf

Tektronix UK Ltd. Beaverton House, Station Approach, Harpenden, Hertfordshire

(*Abbildung Seite 839*)

Der Oszillograf 549 ist mit einer schnellschreibenden bistabilen Sichtspeicherröhre bestückt. Die Darstellungsfläche von 6 cm × 10 cm wird für Anwendung mit geteiltem Bild in zwei unabhängig gesteuerte Speicherschichten von je 3 cm × 10 cm unterteilt; ausserdem besteht links eine nichtspeichernde "Such"-Zone und rechts eine Überschiess-Speicherfläche.

Unabhängige Kontrolle der Speicherschicht für jede Schirmhälfte gestattet Spurspeicherung auf einer Hälfte, während auf der anderen Hälfte entweder konventionell dargestellt, gespeichert oder gedehnt wird.

Eine weitere äussert nützliche Eigenschaft ist die Löschautomatik, die die Darstellung in vorgewählten Zeitintervallen löscht. Die Zeit zwischen Löschvorgängen kann zwischen 0,5 s und 5 s eingestellt werden. Für jede Schirmhälfte ist ausserdem eine manuelle Löschtaste vorhanden; Fernlöschung ist ebenfalls möglich.

Im Typ 549 kann jeder der erprobten Einschübe der Buchstaben-Baureihe oder der 1-Baureihe für Mehrzweck- oder hochspezialisierte Anwendung im Gebiet 0...30 MHz Verwendung finden. Das Gerät akzeptiert auch die neuen Spektralanalysatoreinheiten, wodurch es in einen speichernden Spektralanalysator abgewandelt wird.

Andere Eigenschaften des Types 549 sind u.a. geeichte Ablenkverzögerung von 1 μs...10 s, Einmalablenkung, ×5-Dehnung und vollständige Durchlasstriggerungseinrichtungen.

EE 88 776 *für weitere Einzelheiten*

Zusammenfassung der wichtigsten Beiträge

Ein langsamer Dreikanal-Zähler zum Messen von Alphateilchen von P. B. Roberts

Zusammenfassung des Beitrages auf Seite 784-787

Ein Dreikanal-Instrument zum Messen alphaausstrahlender radioaktiver Stoffe mit Aktivität bis zu einigen hundert Pico-Curie wird beschrieben. Es ist volltransistorisiert und kann mit Batterien oder Netzanschluss betrieben werden. Die Zählschaltungen können auch dort Anwendungsmöglichkeiten haben, wo Vorgänge, die mit niedrigen Raten auftreten, registriert werden sollen.

Spannungsregelung mit Kaltkathodenrelaisröhren von E. P. T. Atkinson und C. F. Hill

Zusammenfassung des Beitrages auf Seite 788-792

Es werden Einzelheiten über die Verwendung von Kaltkathodenrelaisröhren in der Stabilisierung kleiner Gleichrichter-Stromversorgungen mit veränderlichem Ausgang gegeben. Leistungen und Grenzen der Schaltung werden zusammen mit den den Entwurf von Kaltkathodenschaltungen bestimmenden Faktoren beschrieben. Die Z806W mit ihren konstanten, in engen Toleranzen gehaltenen Zündungskenndaten wird erwähnt. Abschliessend wird die grafische Bestimmung der Schaltungswerte für Ausgangsspannungen im Bereich 200 . . . 360 V beschrieben.

Ein vielseitiges Transistor-Reizgerät von W. J. Bannister und R. H. Kay

Zusammenfassung des Beitrages auf Seite 793-795

Ein vielseitiges Transistor-Reizgerät für die neurophysiologische Forschung, das aus drei getrennten Einheiten besteht, wird beschrieben:

(a) Der von der Zeitablenkwellenform der endgültigen Oszillografendarstellung getriebene Verzögerungsteil gibt einen Einzelimpuls oder ein Impulspaar an die Triggerteile (b) und (c) ab. Diese mit der Zeitablenkung synchronisierten Trigger können unabhängig "verzögert", d.h. in jeder Position entlang der X-Achse der Oszillografendarstellung konstant festgehalten werden. Sie bleiben—unabhängig von der Ablenkgeschwindigkeit—in derselben räumlichen Position entlang der Zeitablenkung.

