Commentary

"FOR many people, automation is a terrifying word." These were the opening words of Sir Leon Bagrit, Chairman of Elliott-Automation in this year's Reith Lectures entitled, 'The Age of Automation', which are being broadcast by the BBC.

He went on to say that it conjures up visions of tyrannical machines reducing man to the status of a mere pusher of buttons or watcher of dials and abolishing the need for human thought and judgement.

For many people indeed this is the conception of automation and it is hoped that when Sir Leon has completed his lectures a far better understanding of automation will result.

It is true as he says that most people's fear of automation is that it will "turn them into semi-human or sub-human types: into something very close to robots" but this is because they confuse automation with mechanization.

Mechanization, particularly in many mass production systems does demand no more than "unthinking repetitive motion on the part of the human operator." It is essentially a timing machine which moves the product from place to place in the various stages of manufacture in a given time and during this time a man has to be available to perform a given task. He is "virtually a slave of the machine which fixes his time and movements and he has to produce a series of semi-intelligent and mechanical motions to keep the machine fed and moving."

As is well known, the word "automation" was invented some ten years ago at the Ford Motor Works at Detroit where large scale technological changes in the assembly plant were taking place and it was felt that the word "automatic" was no longer appropriate and "automation" was substituted.

A more preferable word, according to Sir Leon, is "cybernation"—another word of American origin—because it deals with the theory of communications and control.

It is a concept on which a machine system, based on man himself as the model, is "caused to operate with maximum efficiency by means of adequate measurement, observation and control of its behaviour", involving a "detailed and continuous knowledge of the functioning of the system so that the best corrective actions can be applied immediately they become necessary".

Automation of this kind undoubtedly offers the most exciting prospect for the future but it applies at the moment only to certain sections of industry, notably to the steel and oil industries who have modernized their processes by automation methods.

But as Sir Leon points out we shall need to apply our scientific and technological resources to literally every aspect of our society, to our commerce, our industry, our medicine, our agriculture, and our transportation.

Looking at the last of these aspects—namely transportation—there is already a limited degree of automation. We have, for example, the automatic pilot system in aircraft which relieves the human pilot of a lot of his routine tasks and enables him to devote his attention to more skilled activities, but the completely automatic, pilotless aircraft is not yet here.

Nor is the automatic driverless train or bus. In this country, British Railways have installed a number of computers but these are primarily for such administrative applications as general accountancy, revenue, payroll and so on. No computer system has yet been introduced which permits the operation of a driverless train, although it should perhaps be recalled that the British Post Office has for many years been operating its own driverless underground train system in London for parcel transport between the main railway termini.

There is also in this country called, appropriately enough, the robot-tug system which is operating quite satisfactorily.

It can be programmed to follow a prescribed route in a factory or warehouse, stop at any given point along the route, and halt if there is any risk of collision—all this without the necessity of a driver.

The driverless bus seems a very long way ahead although there are certain automatic devices based on radar and other techniques which could enable the driver to function in the same manner as a pilot in an aeroplane. The device would warn him if he has departed from his lane or if he is in danger of collision with a car ahead.

The conductorless bus, if this can be accepted as a form of automation, is much nearer and in this issue, brief details are given of an electronic ticket recognition machine which is now being installed in Manchester and Frankfurt and which, although it places a small additional load on the driver, eliminates the conductor.

These are but a few of the first steps towards complete automation, but complete automation in all the aspects of our society, to which Sir Leon refers in his Reith Lectures, is so far removed that it is difficult to accept that it can "free vast numbers of people from their traditional bondage to nature" and that we have the promise "both of greater leisure and even greater material and intellectual riches".

Automation, says Sir Leon, is a tool of such immense possibilities that no one can yet see the full extent of what it might achieve for mankind, but he promises in the remaining lectures to contribute to our thinking about these possibilities.

A Four Decade Transistor Linear Densitometer

By E. Gordy*, M.D., D.Sc., Peter Hasenpusch* and George F. Sieber*

A solid state linear densitometer circuit having an operating range of optical density from zero to four has been designed using negative dynode voltage feedback combined with biased diode linearity correction. The linearity of output signal with respect to optical density is ±0·5 per cent. A germanium and a silicon transistor version are described.

(Voir page 866 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 873)

LOGARITHMIC response devices (those whose output is a logarithmic function of the input) have long been favoured by designers, for several reasons. Of these, perhaps the most cogent is the large dynamic range which can be covered without need of range changing switches (with their associated attenuators). As a consequence, two advantages are gained immediately. First, within the normal range of use, the device cannot be overdriven (and possibly damaged) due to improper range selection by the user. Second, a single-scale unambiguous readout can be provided, which eliminates errors due to reading the wrong scale. These considerations will naturally apply to a photometric logarithmic response device.

Where the logarithmic response photometer is part of a colorimeter or spectrophotometer, there are two additional factors to be considered. One is the way that light is exponentially attenuated as it passes through a uniformly coloured absorbing medium. The second factor is the way that light is exponentially attenuated, as equal increments of an absorbing substance (e.g. a dye) are added to and dissolved in a transparent medium situated in a suitable container of fixed optical depth.

Strictly speaking, both of these effects are true only for monochromatic illumination. At any single wavelength, these relationships may be expressed mathematically as

$$I = I_0 \, e^{-kcd} \quad \dots \dots \dots \dots \dots \dots (1)$$

where I_0 = initial intensity of light

  I = final intensity of light

  c = concentration of absorbing substance

  d = optical depth of absorbing medium (e.g. solvent plus solute)

  k = a constant which is unique for a particular absorbing substance present in unit concentration and in unit optical depth.

A more useful expression may be obtained from equation (1). Transposing terms and taking the logarithm of both sides gives:

$$\ln I/I_0 = -kcd \quad \dots \dots \dots \dots \dots (2)$$

The minus sign can be removed leading to

$$\ln I_0/I = kcd \quad \dots \dots \dots \dots \dots \dots (3)$$

Now $\ln I_0/I$ is called absorbance (symbol A) and one may now write

$$A = kcd \quad \dots \dots \dots \dots \dots \dots (4)$$

This expression is the basic working equation of colorimetry and spectrophotometry. In practice, the optical path of the absorbing medium is kept constant by using cuvettes (containers with optically transparent end windows) of fixed optical length for holding the optically absorbing samples. The constant k is known for the substance under study, having previously been determined experimentally. Since k and d are fixed, concentration c is directly proportional to absorbance A. An instrument linear in absorbance would provide an output linear in concentration, often a useful feature.

In the field of photographic film sensitometry, it has been the practice since the days of Herter and Driffield, to indicate film characteristics graphically by plotting film blackening against film exposure. In order to cover the large range of exposures and blackening on a single graph, logarithmic compression is employed. Exposure is plotted on a logarithmic time scale and film blackening is expressed as $\ln I_0/I$, where the symbols have the same meaning as in equation (3). In sensitometry, $\ln I_0/I$ is called optical density. In the older literature of spectrophotometry, the term optical density is used in place of the more modern absorbance. For the sake of euphony, and to convey some idea of its broad applicability, the name densitometer is used for the logarithmic photometer about to be described.

Basic Principle of Operation

Consider the photomultiplier as a photo-detector. It offers not only extreme photosensitivity but also the possibility of modifying this photosensitivity simply by vary-

Fig. 1. Functional diagram of dynode negative feedback circuit used as a densitometer
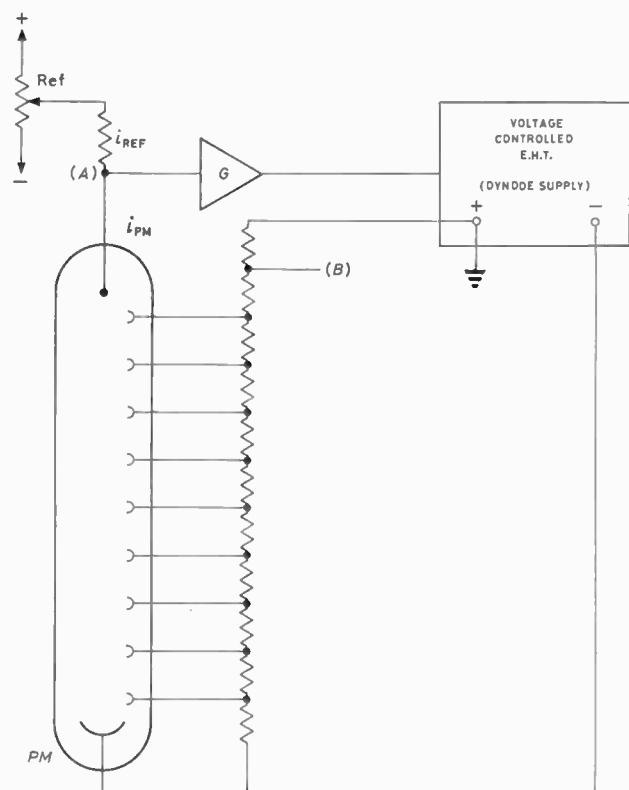
* Roswell Park Memorial Institute, Buffalo, New York, U.S.A.
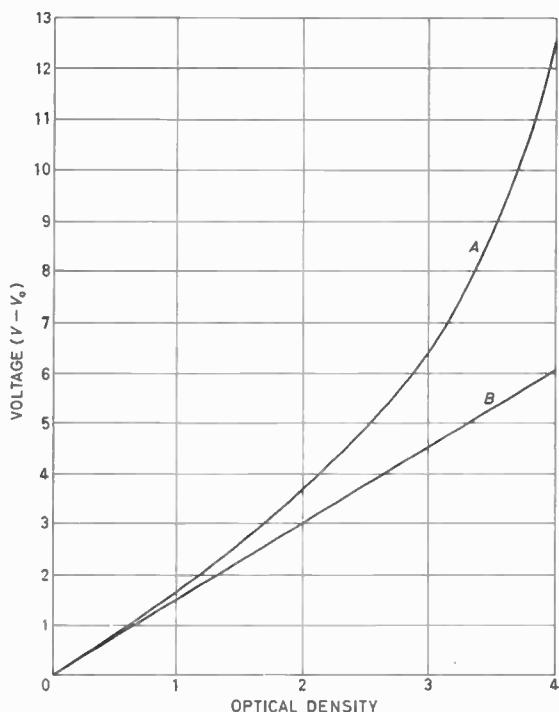
Fig. 2. Graph of output voltage against optical density
Curve A is before linearization
Curve B has been linearized using a biased diode network

ing the voltages supplied to the dynodes. In Fig. 1 is shown an arrangement wherein the photosensitivity of a photomultiplier *PM* is made to vary inversely with the level of photocathode illumination. *G* is an operational amplifier which has a high (negative) transresistance; i.e., a small increase in input current produces a large decrease in output voltage. The operation of the circuit is as follows. The photomultiplier anode current i_{PM} and the reference current i_{REF} are added algebraically at point *A*. Their sum is the input signal to amplifier *G*, whose output voltage is the control signal of a direct-coupled voltage-controlled e.h.t. (dynode) supply.

For any particular intensity of illumination, there will be a particular photomultiplier anode current i_{PM}, and a corresponding e.h.t. level supplying the photomultiplier dynode string and photocathode. With an increase in illumination intensity, i_{PM} tends to increase, which serves to decrease the e.h.t. dynode supply. The decreased dynode (and photocathode) supply level then tends to cause a reduction in i_{PM} due to a decrease in dynode electron-multiplying efficiency. The net result is that the new i_{PM} will be very slightly increased while the e.h.t. level will be reduced.

A decrease in illumination tends to decrease i_{PM}, which causes an increase in the e.h.t. level, thereby tending to increase i_{PM}. The net result of decreased illumination is a slight decrease in i_{PM} and an increase in e.h.t. It will be clear that, with sufficiently high (negative) transresistance in amplifier *G*, i_{PM} tends to remain substantially constant, while the e.h.t. output level is inversely related to the intensity of illumination striking the photocathode of *PM*.

Point *B* of Fig. 1 is a convenient place to pick off an attenuated sample of the e.h.t. level. Curve *A* of Fig. 2 shows the voltage $V - V_0$ at point *B* plotted as a function of $\ln I_0/I$.

Mathematically, curve *A* may be expressed as:

$$V - V_0 = k (\ln I_0/I)^x \ldots\ldots\ldots\ldots (5)$$

where $x > 1$. By the use of a biased diode network, x can be made equal to 1, giving a straight line relationship between device output $(V - V_0)$ and optical density $(\ln I_0/I)$, as shown in curve *B*, Fig. 2.

The exponential relationship between the incident illumination of a photomultiplier and its fed back dynode voltage level, along with linearization by a diode network, was worked out by Sweet[1] using valves in a direct coupled circuit. These ideas have been taken as the starting point for the design of two semiconductor circuit variants, one using germanium, the other, silicon transistors.

In Fig. 3 is shown the functional block diagram of the present linear densitometer applicable to both semiconductor versions. The source of e.h.t. for the photomultiplier dynodes is a d.c. to d.c. convertor, the use of which confers certain advantages. First, the e.h.t. output level is controlled at a low voltage point in the circuit, using amplitude modulation of the collector supply of a push-pull oscillator. Second, with the recent availability of small, low-capacitace high voltage silicon diodes, it is now feasible to operate a solid state convertor efficiently at a frequency of 10kc/s or higher, which permits the use of small e.h.t. filter capacitors. Finally, because of the high frequency and low capacitance of the e.h.t. supply, the system responds rapidly to imposed changes, so that, e.g. an instantaneous increase in optical density from zero to four (full scale optical change) is faithfully indicated at the output terminals within 7msec.

Before turning to the actual densitometer circuits, one further item deserves attention. The success of the design depends on very stable power supplies and in the elimination of drift, mostly thermal in origin. This is not always easy to achieve, and in the case of the germanium transistor model, thermal correction is introduced at several

Fig. 3. Functional diagram of semiconductor linear densitometer
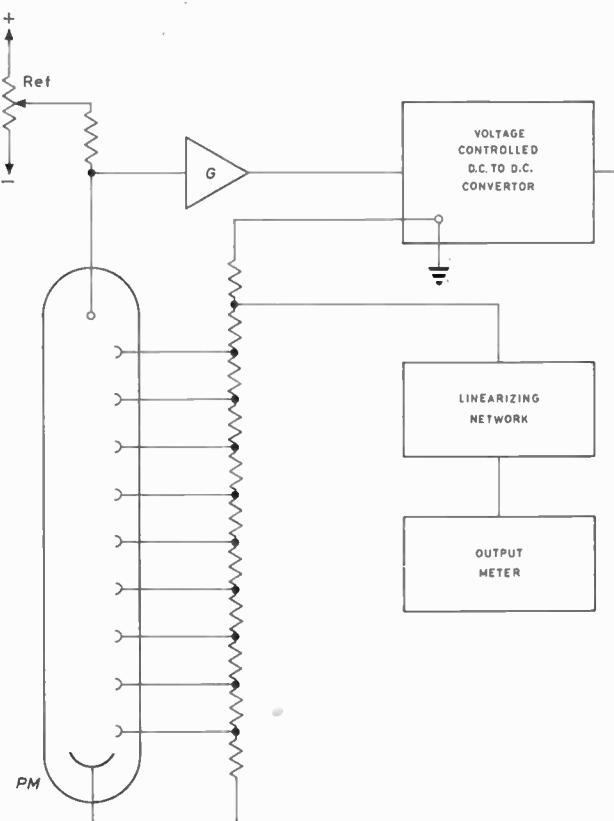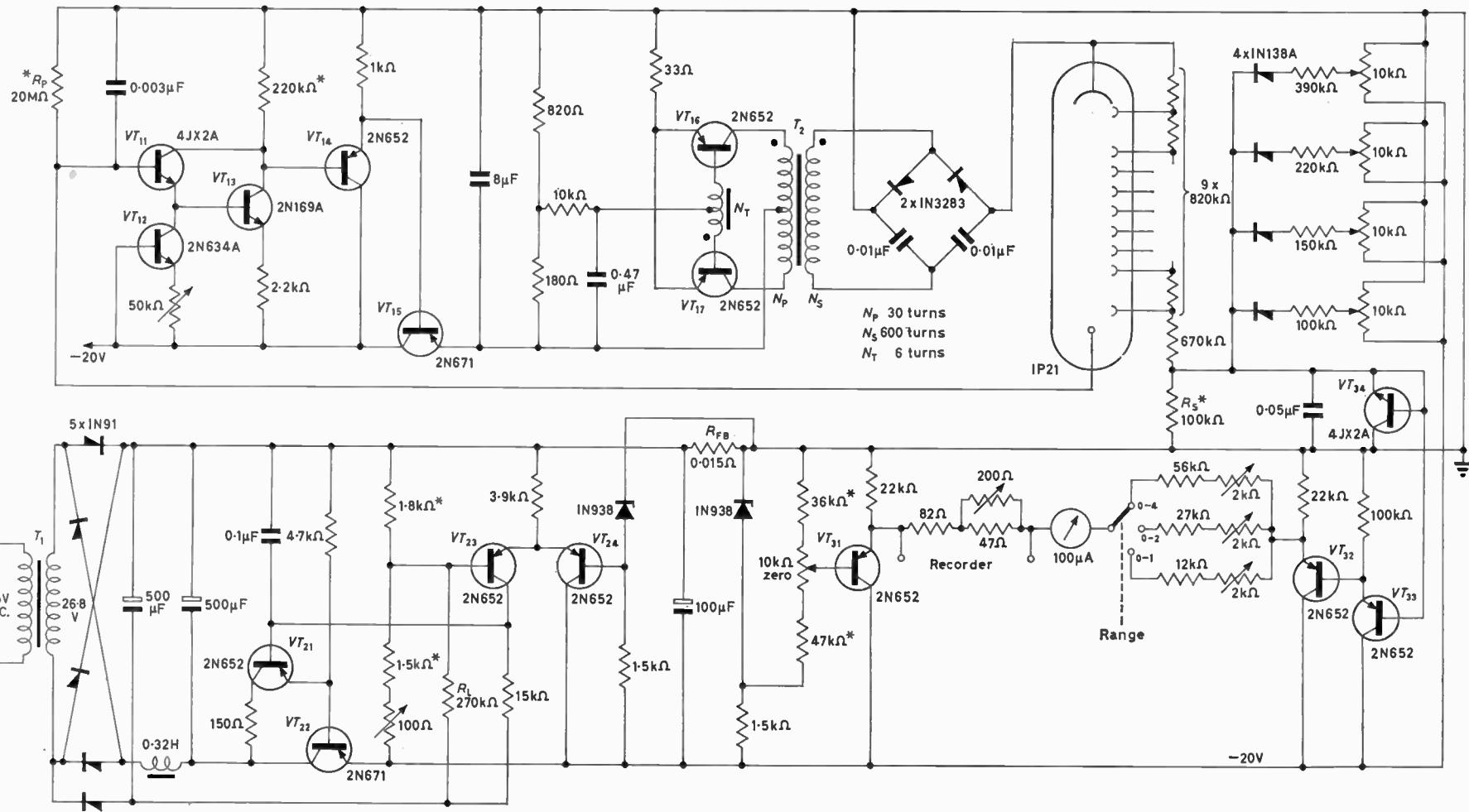
Fig. 4. Linear densitometer using germanium transistors

Fig. 5. Linear densitometer using silicon transistors
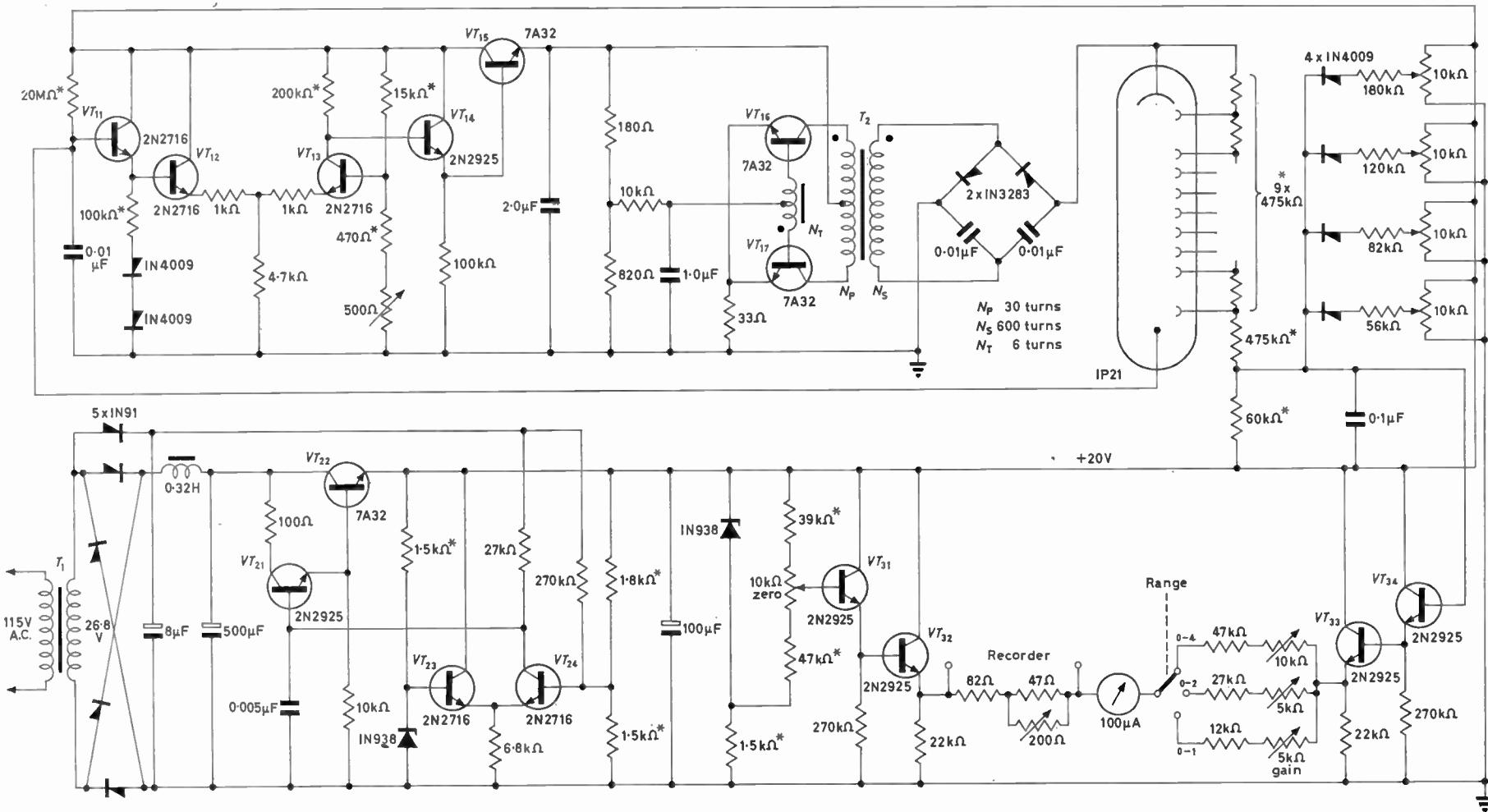
points to compensate for the effects of changing ambient temperatures on the I_{cbo} and V_{be} of sensitive stages.

Germanium Transistor Circuit

Fig. 4 shows the germanium transistor densitometer circuit. Transistors VT_{11} and VT_{13} constitute a compound (Darlington) input stage. Rate grown npn transistors are used here because of their low I_{cbo} values. VT_{11} is a selected version having an I_{cbo} of 10^{-7}A at a V_{be} of 15V at 20°C. This permits the use of a 20MΩ resistor for R_D at the base of VT_{11}. NPN alloy transistor VT_{12} provides thermal compensation for VT_{11} and VT_{13}. A cascade emitter-follower is formed by pnp alloy transistors VT_{14} and VT_{15}. These four transistors constitute a high (negative) transresistance amplifier whose output voltage is the collector supply of a collector-modulated push-pull oscillator made up of VT_{16}, VT_{17} and transformer T_2. This transformer consists of a ferrite pot core wound with a primary winding of 30 turns of 28 gauge wire, a secondary winding of 600 turns of 34 gauge wire and a tertiary winding of 6 turns of 28 gauge wire.

The rectified and filtered output e.h.t. from this oscillator goes directly to the photomultiplier cathode and dynode string. An attenuated sample of this e.h.t. is taken from a 100kΩ resistor R_s in the dynode string. This voltage is subject to thermal compensation by VT_{34} and to linearization by the four biased diodes. The linearized voltage is fed to a meter and associated output terminals via cascaded emitter-follower VT_{33} and VT_{34}. Transistor VT_{31} provides an emitter-follower meter reference level for zero setting (equivalent to subtracting V_o from V). Photomultiplier anode current is fed to VT_{11}, completing the control loop.

No reference current adjustment is provided in this circuit version. Almost all photomultiplier tubes tried (among 1P21 and 1P28 types) had sufficiently low dark current to allow their use in this circuit. Since the amplifier-dynode supply combination only operates (as described above) over a finite range of current and voltage, in densitometry practice, any differences in absolute photosensitivity between various individual photomultiplier tubes are compensated by appropriate adjustment in the intensity of the light source; a fine adjustment is provided by the potentiometer in the base circuit of VT_{31}.

The 20V power supply is conventional. Two features deserve note. Resistor R_L, feeding forward a full wave rectified sample of the secondary voltage of transformer T_1 to the base of VT_{23}, reduces the output ripple by a factor of five and makes the regulated output independent of line voltage over the range 105 to 125V a.c. Resistor R_{FB}, by providing positive current feedback, keeps the output voltage constant to within 1mV, over the full range of output current.

When this circuit is used as a four-decade logarithmic photometer, no range switch is needed or provided. However, in one application, as a monitor for the effluent of a chromatographic column, it was thought desirable to include a range changing switch to provide full scale indication when monitoring dilute solutions; therefore this switch is shown in Fig. 4. Since the recorded optical density is directly proportional to the concentration of effluent-borne solute, one can make direct measurements of the latter from the recorder chart. This particular instrument has been in daily use for over two years.

Silicon Transistor Circuit

With the recent availability of inexpensive planar epitaxial passivated silicon transistors (measured I_{cbo} 10^{-9}A at a V_{be} of 20V and β of 150 to 250), an all silicon transistor densitometer was designed; the circuit is shown in Fig. 5.

Here again a fixed reference current is employed. Input transistor VT_{11} feeds the signal to a high gain difference amplifier VT_{12} and VT_{13}. A variable resistor in the base circuit of VT_{13} provides an adjustment for variations in photomultiplier dark current.

The voltage at the collector of VT_{13} is fed to cascaded emitter-followers VT_{14} and VT_{15}. The emitter voltage of VT_{15} is the collector supply of the push-pull collector-modulated oscillator formed by VT_{16}, VT_{17} and transformer T_2. This again is a ferrite pot core type, wound as indicated in Fig. 5. It will be seen that the remaining circuit geometry is similar to that used in Fig. 4, the germanium transistor circuit. In general, the performance of the silicon transistor circuit is similar to that of germanium transistor circuit, but is obtainable without recourse to the various thermal compensation devices previously required. This is to be expected, in view of the 100 times reduction in I_{cbo} of the sensitive input stage.