(b) Der "Tetanus"-Teil gibt Auftastimpulsreihen an den Triggerteil (c) ab, wenn er durch einen Trigger vom Teil (a) ausgelöst wird. Impulsreihen kann man ausserdem unabhängig von Teil (a) durch Drucktaste manuell einleiten.

(c) Das Reizabgabegerät gibt einen Rechteckimpuls geregelter Amplitude, Dauer und Polarität ab, wenn es durch die Geräteteile (a) oder (b) oder manuell über die Drucktaste getriggert wird. Die Impulsausgänge werden üblicherweise durch einen gut konstruierten, zweifach geschirmten Übertrager isoliert, der die Rechteckimpulsform für Impulse bis zu 10 ms Dauer aufrecht erhält. Ausserdem gibt es für diese und längere Impulse einen Gleichstromausgang.

Frequenzumsetzer für Schnellzähler von P. Adby und E. A. Edis

Zusammenfassung des Beitrages auf Seite 796-802

Der beschriebene Frequenzumsetzer erweitert den Messumfang eines typischen 100-MHz-Zählers auf 900 MHz, wobei die Grundmessgenauigkeit des Zählers nicht beeinflusst wird. Der Entwurf beseitigt die kontinuierliche Abstimmung, die man üblicherweise in solchen Geräten findet, so dass eine Reihe von Frequenzen sehr schnell durch Einstellung eines der sechs umschaltbaren Frequenzbereiche gemessen werden können. Durch Überwachung der Ein- und Ausgangspegel lassen sich unbekannte Signale ohne Mehrdeutigkeit messen. Mit Ausnahme einer Eingangstrennröhre ist die Schaltung durchweg mit Halbleitern bestückt; verschiedene interessante Punkte im Entwurf werden eingehend beschrieben.

Feldeffekt-Transistoren für biologische Verstärker von R. E. Webb

Zusammenfassung des Beitrages auf Seite 803-805

Eine kurze Darstellung der für die Verstärkung bioelektrischer Spannungen bestehenden Anforderungen geht der Information über die Verwendung von Feldeffekt-Transistoren für diesen Zweck voraus. Es werden einige Vergleiche zwischen Vakuum-Elektrometerröhren und Feldeffekt-Transistoren gezogen und Einzelheiten eines hochohmigen Verstärkers gegeben. Schaltungsrauschen und seine Herabsetzung werden auch kurz erwähnt.

Ein tragbarer Elektrokardiosender von A. R. Atkins und E. B. Lock

Zusammenfassung des Beitrages auf Seite 806-808

Ein beschriebener tragbarer Elektrokardiosender wird bei hoher Feuchtigkeit und Tempertatur, z.B. in Klimakammern oder unter Tag in Bergwerken, für physiologische Untersuchungen eingesetzt, um die Elektrokardioimpulse eines arbeitenden Subjektes aufzunehmen und zu einem örtlich entfernten Empfänger zu übertragen.

Bei der Entwicklung des für den Nachweis der Elektrokardioimpulse entwickelten Messwertgebers wurde berücksichtigt, dass die Subjekte möglicherweise nass vom Schwitzen sind.

Der Empfänger wurde so ausgelegt, dass er auf einen von vier Sendern eingestellt werden kann, so dass der Beobachter die Möglichkeit hat, den Herzschlag entweder im Kopfhörer zu zählen oder durch Anschluss eines Elektrokardiografen zu registrieren.

Varistor-Netzwerke als Analog-Rechenelemente von W. G. P. Lamb und M. L. Stones

Zusammenfassung des Beitrages auf Seite 809-811

In diesem Beitrag beschriebene einfache Netzwerke aus nichtlinearen und linearen Widerständen werden so entworfen, dass sie ein Element ergeben, das genau einer Strom-Spannungsbeziehung $i = Kv^n$ folgt, wenn n zwischen eins und drei liegt. Eine Anwendung in bezug auf die Diffusion von Wasserstoffionen in Eisen wird umrissen.