Although the circuit geometry of the 20V regulated power supply in Fig. 5 is similar to the previous model shown in Fig. 4, the need for positive current feedback is eliminated because of the extra loop gain provided by the high β values of the silicon transistor types used. The use of silicon transistors in the present instance has the virtue of allowing circuit simplification with no sacrifice in circuit performance.

The germanium and silicon transistor circuit versions are similar in performance. An optical density range from zero to four is covered (as shown in curve B in Fig. 2) with a linearity of ± 0.5 per cent. This is equivalent to covering a $10^4 : 1$ range of illumination intensity on a single four-decade logarithmic scale. Under laboratory ambient conditions of $26°C \pm 1°$ the equivalent input drift was ·005 in optical density units for the germanium transistor model and less than this for the silicon transistor version, both recorded over a 16 hour period. Neither circuit is affected by line voltage variations of ± 10 per cent.

REFERENCE
1. SWEET, M. H. U.S. Patent 2 492 901.

Ultrasonic Welding of Plastic Materials

A number of synthetic materials can be joined together or welded by the familiar dielectric method in which the sections of the material to be welded are placed between two electrodes to which a high-frequency voltage is applied.

Certain plastic materials such as polyvinyl chloride can be welded by this dielectric loss method but polyvinyl chloride leaves a residue of chlorine after welding and is not altogether suitable for the packaging of food or similar products.

Polyvinyl chloride can be replaced by other plastic materials such as polyethylene, polyamide or polystyrene but these materials cannot be welded by the existing methods due to their low dielectric loss.

An alternative method of welding of these low dielectric loss materials has now been devised whereby the materials can be fused at ultrasonic frequencies.

In this method the conventional high-frequency electrodes are replaced by welding tools of special design and shape—known as sonotrodes—which exert a static pressure on the material to be welded and which at the same time are fed from an ultrasonic frequency generator.

Thus the material between the sonotrodes is subjected to an alternating pressure at ultrasonic frequency and this alternating pressure will give rise to heat due to the frictional losses in the material, and, if sufficient, will produce welding action.

The heat generated is independent of the dielectric properties of the material so that virtually all thermoplastic synthetic materials can be welded by this process.

A complete welding unit operating on this principle has now been developed by Korting Radio Werke. G.m.b.H. of Grassau, Chiemgau in West Germany with an output up to 1kW at a frequency of about 20kc/s.

An Experimental Electronic Ignition System

W. Alexander*, Ph.D., M.Sc., M.I.E.E. and P Sample†, B.Sc.

An experimental investigation into the advantages and disadvantages of an electronic ignition system for automobile engines is described. In the method employed a capacitor is discharged into the primary of a transformer with s.c.r. used for switching. It is shown that an increased performance can be obtained from the engine.

IN recent years there has been an increasing interest in the use of so called electronic methods of ignition for petrol engines. The work described here consists of a short investigation into the advantages and disadvantages of a typical electronic ignition system, but it is in no way meant to represent an ideal approach.

Before considering the justification for electronic systems, it is necessary to consider the disadvantages of normal ignition systems. There are two methods in common use today, namely, the battery-coil and magneto. Only the former will be considered, since this is used almost universally on automobile engines and it is in this sphere that the main interest in electronic ignition lies.

The circuit of a battery-coil ignition system is shown in Fig 1. When the ignition switch is closed, the primary winding of the induction coil is connected across the battery during the 'closed' period of the contact breaker. During the 'open' period the primary is disconnected and no current flows. The rate of growth of current in the primary during the 'closed' period is very important, since this determines the magnitude of the primary current broken when the contacts are separated. At the moment when the contacts begin to separate, the voltage rises abruptly in the secondary winding due to the collapse of the magnetic field previously built up by the growth of current in the primary winding.

The disadvantages of the above system can be listed as follows:

(1) The large value of current being interrupted, by the contact-breaker points, causes excessive wear.

(2) The moving arm of the contact-breaker tends to 'bounce' at high speeds, thus shortening the time of the 'closed' period and so causing contact wear.

(3) A large reduction occurs in the output spark voltage with increasing spark repetition frequency i.e. engine speed. This is a common restriction in the design of high speed multi-cylinder engines.

(4) This system is highly inefficient at low speeds due to the high current required.

Therefore an improved system should have reduced current in the contact-breaker, or better still this contact should be eliminated completely. Also, the output voltage should be approximately constant over the entire range of engine speeds and the current drain, particularly at low speeds, should be reduced. In considering how well these conditions can be achieved by the use of an electronic system, two approaches are possible. Firstly the high voltage spark can be generated by some form of triggered blocking-oscillator using a high voltage transistor and
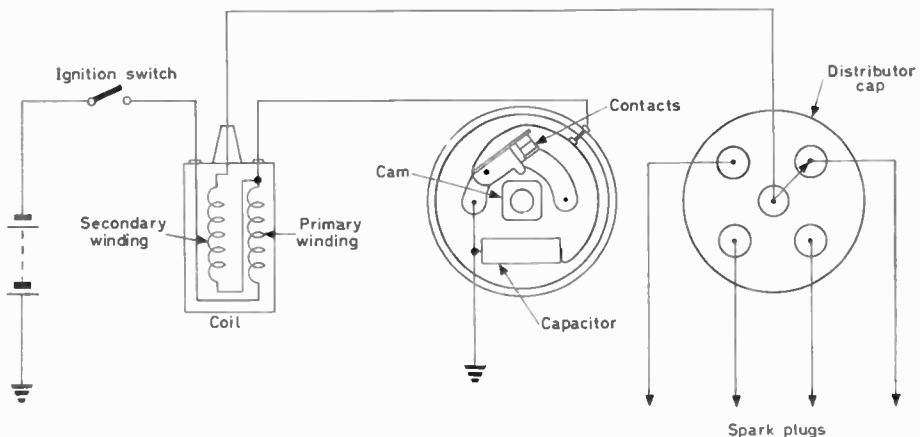

Fig. 1. Battery-coil ignition circuit

transforming up the pulse thus produced. However this requires the use of a transistor with a voltage rating of the order of 200 to 300V, which was not readily available at the time when this work was carried out. Secondly, spark generation is possible by discharging a capacitor into the primary of a transformer. It is this method that was chosen, the necessary switching being performed by a silicon controlled rectifier (thyristor).

Basic Spark Generating Circuit

The circuit used for generating the high voltage pulse, required for ignition, is shown in Fig. 2. When the silicon controlled rectifier is not conducting, capacitor C_1 will charge through R_1 to voltage E. If a positive pulse is applied to the s.c.r. gate then C_1 will discharge into the primary of T_1, thus producing a voltage pulse in the secondary circuit. The load imposed by the spark plugs and stray losses is represented by resistance R_2 and capacitance C_2.

When C_1 is discharged into the transformer primary, the voltage rise on the windings is not instantaneous, due to the secondary stray capacitance, C_2. Referred to the primary winding, this capacitance may be expressed as $n^2 C_2$, where n is in the transformer turns ratio. Hence before the voltage E can be applied to L_1, assuming C_1 is fully charged, the capacitor $n^2 C_2$ must first be charged. If this operation is so rapid that negligible current flows in L_1 while it occurs, then, since the total quantity of stored electricity is the same before and after contact,

$$EC_1 = (C_1 + n^2 C_2) V_1$$
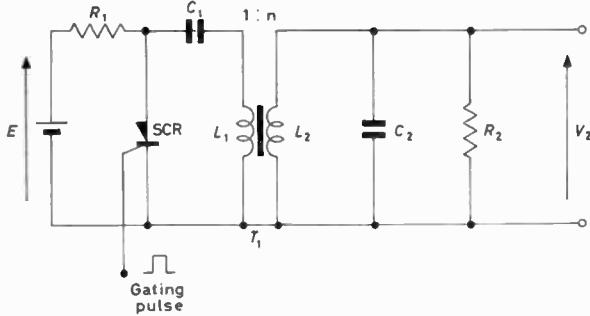
Fig. 2. Basic spark generating circuit

charged, this being determined by the time-constant C_1R_1.

It has been stated[1] that the voltage maximum should be reached as quickly as possible to obtain consistent ignition and to make the ignition circuit less dependent on external loading. This factor is governed by the natural frequency, ω_1 of the spark generating circuit and is dependent on C and L_2.

On examination of the circuit of Fig. 2, it will be seen that during the time that the s.c.r. is switched on, there is a constant drain on the source E through R_1. In order that this drain does not cause an excessive loss of power, the 'switch-on' time must be kept short; a value of about 100μsec being used in practice. Also the current through the s.c.r. must be kept sufficiently small so that it switches off at the end of the pulse. This fixes the minimum value
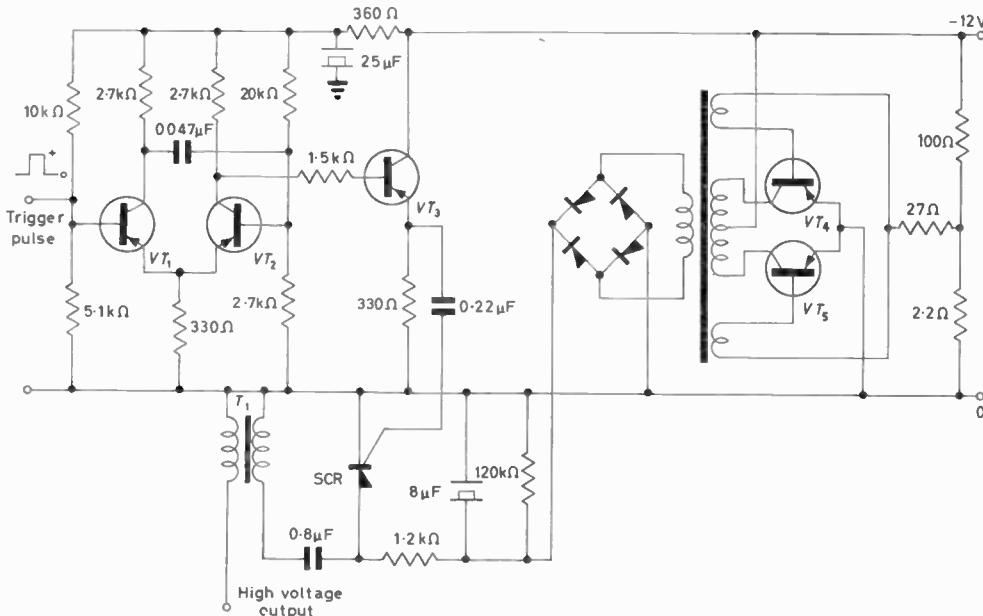
Fig. 3. Complete circuit of electronic ignition unit

or $V_1 = \dfrac{EC_1}{C_1 + n^2 C_2}$ volts

where V_1 is the voltage on C_1 and n^2C_2, immediately following the switching of the s.c.r.

The differential equation for the current in the secondary circuit is:

$$L_2(di_2/dt) + M(di_1/dt) + V_2 = 0$$

where i_1, i_2 and v_2 represent instantaneous values, and M is the mutual inductance between the transformer windings.

Now if the time t is taken to be zero when the capacitance $(C_1 + n^2 C_2)$ has reached voltage V_1 and if R_2 is assumed to be large, then the above equation yields the following solution:

$v_2 = nV_1 \exp(-t/CR_2) \cos \omega t$

where $\omega \simeq (1/L_2C)$

and $C = C_2 + (C_1/n^2)$

This gives the secondary voltage which will produce sparking. Its peak value will be nV_1. Also C_1 must be sufficiently large, compared with nC_2, so that V_1 is almost equal to E, when C_1 is fully charged. The output voltage versus spark repetition characteristic of the circuit will then be dependent simply on the voltage to which C_1 is
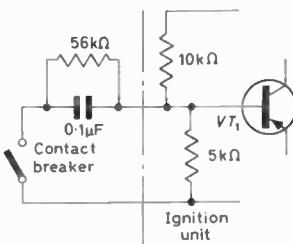

Fig. 4 (left). Recycling circuit for triggering from contact breaker

of R_1. However, since the discharge current is oscillatory, it was found that the s.c.r. was effectively switched off by the negative going part of the oscillatory current in the circuit, even when the d.c. in the s.c.r. was several times the holding current[2].

Practical Circuit

So far, no mention has been made of the actual values of voltage required for the ignition circuit. However, since it was not convenient to wind a suitable transformer at the time when this work was carried out, a conventional component as used in a battery-coil ignition system was obtained, and measurements first made on this. It was found that the output voltage was 27kV at low spark repetition frequencies and as this represented a fairly

typical value for modern ignition systems this value was also chosen for the proposed electronic ignition system.

The upper spark repetition frequency of the electronic system at which a useful spark is still produced, was fixed at 1kc/s to bring the unit into line with commercial electronic units at present on the market[3]. The only disadvantage in using the above type of transformer is that it has much higher values of winding inductance than is required in the proposed electronic system, hence the spark voltage rise time tends to be slower than that which could be obtained with a correctly designed component.

The transformer turns ratio, n, was found to be 100:1 and, taking account of losses, it was found necessary to use a value of E equal to 280V. Values of R_1 and C_1 were then chosen to obtain the required output voltage characteristic.
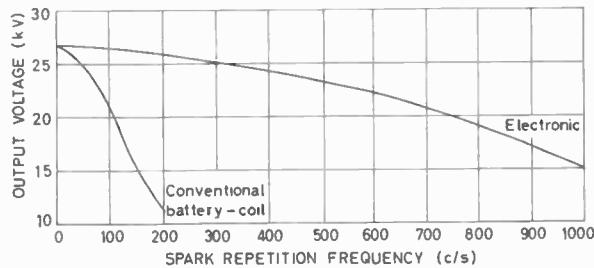

Fig. 5. Output voltage from ignition units
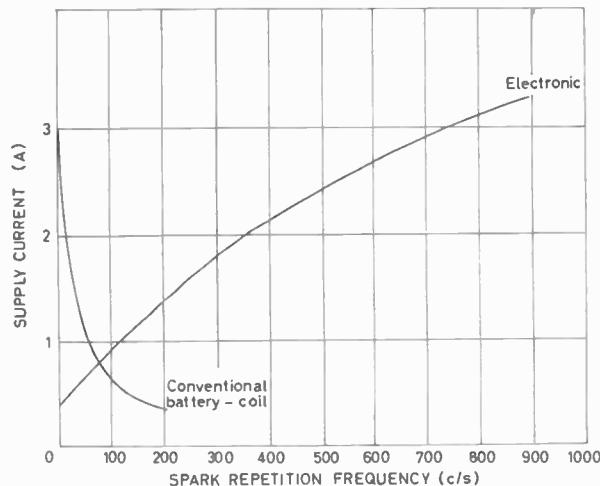
Fig. 6. Current drawn by ignition units from 12V supply

The complete circuit is shown in Fig. 3. Transistors VT_1 and VT_2 constitute an emitter coupled monostable multivibrator which is triggered by some form of pick-up, thus producing a pulse which is synchronously timed to the engine speed. Sufficient drive to switch the s.c.r. is provided by the transistor VT_3. A conventional transformer coupled d.c. to d.c. convertor is used to produce the charging supply of 280V from the usual 12V available from the engine electrical system.

The requirement of the pick-up system needed to switch the multivibrator, is that it should produce a positive pulse of value greater than 1V with a reasonably fast rise time, A suitable component would be of the electromagnetic type and has been adequately covered elsewhere[4].

Alternatively it is possible to trigger the circuit by shorting the input terminals. This can be carried out by the usual contact-breaker. However, it is necessary to include a circuit to give the unit a definite recycling time to

eliminate the effects of breaker ' bounce '; this is shown in Fig. 4. This arrangement was used to test the electronic ignition unit on a test-bench mounted engine.

Performance

With the electronic ignition unit triggered from a variable frequency oscillator, measurements of output voltage and current were made and are shown plotted against frequency in Figs. 5 and 6 respectively. Also plotted are similar measurements made on a conventional battery-coil ignition system.

Tests were made on an engine mounted on a test bench, using the contact-breaker to trigger the ignition unit. The
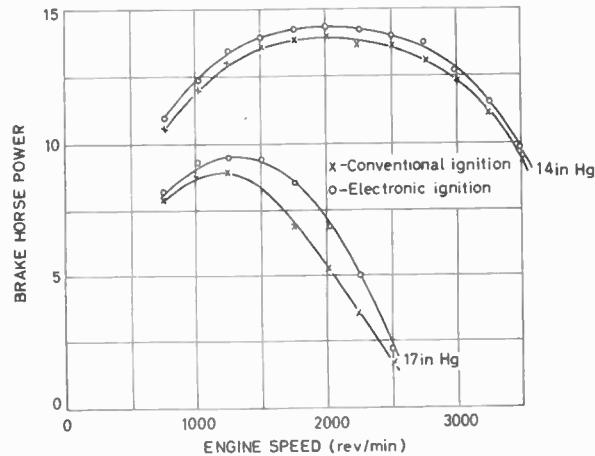

Fig. 7. Variation of b.h.p. with speed for Vauxhall engine at low throttle settings
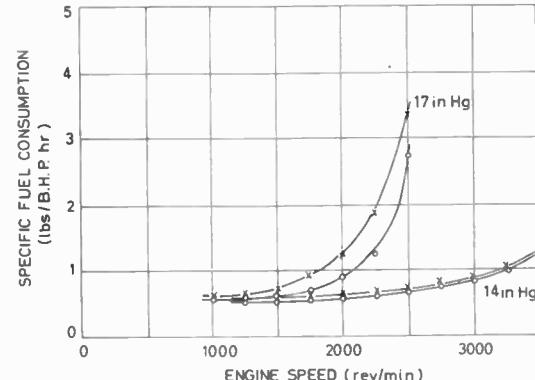
Fig. 8. Variation of specific fuel consumption with speed for Vauxhall engine at low throttle settings

engine used was a Vauxhall E.1X four cylinder engine. This could be loaded by a Heenan and Froude hydraulic dynamometer. Measurements were made of brake horsepower and brake specific fuel-consumption over the engine speed range for low throttle settings, since it was under these conditions that improvements would be most likely. The results obtained are plotted, together with those obtained with a conventional ignition system in Figs. 7 and 8.

The increase in b.h.p. and reduction in brake specific fuel consumption which were found, were no doubt due to the much better waveform of the electronic unit, leading to more consistent ignition. Subsequent measurements showed that the output voltage rise-time of the electronic unit was approximately half that of the conventional unit, and to this is attributed the improvements in engine performance, together with the higher values of spark voltage obtained over the complete speed range of the engine.

Conclusions

It has been shown that it is possible to build an electronic ignition unit which is capable of operating over a much wider speed range than is possible with the conventional battery-coil ignition system. Coupled with this, the advantages of reduced current drain on the battery at low speeds, faster voltage rise of the spark and ease of maintenance go together to make the electronic ignition unit a very attractive proposition in the design of higher speed petrol engines.

The major disadvantage of an electronic system is cost, but this could no doubt be reduced by the alternative use of a transistor circuit, as mentioned in the introduction. Here there would be fewer components and the overall cost would be less than a circuit using an s.c.r. It is thought that an identical performance to that given here could be obtained, with a transistor circuit.

An alternative approach, using an s.c.r., was tried. This utilized a triggered d.c. to d.c. convertor, so that capacitor C_1 could be charged during the time interval between successive sparks. This circuit had a constant voltage output over the required frequency range, but was not so reliable as the circuit described above and so was abandoned.

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The Demodulation of a Frequency Modulated Signal Modulated by Two Tones

By G. May*

The modulation process and the crowded spectrum of side frequencies that appears as a result of frequency-modulating a carrier with two tones has been discussed in several texts[1,2,3,4]. The converse process of demodulation, showing how the f.m. signal is converted to an amplitude-modulated wave in a typical discriminator has received, surprisingly, very little attention.

It is the purpose of this article to show how these side frequencies, particularly the combination frequencies, are used during demodulation.

(Voir page 867 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 873)

THE frequency spectrum of an f.m. wave modulated by two tones can be represented by the expression:

$$e = E_0 \sin (\omega_c t - m_{f1} \cos \omega_{m1} t - m_{f2} \cos \omega_{m2} t) \quad \ldots \ldots (1)$$

where $\omega_0/2\pi$ is the unmodulated carrier frequency, and m_{f1} and m_{f2} are the modulation indices due to the two modulating tones having frequencies of $\omega_{m1}/2\pi$ and $\omega_{m2}/2\pi$ respectively.

As there are inaccuracies in the expansions and expressions given by Crosby[1] and Starr[2] a complete expansion of equation (1) is given in Appendix (1), where it will be seen that the individual terms consist of side-waves having the form:

$$E_0 = J_p(m_{f1}) J_q(m_{f2}) [\sin(\omega_0 + p\omega_{m1} \pm q\omega_{m2}) t + \sin(\omega_c - p\omega_{m1} \pm q\omega_{m2}) t] \quad \ldots \ldots (2)$$

Here p and q are integers. When the sum of p and q is odd the sine terms become cosine terms.

It is also worth noting when comparing the generation of combination tones in two-tone f.m. and two-tone a.m. signals, that Black's[4] comparison, where he says that the super-position theorem is applicable to the a.m. case, is incorrect.

Demodulation of the Two Tone Modulated F.M. Wave

As shown in an earlier article[5], in a typical f.m. discriminator, the demodulation process is basically one of differentiation and it is achieved by using a transfer impedance that is a linear function of frequency.

If the modulated f.m. signal is considered to consist of sidewaves of the type given in expression (2) it is found that for every side-wave present at the input terminals of the transfer impedance there will be an output at the same frequency but its amplitude will be adjusted in such a way that it can be shown to be made up of five components (see Appendix (2)) having relative amplitudes as follows:

(1) An amplitude component containing the original Bessel coefficients.

(2) A component whose amplitude is equal to half the modulation factor m_{a1}. This is the lower side-frequency amplitude-modulation component of $f_c + (p + 1)f_{m1} + qf_{m2}$.

(3) Another component whose amplitude is $m_{a1}/2$. This is the upper side-frequency amplitude-modulation component of $f_0 + (p - 1)f_{m1} + qf_{m2}$.

(4) and (5). Two components each having an amplitude equal to half the modulation factor m_{a2}. These are the lower and upper side-frequency amplitude-modulation components of $f_0 + pf_{m1} + (q+1)f_{m2}$ and $f_0 + pf_{m1} + (q-1)f_{m2}$.

When the whole train of these components is related to the associated 'carriers' it will be found that one has an array of amplitude-modulated signals. Every side-wave is itself amplitude-modulated by the two tones f_{m1} and f_{m2}. The amplitude-modulation factors m_{a1} and m_{a2} are given by:

$$m_{a1} = sm_{c1} = sf_{d1}/f_c \quad \ldots \ldots \ldots \ldots (3)$$

$$m_{a2} = sm_{c2} = sf_{d2}/f_c \quad \ldots \ldots \ldots \ldots (4)$$

where f_{d1} and f_{d2} are the individual frequency deviations associated with f_{m1} and f_{m2}, respectively.

The side-waves at the combination frequencies

* Royal Military College of Science.
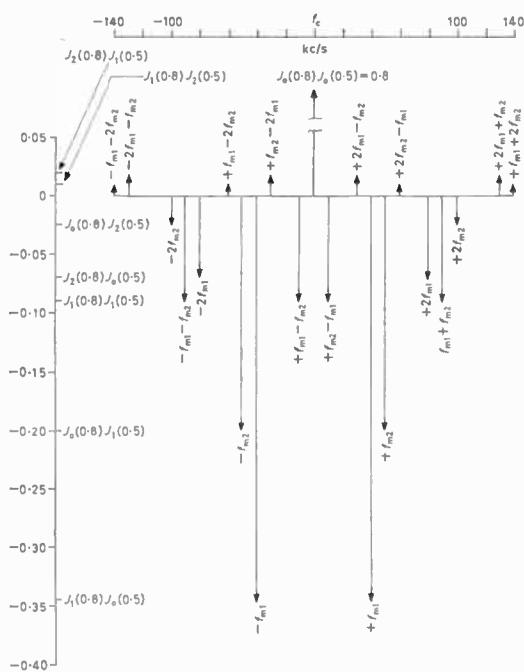
Fig. 1. Relative amplitudes of side-waves at input to the f.m. discriminator
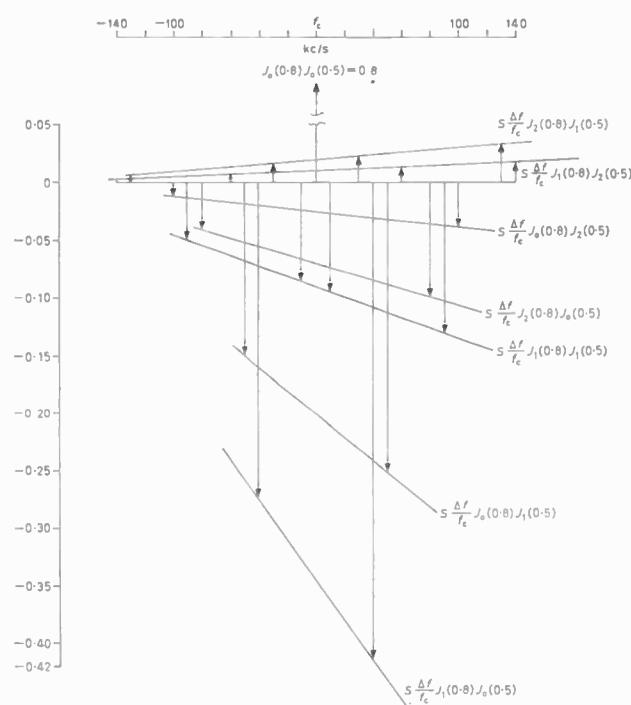
Fig. 2. Relative amplitude of side-waves applied to diode of Foster-Seeley discriminator

$f_c + pf_{m1} + qf_{m2}$, are seen to be a necessary addition to those at $f_c + pf_{m1}$ and $f_c + qf_{m2}$ in order to supply 'carriers' and side-frequencies to complete the array of amplitude-modulated signals.

On summing the voltage components of the array (see Appendix (3)) it will be found that the expression for the amplitude-modulated f.m. signal is:

$$e = E_0 (1 + m_{a1} \sin \omega_{m1} t + m_{a2} \sin \omega_{m2} t)$$
$$[\sin(\omega_c t - m_{f1} \cos \omega_{m1} t - m_{f2} \cos \omega_{m2} t)] \quad \ldots \ldots (5)$$

which when demodulated will produce the two original audio tones.