Astabile Transistor-Multivibratoren für logische Mehrzweckelemente von G. H. Stearman

Zusammenfassung des Beitrages auf Seite 812-817

Die wünschenswerten Eigenschaften eines als Taktgeber mit einem Mehrzwecksatz logischer Elemente eingesetzten astabilen Multivibrators werden beschrieben. Die Nachteile der konventionellen Schaltung, mehrerer Abwandlungen derselben und einiger nichtkonventioneller Schaltungen werden aufgezeigt und anschliessend einige neue Abwandlungen der Verknüpfung vorgestellt. Es wird gezeigt, dass die letzteren den gestellten Anforderungen besser genügen, als die bereits früher verfügbaren Schaltungen.

NF-Leistungsverstärker mit Übergangsrückkopplung von A. R. Bailey

Zusammenfassung des Beitrages auf Seite 818-821

Konventionell entworfene NF-Leistungsverstärker für hochwertige Wiedergabe sind für viele Laboraufgaben unzureichend. Besonders Verzerrungen sind häufig in einem Teil des Frequenzspektrums übermässig hoch, und die Stabilität wird oft bei kapazitiver Belastung nicht aufrecht erhalten. Dieser Beitrag beschreibt die Entwurfsphilosophie und Leistung eines für Einsatz im Labor entwickelten Verstärkers, der viel bessere technische Daten hat als die üblicherweise akzeptierten.

Die volle Nennleistung kann über den ganzen Tonfrequenzbereich mit einem Verzerrungsniveau von unter 0,1 Prozent erreicht werden. Ausserdem ist der Verstärker für alle Belastungswerte bedingungslos stabil und hat gute Erholungseigenschaften nach Überlastung.

Eine Drossel-Zenerdioden-Filterschaltung von J. M. Diamond

Zusammenfassung des Beitrages auf Seite 822-825

Die Vorteile der Drossel-Zenerschaltung, zu denen Kurzschlussschutz, Stabilisierung und Brummsieben gehören, werden besprochen. Bei Nachschaltung eines konventionellen gegengekoppelten Konstanthalters gibt die Drossel-Zenerschaltung einen äusserst einfachen und narrensicheren Kurzschlussschutz und ermöglicht äusserst niedrige Werte für Stromversorgungsregelung und Brummpegel. Die Schaltung wird auf Verlustleistung in den verschiedenen Elementen, einschliesslich des Durchlasselementes des gegengekoppelten Konstanthalters, und unter den schlimmsten Netz- und Belastungsbedingungen analysiert.

Ein wiederholender Elektronenschalter für die Isotopenforschung von E. van der S. Lotz

Zusammenfassung des Beitrages auf Seite 825-827

In Versuchen mit kurzlebigen Isotopen und für gewisse statistische Untersuchungen muss man oft einen kerntechnischen Impulszähler viele Male hintereinander für kurze Perioden in Betrieb bringen. Beschrieben wird ein Elektronenschalter, der solche Zähler wiederholt für vorgewählte Perioden von 1, 2, 4, 8 oder 16 Sekunden mit Ruheperioden betreibt. Die gewünschte Schaltfolge wird durch eine 12-Kathoden-Glimmentladungsröhre erzeugt, deren Steuerimpulse von gleichgerichteten, ungefilterten 50-Hz-Sinuswellen abgeleitet werden. Sowohl Voll- wie Einweggleichrichter finden Verwendung, und die Umschaltung von einer Arbeitsweise zur anderen erfolgt mittels einer kleinen Gastetrode.

Eine Hochleistungs-HF-Reiz-Trenneinheit für Anwendung in der Elektrophysiologie von W. T. Catton und R. G. Farrier

Zusammenfassung des Beitrages auf Seite 828

Rechteckimpulse von einem herkömmlichen Reizgerät werden zur Schirmgittermodulation eines bei 5 MHz arbeitenden, elektronengekoppelten Tetrodenoszillators benutz. Die entstehenden HF-Impulse werden durch ein verlustarmes Koaxialkabel in einen abgeschirmten Vollwegdemodulator gespeist, dessen Gleichstromausgang gegen Erde isoliert ist. Leistungsumsetzung erfolgt mit gutem Wirkungsgrad, und es wird eine viel grössere Leistung abgegeben, als mit früheren Trenneinheiten möglich war.