Taking as an example the passive network of a Foster-Seeley discriminator having a centre frequency, f_0, of 4·5Mc/s and a slope factor $s = 22·5$, and assuming the following signal and modulation conditions:

|		
Unmodulated carrier, f_c, equal to the discriminator centre frequency, f_0	$f_c = f_0 = 4·5\text{Mc/s}$	
Frequencies of modulating tones	$f_{m1} = 40\text{kc/s}$	$f_{m2} = 50\text{kc/s}$
Frequency deviations	$f_{d1} = 32\text{kc/s}$	$f_{d2} = 25\text{kc/s}$
Modulation indices	$m_{f1} = 0·8$	$m_{f2} = 0·5$
Modulation coefficients	$m_{c1} = 32/4500$	$m_{c2} = 25/4500$
Modulation factors	$m_{a1} = 0·16$	$m_{a2} = 0·125$
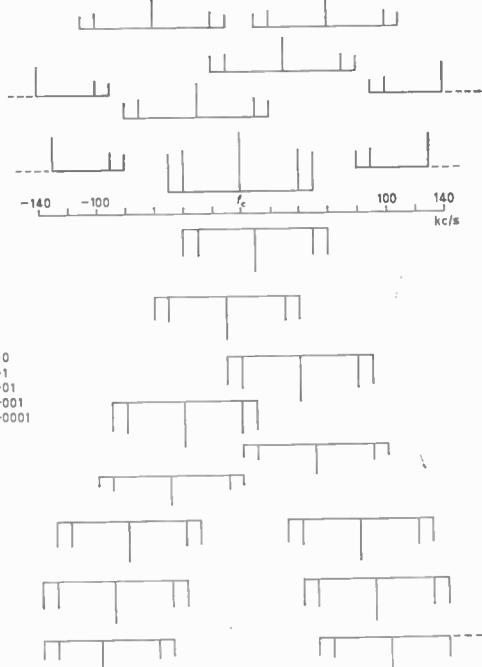
Fig. 3. Resolution of side-waves into a resultant array of two-tone amplitude modulated waves

The relative amplitudes of the side-waves are shown graphically, using linear scales, in Figs. 1 and 2. Fig. 3 shows how the side-waves of Fig. 2 can be displayed as a resultant array of amplitude-modulated waves. Note that the side-waves having Bessel coefficients of $J_3(0·8) J_0(0·5)$, $J_3(0·8) J_1(0·5)$, $J_1(0·8) J_3(0·5)$ and $J_2(0·8) J_2(0·5)$ have not been plotted in Figs. 1 and 2 as their relative amplitudes are less than 1 per cent and consequently difficult to display without using inordinately large scale factors. Although these side-waves are so small, they are required for the formation of side-frequency components for the 'carriers' having coefficients of $J_2(0·8) J_0(0·5)$, $J_2(0·8) J_1(0·5)$ and $J_1(0·8) J_2(0·5)$.

In Fig. 3, however, a logarithmic scale has been used and all the side-waves necessary for 99·9 per cent of the total energy[6] to be retained are displayed.

All the side-waves within the frequency bandwidth limits, in this particular case $2(f_{m1} + 2f_{m2})$, are necessary in the demodulation process if distortion is to be avoided.

The smallness of the resulting amplitude modulation factors, 0·16 and 0·25, emphasizes both the value of increasing the modulation coefficient, by translating the f.m. signal to a low intermediate frequency prior to detection and the advantage of keeping the slope factor, s, as large as possible consistent with discriminator linearity requirements.

APPENDIX

(1) EXPANSION OF THE EXPRESSION FOR A TWO-TONE MODULATED F.M. SIGNAL

The expression to be expanded is

$$e = E_0 \sin(\omega_c t - m_{f1}\cos\omega_{m1}t - m_{f2}\cos\omega_{m2}t) \quad \dots (6)$$

This can be expressed as

$$e = E_0 \left\{ \begin{array}{l} \sin\omega_c t \cos(m_{f1}\cos\omega_{m1}t)\cos(m_{f2}\cos\omega_{m2}t) \\ + \cos\omega_c t \sin(m_{f1}\cos\omega_{m1}t)\cos(m_{f2}\cos\omega_{m2}t) \\ + \cos\omega_c t \cos(m_{f1}\cos\omega_{m1}t)\sin(m_{f2}\cos\omega_{m2}t) \\ - \sin\omega_c t \sin(m_{f1}\cos\omega_{m1}t)\sin(m_{f2}\cos\omega_{m2}t) \end{array} \right\} \quad \dots (7)$$

Using the Bessel function expansions[7]:

$$\cos(x\cos\phi) = J_0(x) - 2J_2(x)\cos 2\phi + 2J_4(x)\cos 4\phi - \dots$$
$$\sin(x\cos\phi) = 2J_1(x)\cos\phi - 2J_3(x)\cos 3\phi + 2J_5(x)\cos 5\phi - \dots$$

then after straightforward but somewhat tedious manipulation one obtains:

$e = E_0$

$$\left\{ \begin{array}{l}
+ J_0(m_{f1})J_0(m_{f2})\sin\omega_c t \\
- J_1(m_{f1})J_0(m_{f2})\cos(\omega_c \pm \omega_{m1})t \\
- J_2(m_{f1})J_0(m_{f2})\sin(\omega_c \pm 2\omega_{m1})t \\
+ J_3(m_{f1})J_0(m_{f2})\cos(\omega_c \pm 3\omega_{m1})t \\
+ J_4(m_{f1})J_0(m_{f2})\sin(\omega_c \pm 4\omega_{m1})t \\
- \dots\dots\dots\dots\dots\dots\dots \\
- J_0(m_{f1})J_1(m_{f2})\cos(\omega_c \pm \omega_{m2})t \\
- J_0(m_{f1})J_2(m_{f2})\sin(\omega_c \pm 2\omega_{m2})t \\
+ J_0(m_{f1})J_3(m_{f2})\cos(\omega_c \pm 3\omega_{m2})t \\
+ J_0(m_{f1})J_4(m_{f2})\sin(\omega_c \pm 4\omega_{m2})t \\
- \dots\dots\dots\dots\dots\dots \\
- J_1(m_{f1})J_1(m_{f2})\sin(\omega_c + \omega_{m1} \pm \omega_{m2})t \; + \sin(\omega_c - \omega_{m1} \pm \omega_{m2})t \\
+ J_1(m_{f1})J_2(m_{f2})\cos(\omega_c + \omega_{m1} \pm 2\omega_{m2})t + \cos(\omega_c - \omega_{m1} \pm 2\omega_{m2})t \\
+ J_1(m_{f1})J_3(m_{f2})\sin(\omega_c + \omega_{m1} \pm 3\omega_{m2})t \; + \sin(\omega_c - \omega_{m1} \pm 3\omega_{m2})t \\
- \dots\dots\dots\dots\dots\dots\dots\dots \\
+ J_2(m_{f1})J_1(m_{f2})\cos(\omega_c + 2\omega_{m1} \pm \omega_{m2})t + \; \cos(\omega_c - 2\omega_{m1} \pm \omega_{m2})t \\
+ J_2(m_{f1})J_2(m_{f2})\sin(\omega_c + 2\omega_{m1} \pm 2\omega_{m2})t + \; \sin(\omega_c - 2\omega_{m1} \pm 2\omega_{m2})t \\
- J_2(m_{f1})J_3(m_{f2})\cos(\omega_c + 2\omega_{m1} \pm 3\omega_{m2})t + \cos(\omega_c - 2\omega_{m1} \pm 3\omega_{m2})t \\
- \dots\dots\dots\dots\dots\dots\dots\dots \\
+ J_3(m_{f1})J_1(m_{f2})\sin(\omega_c + 3\omega_{m1} \pm \omega_{m2})t + \; \sin(\omega_c - 3\omega_{m1} \pm \omega_{m2})t \\
- J_3(m_{f1})J_2(m_{f2})\cos(\omega_c + 3\omega_{m1} \pm 2\omega_{m2})t + \cos(\omega_c - 3\omega_{m1} \pm 2\omega_{m2})t \\
- J_3(m_{f1})J_3(m_{f2})\sin(\omega_c + 3\omega_{m1} \pm 3\omega_{m2})t + \; \sin(\omega_c - 3\omega_{m1} \pm 3\omega_{m2})t \\
+ J_3(m_{f1})J_4(m_{f2})\cos(\omega_c + 3\omega_{m1} \pm 4\omega_{m2})t + \cos(\omega_c - 3\omega_{m1} \pm 4\omega_{m2})t \\
+ \dots\dots\dots\dots\dots\dots\dots\dots \\
- J_4(m_{f1})J_1(m_{f2})\cos(\omega_c + 4\omega_{m1} \pm \omega_{m2})t + \; \cos(\omega_c - 4\omega_{m1} \pm \omega_{m2})t \\
- J_4(m_{f1})J_2(m_{f2})\sin(\omega_c + 4\omega_{m1} \pm 2\omega_{m2})t + \; \sin(\omega_c - 4\omega_{m1} \pm 2\omega_{m2})t \\
+ J_4(m_{f1})J_3(m_{f2})\cos(\omega_c + 4\omega_{m1} \pm 3\omega_{m2})t + \cos(\omega_c - 4\omega_{m1} \pm 3\omega_{m2})t \\
+ \dots\dots\dots\dots\dots\dots\dots\dots
\end{array} \right\}$$

$$\dots\dots\dots (8)$$

As there is some confusion over the occurrence of sine and cosine terms in the expansions quoted in references 1 and 2 it is perhaps of value to list the occurrence of these terms for various combinations of carrier and modulating signals.

Modulated wave	Nature of terms in expansion
$\sin(\omega_c + m_{f1}\cos\omega_{m1} + m_{f2}\cos\omega_{m2})t$	Sines and cosines
$\cos(\omega_c + m_{f1}\sin\omega_{m1} + m_{f2}\sin\omega_{m2})t$	All cosines
$\cos(\omega_c + m_{f1}\cos\omega_{m1} + m_{f2}\cos\omega_{m2})t$	Sines and cosines
$\sin(\omega_c + m_{f1}\sin\omega_{m1} + m_{f2}\sin\omega_{m2})t$	All sines

(2) DERIVATION OF THE RELATIVE AMPLITUDES OF THE AMPLITUDE-MODULATION SIDE-FREQUENCY COMPONENTS

Consider the current component corresponding to one of the four voltage components of expression (2), viz:

$$i = I_c J_p(m_{f1})J_q(m_{f2})\sin(\omega_c + p\omega_{m1} + q\omega_{m2})t \quad \dots (9)$$

Applying this to the transfer impedance[5], Z_{TA}, given by

$$|Z_{TA}| = aR_D[1 + s\Delta f/f_0] \quad \dots (10)$$

R_D, a and s are constants for the particular discriminator and Δf is the difference between the frequency concerned and the centre frequency, f_0.

$$e = E_0[1 + (s\Delta f/f_0)]J_p(m_{f1})J_q(m_{f2})\sin(\omega_0 + p\omega_{m1} + q\omega_{m2})t$$
$$\dots\dots (11)$$

Replacing aI_cR_D by E_0, Δf by $(pf_{m1} + qf_{m2})$ and letting the carrier frequency f_c coincide with f_0 equation (11) can be re-written as:

$$e_0 = E_0[1 + s(pf_{m1} + qf_{m2})/f_c]J_p(m_{f1})J_q(m_{f2})$$
$$\sin(\omega_c + p\omega_{m1} + q\omega_{m2})t \quad \dots (12)$$

From equation (12) one can extract the term:

$$E_0[s(pf_{m1} + qf_{m2})/f_c]J_p(m_{f1})J_q(m_{f2})\sin(\omega_c + p\omega_{m1} + q\omega_{m2})t$$

In order to convert the amplitude function $s(pf_{m1} + qf_{m2})/f_c$ of this term into the desired modulation factors it is necessary to employ the Bessel identity[7]:

$$J_p(m_{f1}) = (m_{f1}/2p)[J_{(p+1)}(m_{f1}) + J_{(p-1)}(m_{f1})] \quad \dots (13)$$

Multiplying both sides of the identity by spf_{m1}/f_c and remembering that $m_{f1} = f_{d1}/f_{m1}$:

$$(spf_{m1}/f_c)J_p(m_{f1}) = (spf_{m1}/f_c)(f_{d1}/2pf_{m1})$$
$$[J_{(p+1)}(m_{f1}) + J_{(p-1)}(m_{f1})] \quad \dots (14)$$

Also it has been shown that $sf_{d1}/f_{m1} = sm_{c1} = m_{a1}$ so that equation (14) can be re-stated as:

$$(spf_{m1}/f_c)J_p(m_{f1}) = (m_{a1}/2)[J_{(p+1)}(m_{f1}) + J_{(p-1)}(m_{f1})]$$
$$\dots\dots\dots (15)$$

and similarly

$$(sqf_{m2}/f_c)J_q(m_{f2}) = (m_{a2}/2)[J_{(q+1)}(m_{f2}) + J_{(q-1)}(m_{f2})]$$
$$\dots\dots\dots (16)$$

Substituting equations (15) and (16) into equation (12) five voltage components are obtained all having the same frequency $f_c + pf_{m1} + qf_{m2}$.

$e_0 = E_0$

$$\left\{ \begin{array}{l}
J_p(m_{f1})J_q(m_{f2})\sin(\omega_c + p\omega_{m1} + q\omega_{m2})t \\
+ (m_{a1}/2)J_{(p+1)}(m_{f1})J_q(m_{f2})\sin(\omega_c + p\omega_{m1} + q\omega_{m2})t \\
+ (m_{a1}/2)J_{(p-1)}(m_{f1})J_q(m_{f2})\sin(\omega_c + p\omega_{m1} + q\omega_{m2})t \\
+ (m_{a2}/2)J_p(m_{f1})J_{(q+1)}(m_{f2})\sin(\omega_c + p\omega_{m1} + q\omega_{m2})t \\
+ (m_{a2}/2)J_p(m_{f1})J_{(q-1)}(m_{f2})\sin(\omega_c + p\omega_{m1} + q\omega_{m2})t
\end{array} \right\}$$
$$\dots\dots\dots (17)$$

Thus in addition to the component, a 'carrier', containing the original Bessel coefficient $J_p(m_{f1})J_q(m_{f2})$, there are generated the four amplitude-modulation side-waves associated with the carrier waves having frequencies of $f_c + (p \pm 1)f_{m1} + qf_{m2}$ and $f_c + pf_{m1} + (q \pm 1)f_{m2}$.

Note that in using the Bessel identity (13) the following further identity[7] will be required in some instances:

$$J_{-n}(m_f) = (-1)^n J_n(m_f)$$

(3) SUMMATION OF THE VOLTAGE COMPONENTS DEVELOPED ACROSS THE DISCRIMINATOR TRANSFER IMPEDANCE

Taking as an example a specific current component:

$$i_{11} = -I_c J_1(m_{f1})J_1(m_{f2})\sin(\omega_0 + \omega_{m1} + \omega_{m2})t \quad \dots (18)$$

from the expansion of $i = I_c\sin(\omega_c t - m_{f1}\cos\omega_{m1}t - m_{f2}\cos\omega_{m2}t)$ instead of the general term given by equation

(9) it is found from equation (17) that the voltage due to this component is:

$$e_{11} = E_0 \left\{ \begin{array}{l} -J_1(m_{f1}) J_1(m_{f2}) \sin(\omega_c + \omega_{m1} + \omega_{m2})t \\ -(m_{a1}/2) J_2(m_{f1}) J_1(m_{f2}) \sin(\omega_c + \omega_{m1} + \omega_{m2})t \\ -(m_{a1}/2) J_0(m_{f1}) J_1(m_{f2}) \sin(\omega_c + \omega_{m1} + \omega_{m2})t \\ -(m_{a2}/2) J_1(m_{f1}) J_2(m_{f2}) \sin(\omega_c + \omega_{m1} + \omega_{m2})t \\ -(m_{a2}/2) J_1(m_{f1}) J_0(m_{f2}) \sin(\omega_c + \omega_{m1} + \omega_{m2})t \end{array} \right\}$$

$$\ldots\ldots\ldots\ldots (19)$$

If this process is repeated it will be seen that for every one of the side-waves forming the two-tone modulated f.m. signal applied to the discriminator there emerges a carrier complete with four amplitude-modulation side-frequency components.

The carrier component of equation (19) will be found to have its corresponding amplitude-modulation components

$$-(m_{a1}/2) E_0 J_1(m_{f1}) J_1(m_{f2}) \cos(\omega_c + \omega_{m2})t$$
$$+(m_{a1}/2) E_0 J_1(m_{f1}) J_1(m_{f2}) \cos(\omega_c + 2\omega_{m1} + \omega_{m2})t$$
$$-(m_{a2}/2) E_0 J_1(m_{f1}) J_1(m_{f2}) \cos(\omega_c + \omega_{m1})t$$
$$+(m_{a2}/2) E_0 J_1(m_{f1}) J_1(m_{f2}) \cos(\omega_c + \omega_{m1} + 2\omega_{m2})t$$

and the amplitude-modulation components of equation (19) are the corresponding side-frequencies of the following 'carrier' components

$$+E_0 J_2(m_{f1}) J_1(m_{f1}) \cos(\omega_c + 2\omega_{m1} + \omega_{m2})t$$
$$-E_0 J_0(m_{f1}) J_1(m_{f1}) \cos(\omega_c \qquad + \omega_{m2})t$$
$$+E_0 J_1(m_{f1}) J_2(m_{f2}) \cos(\omega_c + \omega_{m1} + 2\omega_{m2})t$$
$$-E_0 J_1(m_{f1}) J_0(m_{f2}) \cos(\omega_c + \omega_{m1} \qquad)t$$

When the sum of all the terms similar to e_{11} of equation (19) is considered it will be found that one has a set of terms equivalent to the set formed by multiplying each term of the expansion given in equation (8) by the two-tone amplitude-modulation term $(1 + m_{a1} \sin \omega_{m1}t + m_{a2} \sin \omega_{m2}t)$. The complete expression for the voltage e_t developed across the transfer impedance is thus

$$e_t = E_0 (1 + m_{a1} \sin \omega_{m1}t + m_{a2} \sin \omega_{m2}t)$$
$$\sin (\omega_c t - m_{f1} \cos \omega_{m1}t - m_{f2} \cos \omega_{m2}t) \ldots (20)$$

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Electron Spin Resonance Equipment

Electron spin resonance equipment valued at £85 000 has been ordered from Decca Radar Ltd during the past year. Installations will be completed in 1964 at the Universities of Bristol, Glasgow and Nottingham, and at the National Chemical Laboratory, Teddington. The supply of systems intended for control purposes in certain production processes is also under discussion.

In the design and manufacture of electron spin resonance equipment Decca has followed closely the recommendations of the panel, specially convened by the Department of Scientific and Industrial Research, to prepare a specification for spectrometers which would adequately meet most needs of research workers who require these instruments.

One of the panel's recommendations was for a particularly stable source of microwave energy derived from a crystal-controlled klystron generator. In the Decca spectrometers the klystron is phase-locked to the crystal source, resulting in a low noise content and absence of drift. In addition the latest techniques in low noise receivers have been utilized together with phase sensitive detectors of high conversion efficiency.

Decca Radar Ltd currently has in production a range of equipment in unit form which can be used in a 'building block' concept to provide the performance required from a finished system. Each unit in the range is completely independent and self-powered, and may be added to existing systems or interchanged between different systems. Items in the range include receivers, oscillators, klystron power supplies and phase locking units.

The complete system incorporates both straight and super-heterodyne microwave detection with switched change-over and modulation and phase-sensitive detection at 100kc/s and 33c/s. In some applications where high microwave powers are used it is necessary to compensate for small variations in cavity tuning, and there are facilities for controlling the klystron frequency in any of three ways, as selected by the operator:

(1) By phase-locking to a harmonic derived from a quartz crystal-controlled oscillator, as described above.

(2) By a combination of both cavity-locking and phase-locking retaining the stability derived from the latter but correcting for the effects of cavity drift.

(3) In the simple detection system, by locking to the sample cavity alone.

The units required for any locking system can also be supplied separately, and any combination of units may be specified.

Five different types of magnet systems are available, four with electromagnets and one with a permanent magnet.

Electron spin resonance spectrometers can be used to investigate minute samples in liquid, solid or even gaseous forms, and although this class of equipment is mainly in use in research laboratories it is already having important applications in industry. Decca Radar Ltd has been investigating and developing electron spin resonance techniques at their own research laboratories, and participation in work of this kind is a natural extension of the Company's specialized activities in microwave research. Considerable experience has already been gained in the design of high sensitivity, very low noise receivers for high performance radar and for space communications projects—a technique closely allied to the development of electron spin resonance equipment.

An Electronic Spectrometer for Use at Audio and Sub-Audio Frequencies

By D. Llanwyn Jones*

The theory of a simple spectrometer for use at audio and sub-audio frequencies is developed and the circuit selectivity and stability characteristics are deduced. A practical circuit is given and discussed with reference to the results of the theoretical analysis.

(Voir page 867 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 874)

ELECTRONIC systems designed to perform a spectrum analysis of a given electrical signal may be divided into two main groups. Firstly systems making use of a heterodyne principle may be considered. In these the input signal is mixed with a signal derived from a local oscillator and the spectrum analysis is accomplished by means of intermediate frequency filters consisting of tuned LC or crystal elements. The second large group contains the constant-Q systems in which a direct analysis of the input signal is achieved by means of suitable filters. The system considered here falls into this second group.

In the frequency band under consideration (0·1c/s to 10kc/s) it is not possible to employ a heterodyne system due to the fact that the carrier frequency cannot be adequately rejected by the intermediate frequency filters when the signal frequency is very low. In the audio frequency range it is possible to employ simple LC filters but the Q and off-tune selectivity of these is generally rather poor, particularly at the lower frequencies. At sub-audio frequencies such filters are impracticable due to the large component values required.

The spectrometer to be described employs feedback techniques to achieve a high selectivity, the frequency of operation being determined by an RC feedback network. The system is examined from the point of view of the selectivity attainable and from the point of view of the inherent stability. In the particular application of the author the spectrum of electromagnetic noise was being studied but the system is equally suited to other fields of application.

General Considerations

The basic requirement to be met by any spectrometer is that it should isolate a single frequency component of the received signal. In the case of constant-Q spectrometers this implies that the Q of the system should be as large as possible since Q is inversely related to the effective (noise) bandwidth of the system. In practice there comes a point when increasing the Q does not affect the output of the system[1], provided that the width of the spectrum of the signal is finite. If the input signal consists of an infinite train of similar waveforms the spectrum is a line spectrum, the width of each line being infinitesimally small. In this case the output of the spectrometer is always proportional to the Q of the system, however great this may be.

Besides meeting the condition that the Q of the system should be sufficiently large it is necessary to ensure that the 'skirts' of the gain-frequency characteristics have as large a slope as possible. This follows from a consideration of the fact that the output of the system at a given frequency f, is determined by the product of the transmittance of the system at the frequency f. $Y(f)$, and the spectrum component of the signal at frequency f, $G(f)$. If the spectrum component at a frequency f_0 is to be deter-

* King's College, University of London.

mined then it is necessary that:

$$G(f_0)Y(f_0) \gg G(f)Y(f) \text{ for all values of } f$$

In practice a selectivity of 40dB/decade or preferably 60dB/decade will ensure that the above condition is satisfied for the great majority of signals.
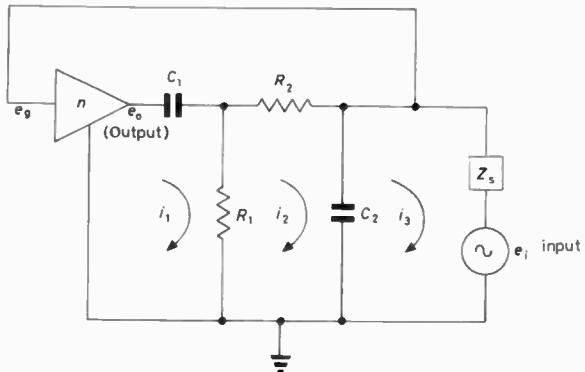

Fig. 1. Arrangement of spectrometer
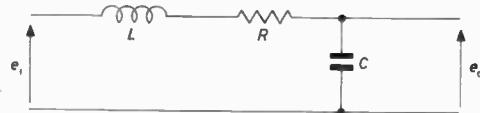
Fig. 2. Series L, C, R circuit (simple filter)

Theory of the Spectrometer

The schematic circuit of the spectrometer considered here is shown in Fig. 1.

An amplifier of forward gain n is connected to a positive feedback network $C_1R_1C_2R_2$ which produces a feedback voltage e_g. The signal to be analysed e_i is fed into the positive feedback loop via a series impedance Z_s. Mesh currents i_1, i_2 and i_3 flow as defined in the figure. It is assumed that the input impedance of the amplifier is large and the output impedance small compared with the impedances in the feedback loop. Hammond[2] has considered the case in which a differentiating-integrating network is employed in the feedback loop, but in what follows no restrictions are placed on the relative value of C_1, R_1, C_2 and R_2.

By application of Kirchoff's laws the following five equations defining the five unknowns i_1, i_2, i_3, e_g and e_0 are obtained.†

(a) $e_0' = ne_g'$

(b) $e_0' = n(i_2' - i_3')/pC_2$

(c) $e_0' = i_1'(1 + pT_1)/pC_1 - i_2'R_1$

† *The prime e.g. e_0' is employed to indicate a Laplace transformation.*

(d) $e_o' = i_1'/pC_1 + i_2'(1 + pT_2)/pC_2 - i_3'/pC_2$

(e) $e_i' = -i_3'(1 + pT_s)/pC_2 + i_2'/pC_2$

where $T_1 = C_1R_1$, $T_2 = C_2R_2$ and $T_s = C_2Z_s$

Solving for e_o' in terms of e_i' one obtains for the transfer function of the circuit:

$$e_o'/e_i' = nC_2[R_2(1+pT_1)+R_1] \begin{Bmatrix} p^2T_1T_2T_s + p(T_1T_2 + T_1T_s(1-n) \\ \ldots + T_2T_s + T_sC_2R_1) + T_2 + T_s + C_2R_1 \end{Bmatrix}^{-1} \cdots (1)$$

Considering now the series LCR circuit illustrated in Fig. 2 it may readily be shown that the transfer function is as follows:

$$e_o'/e_i' = \{(p^2/\omega_o^2) + (p/\omega_oQ) + 1\}^{-1}$$

with

$$\omega_o = 1/LC \text{ and } Q = \omega_oL/R \ldots\ldots (2)$$

This transfer function is illustrated in Fig. 3 and is known as the 'Quadratic Lag'. The selectivity of the circuit shown in Fig. 2 depends on the value of $Q = \omega_oL/R$ in the immediate vicinity of ω_o but the wide-band
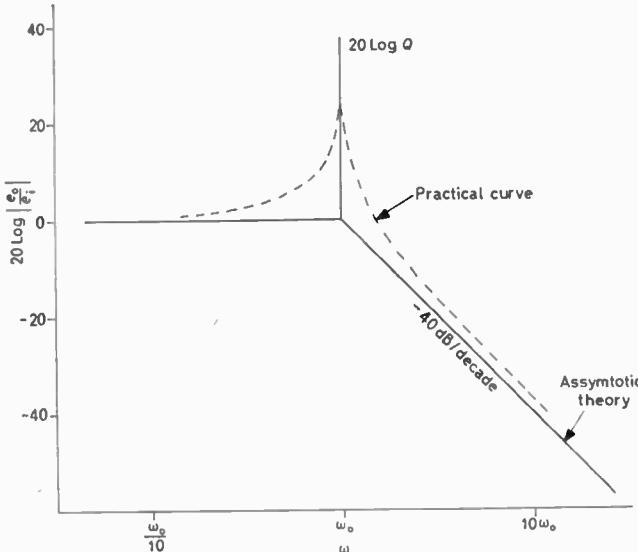
Fig. 3. The quadratic lag

selectivity has a value of 0dB/decade below ω_o and -40dB/decade above ω_o as shown in Fig. 3.

Comparing equations (1) and (2) it may be noted that the bracketed term in equation (1) is of the quadratic lag form. Since the initial term is only slowly varying with frequency in the region where the quadratic lag term reaches its maximum value one can consider the selectivity in the region of the tuned frequency to be determined by the bracketed term while the wide-band selectivity will be controlled by the selective properties of both the initial and bracketed terms.

In particular the case in which $C_1 = C_2 = C$, $R_1 = R_2 = R$, $T_1 = T_2 = T$ is now considered and it is assumed that $T_s \gg T$. Equation (1) now reduces to the following form:

$$e_o'/e_i' = \frac{nT(2 + pT)}{T_s} \{p^2T^2 + pT(3-n) + 1\}^{-1} \cdots (3)$$

Comparing equations (2) and (3) it can be seen that if the initial factor is neglected for the moment, the transfer function of the spectrometer circuit is similar to that of the series LCR circuit with:

$$\omega_o = 1/T = 1/CR \text{ and } Q = 1/(3 - n) \ldots\ldots (4)$$

Thus the Q of the circuit can be very large as $n \to 3$. For $n > 3$ the circuit will oscillate with an amplitude controlled by non-linearities.

Equation (3) may thus be written in the form:

$$e_o'/e_i' = \eta \{Q.L.\}$$

η is a frequency dependent term which is slowly varying with respect to the frequency variation of the quadratic lag ($Q.L.$). The wide-band selectivity in part depends on frequency variation of η, i.e. on the choice of the feeding impedance Z_s-resistive, capacitive or inductive. The choice of Z_s has a profound effect on the behaviour of the circuit particularly if the tuned frequency is to be continuously variable by employing ganged potentiometers for R or tunable capacitors for C. Two cases are considered below.

Case 1, Z_s Resistive, $Z_s = R_s$

In this case it follows from equation (1) that the transfer function is now:

$$e_o'/e_i' =$$
$$\frac{nR(2 + pCR)}{2R + R_s} \begin{Bmatrix} \frac{p^2C^2R^2R_s}{2R + R_s} + p\frac{[CR^2 + CRR_s(3-n)]}{2R + R_s} + 1 \end{Bmatrix}^{-1}$$

By comparison with equation (2) the resonant frequency and effective Q of the circuit are as follows:

$$\omega_o^2 = \frac{2R + R_s}{C^2R^2R_s} \text{ and } Q = \frac{\sqrt{[R_s(2R + R_s)]}}{R + R_s(3 - n)}$$

which reduce to the forms given in equation (4) for $R_s \gg R$. For $R_s \gg R$:

$$\eta = 2nR/R_s(1 + pCR/2)$$

as already stated, the overall wide-band selectivity of the system is determined by $\eta(Q.L.)$ and is thus zero below resonance and -20dB/decade above resonance.

At $\omega = \omega_o$ one may write $\eta = \eta_o$ where:

$$|\eta_o| = (nR/R_s)\sqrt 5$$

The ratio e_o'/e_i' at resonance is $|\eta_o|Q$ and is thus:

$$|e_o'/e_i'|_{\omega = \omega_o} = (nQR/R_s)\sqrt 5 \simeq \frac{3\sqrt 5 . QR}{R_s} \quad (5a)$$

The important point here is that the resonant output of the system is proportional to R. If the spectrometer is continuously tuned by varying R then the output of the system to a sinusoidal input will be inversely related to the resonant frequency $f_o = 1/2\pi CR$. If, however, the system is tuned by means of a variable capacitor then the output at resonance will be independent of f_o. At low frequencies potentiometers have to be employed for tuning since the required capacitance is large. The variation of the output with f_o is likely to be a disadvantage and for this reason the case of capacitive injection of the signal is considered.

Case 2, Z_s Capacitive, $Z_s = 1/pC_s$

From equation (1) with $T_s = C_2/pC_s$ it follows that:
$$e_o'/e_i' = npC_sR(2 + pCR)\{p^2T^2(1 + C_s/C) + pT[3 - n + 2C_s/C] + 1\}^{-1}$$

and hence:

$$\omega_o^2 = \frac{1}{T^2(1 + C_s/C)} \text{ and } Q = \frac{\sqrt{(1 + C_s/C)}}{3 - n + 2C_s/C}$$

In this case η is given by:

$$\eta = 2npC_sR(1 + pCR/2)$$
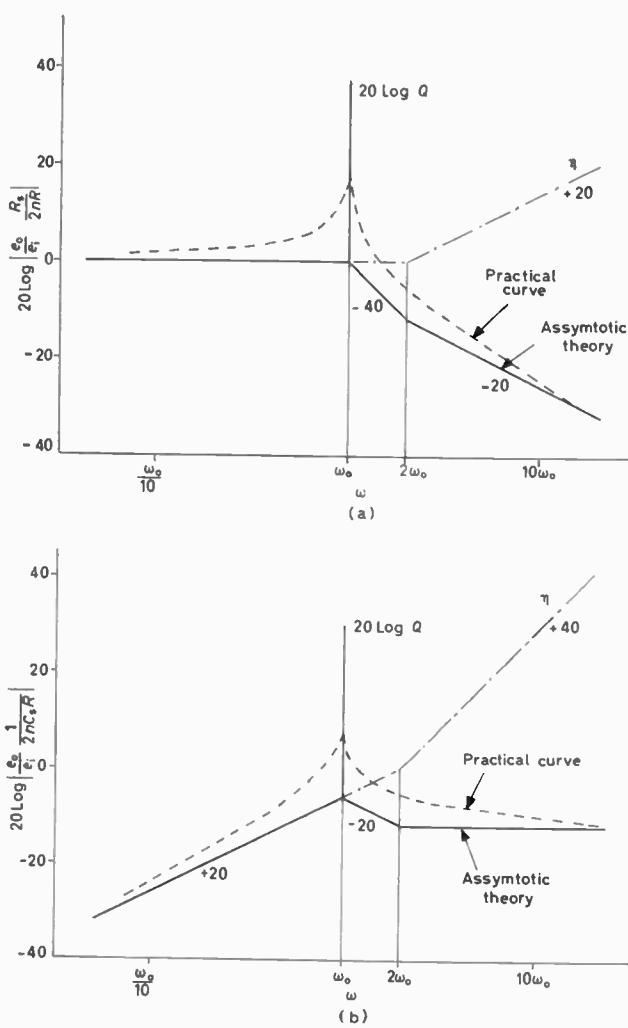
Fig. 4. Gain-frequency characteristic of spectrometer
(a) Z_s resistive
(b) Z_s capacitive

The overall selectivity of the system determined by $\eta(Q.L.)$ is thus $+20$dB/decade for $\omega \ll 2/T$ and is zero for $\omega \gg 2/T$.

As before, writing η_0 for the value of η at resonance:

$$|\eta_0| = (nC_s/C)\sqrt{5} \simeq \frac{3\sqrt{5}.C_s}{C}$$
$$\ldots\ldots\ldots (5b)$$

It can be seen that in the case of capacitive injection of the input signal the output of the system at resonance is unaffected by the value of R. In other words the spectrometer may be tuned by varying R without affecting the output. If C is varied the output of the system at resonance is inversely proportional to C.

The frequency characteristics of the circuits considered in the two cases above are shown in Figs. 4(a) and (b). These figures illustrate the points made regarding the wide-band selectivity of the two systems.

Improvement of Skirt Selectivity

Generally a skirt selectivity of 40 or, better, 60dB/octave will be necessary in order to prevent signal breaking through to the output of the spectrometer. When only single frequency operation is required the desired skirt selectivity may be attained by employing a cascaded series of RC low-pass and high-pass filters in front of the spectrometer. In practice such filters will have a sharp enough cut-off provided that the impedance of each filter is ten times the impedance of the preceding filter. Cathode- or emitter-followers will generally be necessary in order to avoid loading problems.

If the spectrometer is to be tuned over a range of frequencies it is necessary to design a bandpass filter to give the required skirt selectivity while allowing frequencies within the bandpass to be transmitted. Again RC filters are suitable.

Stability

In designing a suitable system the chief problem to be overcome is that of obtaining a stable Q since the skirt rejecting properties of the circuit are stable due to the nature of the components employed.

From equation (4) it follows that the stability of the Q of the system is related to the stability of the forward gain of the amplifier, n. If n changes by an amount Δn, Q will change by an amount ΔQ where:

$$\Delta Q = Q^2 \Delta n$$

The forward gain of the amplifier will generally be stabilized by employing negative feedback. If an amplifier of open-loop gain α is used and a fraction-β of the output voltage is fed back to control the gain then it is well known that the closed-loop gain is:

$$n = \alpha/(1 + \alpha\beta)$$

If it is supposed that the change Δn arises due to changes $\Delta\alpha$ in α and $\Delta\beta$ in β it then follows that ΔQ is given by:

$$\Delta Q/Q = Qn^2/\alpha \left\{(\Delta\alpha/\alpha) - \alpha\Delta\beta\right\} \ldots\ldots (6)$$

The problem of achieving a stable Q is thus that of stabilizing α and β. As discussed below $\Delta\beta$ will normally be

Fig. 5. Practical spectrometer circuit
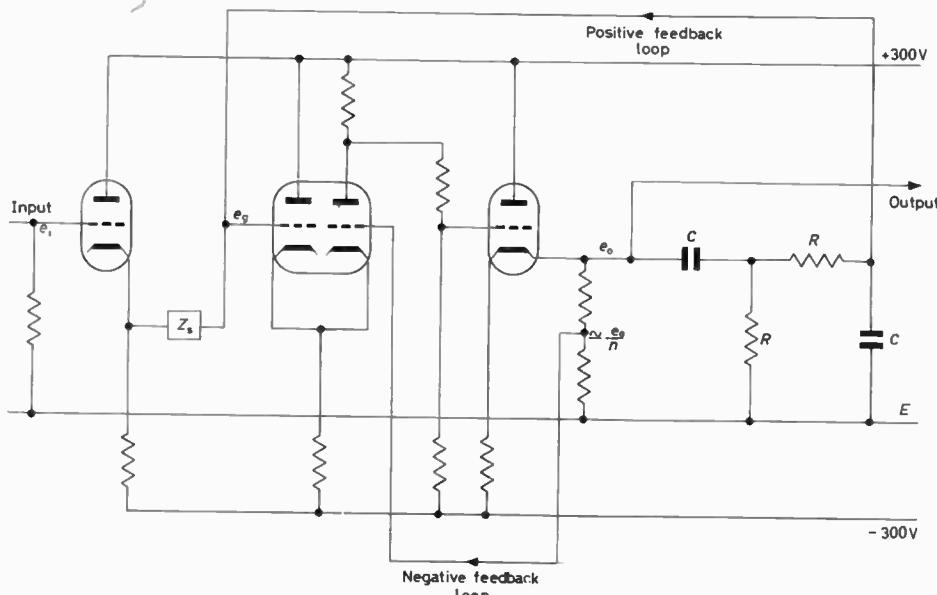
negligible and thus a stable Q will be achieved when α is large.

Practical Circuit

A simple practical circuit is shown in Fig. 5.

The circuit employs a long-tailed cathode-coupled triode amplifier with cathode-follower output. The input signal is fed into the system via a buffer stage. The forward gain and hence the Q of the circuit is controlled by the degree of negative feedback applied by the potential divider in
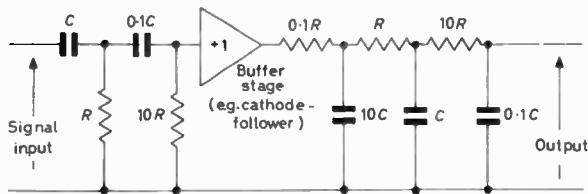

Fig. 6. Arrangement of pre-filter

the cathode circuit. In the spectrometer employed by the author the tuned frequency was automatically swept through one decade of the spectrum by employing a precision dual gang potentiometer for R and rotating it by means of a synchronous motor. The spectrum was plotted by means of a pen recorder connected to the output.

The effective Q of the circuit may be measured and set to the required value by measuring the ratio $|e_o/e_i|$ at resonance. From equation (5a) and (5b) it follows that:

$$Q = R_s/R3\sqrt{5}\ |e_o'/e_i'|_{\omega=\omega_o} \text{ for } Z_s = R_s$$

$$Q = C/C_s3\sqrt{5}\ |e_o'/e_i'|_{\omega=\omega_o} \text{ for } Z_s = 1/pC_s$$

From which Q may be determined.

The stability of the system limits the Q that can be attained. Using stabilized power supplies variations in α are of the order of 1 per cent on a short time basis and variations in β are made very small by using resistors with equal resistance-temperature coefficients in the negative feedback potential divider. Taking α as 50 and assuming the variations of 10 per cent in Q are the maximum allowable it follows from equation (6) that the maximum Q that can usefully be employed is about 50. Higher stability may be achieved by employing an amplifier of higher gain, e.g. by adding a pentode amplifier after the double triode and taking the signal from the first anode of the double triode rather than the second to maintain the required phase relations. In the circuit shown a Q value of around 30 stable to within 5 per cent over a period of 24 hours could be achieved.

In order to obtain a skirt selectivity of 60dB/decade the pre-filter shown in Fig. 6 may be used. This circuit is designed for use with a capacitive Z_s.

Acknowledgments

The author wishes to thank Dr. F. W. Chapman for many helpful suggestions incorporated in this article.

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An Electronic Ticket Recognition System

An electronic ticket recognition passenger control system for public transport has recently been introduced by Ticket Equipment Ltd of Cirencester, Gloucestershire, and is now being installed in the transport services of Manchester and Frankfurt-on-Main, West Germany.

The Unimat automatic ticket issuing machine (left) with the electronic ticket cancelling machine at the entrance to a bus on the right
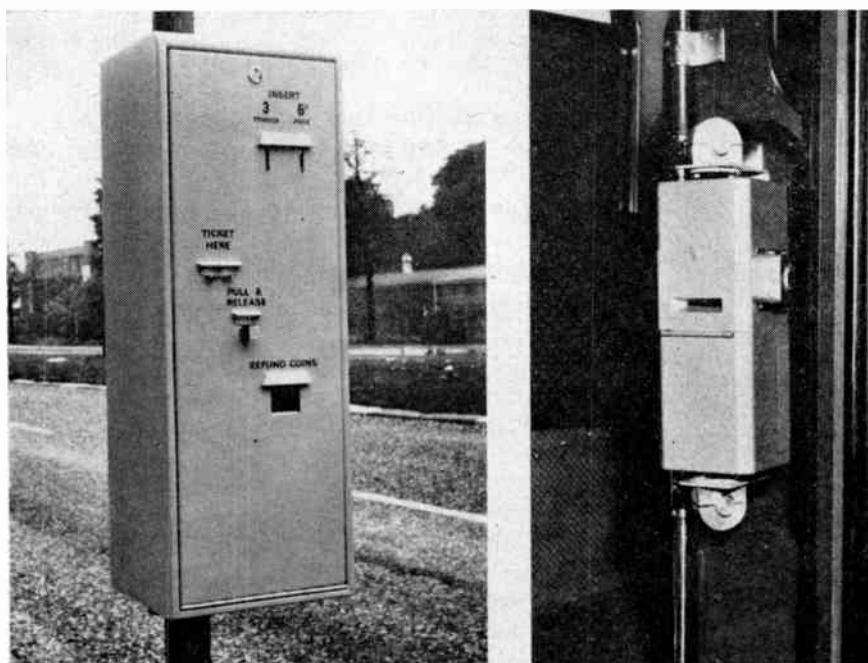
The system is designed for operation on one man driver operated buses and passengers will buy their tickets in advance from automatic ticket issuing machines, which are located at termini and intermediate points along the bus route.

On entry to the bus, the passenger inserts the ticket, which is of a special forgery-proof design and carries a metallic spot at one end, into a conspicuously marked slot on the electronic ticket-recognition machine installed at the bus entrance near to the driver. The machine 'reads' the metallic spot and if the ticket is valid, a small guillotine automatically cuts off the end of the ticket with the metallic spot which is retained for later checking.

At the same time the ticket is stamped with the route number, date, time and fare stage, the combination of which makes it impossible for an invalid ticket to pass the scrutiny of travelling inspectors at any stage of the journey.

If a ticket is inserted incorrectly, or after the spot has been cut off, or if any other piece of card is used, the driver is alerted by a warning device. Passengers who board without pre-purchased tickets are issued with conventional tickets from a driver operated machine.

When transfer or multiple journey tickets are used, additional metallic spots are printed on the ticket, one spot being guillotined on each journey.

It is expected that this system applied to one man driver buses operating on selected routes will considerably reduce the amount of time taken by the driver in issuing tickets including the taking of fares and the giving of change, and so speed up passenger flow, increase staff economies and prevent the use of fraudulent tickets.

SECAM Signal Decoding for Single-Gun Tubes

By P. Cassagne* and G. Melchior*

The use of the SECAM signal in single-gun tube receivers is examined. A new circuit for direct conversion, without demodulation of the chrominance signals, to CCS signals is described. This brings a simple and at least as effective a solution as that used with the NTSC signal.

In addition the same arrangement permits the construction of high-grade SECAM transcoders.

(Voir page 867 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 874)

ALTHOUGH so far the only three-colour reproduction tube in production is the three-gun masked tube, many laboratories[1,2,3,4] have been endeavouring to find solutions for single-gun tubes. By their very principle such tubes involve sequential display of the picture three-colour components, so necessitating that the viewer's eye shall integrate these components in space and time to ensure satisfactory reconstitution of the picture.

The corresponding circuit diagram is illustrated in Fig. 1 showing the cyclic switch which connects the electron
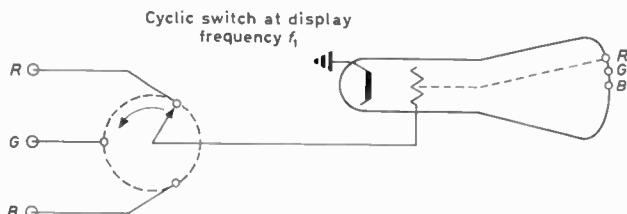
Fig. 1. Cyclic switch for single-gun tube

beam control electrode to the channel of the primary colour corresponding to the phosphor being scanned at that instant.

At this stage there is no apparent difference between the NTSC or SECAM transmission coding since each is in a position to supply the necessary three-colour components.

But such an arrangement has several practical drawbacks which, in particular, are due to the principle of the switch which does not function at constant luminance. The analysis of the multiplexing obtained in this way suggests other and more effective solutions.

By decomposing the signal in a Fourier series it is found[5] that the sampled signal obtained by this process is equivalent to a signal obtained by adding in equal proportions the primary components, on which there is superimposed a series of modulation components of three carriers whose mutual phase angle is 120°. Each of these carriers is modulated by each three-colour signal so that their sum constitutes a vector modulated in amplitude and phase which is cancelled for reference white. This decomposition is shown diagrammatically in Fig. 2, where it is assumed that the display frequency f_1 is sufficiently high to ensure that the partial spectra do not overlap.

Strictly speaking there are also components at harmonic frequencies of f_1 which can be neglected if sampling times are sufficiently short.

The results of the analysis show that it is possible to ensure faithful colours in the reproducing system by means of a such a composite signal, if:

(1) The amplitudes of the low and high frequencies are in the ratio 1:2.

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* Compagnie Française de Télévision.

(2) The carrier frequency and phase are closely related to the repetition period of the spot over the phosphor triplets.

When these conditions are satisfied the tube acts as a true decoder of the composite signal applied to its control electrode.

So, on the basis of this correspondence, it is possible to envisage the construction of receivers which will not
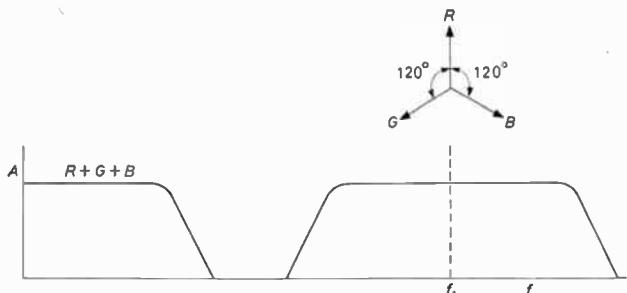

Fig. 2. Decomposition of colour signal

require the reconstitution of the primary signals and the corresponding switch. It will be apparent that the type of coding used for transmission may introduce important differences in circuits designed to convert the received composite signal.

The purpose of this note is to describe a mode of conversion obtainable from the SECAM signal, which considerably simplifies the processes so far considered and which compares favourably with the corresponding devices applicable to the NTSC signal.

Principle of Single-Gun Tube Receivers

The signals required for application to a single-gun tube consist, in most cases, of:

(1) A monochrome signal carrying the high definition,

$$E_M' = (E_R' + E_G' + E_B')/3$$

(2) A colour signal consisting of a sub-carrier simultaneously modulated in amplitude and phase:

$$E_k' = E_R' \sin \omega_1 t + E_G' \sin (\omega_1 t - 2\pi/3) + E_B' \sin (\omega_1 + 2\pi/3)$$

So receivers fitted with single-gun tubes will have to include a 'convertor' delivering, from the transmitted signals, the signals expressed above to be applied to the tube.

The signal E_M' is obtained by adding to the transmitted luminance signal E_Y' the difference signal:

$$E_D' = E_M' - E_Y'$$

where E_D' is a linear combination of the three primary signals, and which being cancelled in the absence of colour is therefore a component of chrominance which will always be easily obtained by a matrix operation from the two

received chrominance components. For example it is possible to calculate in terms of the red and blue chrominance components:

$$E_D' = 0.17 (E_R' - E_Y') + 0.27 (E_B' - E_Y')$$

Thus a convertor must essentially consist of:

A channel delivering E_D'

A colour channel delivering E_k'

these two signals having to be obtained from the signals transmitted by the system's colour channel.

Two solutions are available for setting up the convertor colour channel:

(1) Demodulation of chrominance and remodulation of a local sub-carrier.

(2) Direct conversion without demodulation.

In the latter case the principle of the convertor depends essentially on the modulation process applied to the colour sub-carrier proper to the colour television system.
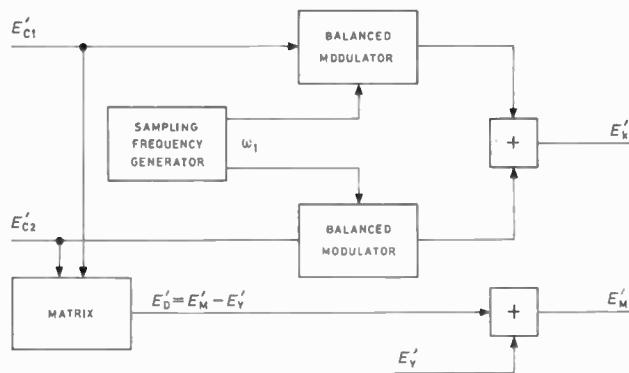
Fig. 3. Demodulation-remodulation convertor principle

Demodulation and Remodulation Convertor

GENERAL PRINCIPLE

The general principle of a convertor with demodulation and remodulation of chrominance is, for all colour television systems, shown in Fig. 3.

The incoming signals E_{c1}' and E_{c2}' are the two chrominance components transmitted and received by demodulation of the colour sub-carrier. They depend on the particular television system used. It may happen that one of these components corresponds to E_D', in this case the matrix is eliminated.

However, this matrix is simple and the main difficulty met with for such a convertor is undoubtedly the design of stable balanced modulators, and this difficulty remains whatever the system considered.

On the whole there remains only the sampling sub-carrier frequency generator which may substantially affect the cost price of a convertor, depending on the system used.

SAMPLING FREQUENCY GENERATOR

For certain tubes (index tubes), sampling has to be related to the co-ordinates of the impact of the beam on the screen. The sampling frequency is then obtained from a signal emitted by the tube itself, and in this case the circuits used will be the same for all systems.

For other tubes (e.g. chromatron) it is necessary to add to the convertor circuits for generating a sampling frequency whose phase is 'interlaced' with respect to the scanning periods.

Direct use of the colour sub-carrier frequency is not recommended since it is too low, but with the NTSC

system (more generally with any system in which the sub-carrier frequency is determined with respect to scanning) the frequency f_1 can be obtained relatively simply by multiplying the sub-carrier frequency.

The problem is different in the case of the SECAM system in which the sub-carrier is not directly usable on account of its modulation. Jackson[6] has proposed the use of an oscillator with its phase locked on the sub-carrier during line blanking (the sub-carrier frequency then being constant). Another solution giving an oscillation with its phase interlaced with respect to scanning seems to be simpler and has proved itself.

The oscillation is imposed on a crystal, but this crystal is not part of an oscillator, it is simply excited at each line by a short pulse. To obtain phase interlacing the crystal's proper frequency is an old multiple of line half-frequency and the exciting pulse is alternately positive and negative. It is obtained by differentiating a rectangular signal at
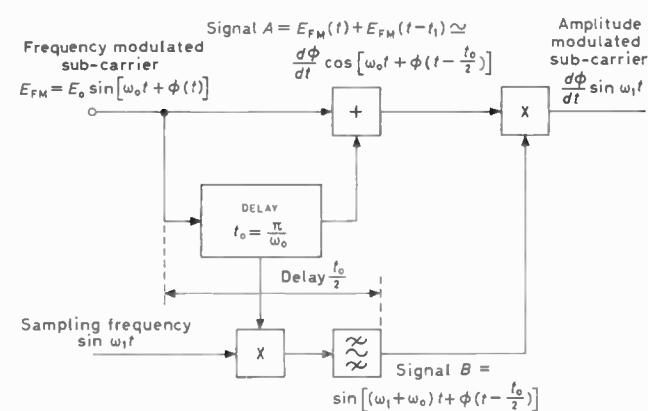
Fig. 4. F.M./A.M. convertor

line half-frequency which is always available in a SECAM receiver (switch control signal).

The only difference between the NTSC and SECAM systems is concerned with the sampling frequency generator for tubes of the chromatron type, obtainable in both cases by simple means, as shown above.

Conversion Without Demodulation

In order to avoid the use of balanced modulators in the receiver, it seems preferable to convert directly, i.e. without intermediate demodulation and remodulation, the colour sub-carrier transmitted to a 'circular modulation' CCS sub-carrier suitable for application to the single-gun tube. Ten years ago Loughlin[7] gave a solution applicable to the NTSC system. Direct conversion is also possible in the SECAM case in which the colour sub-carrier is frequency modulated sequentially for the two chrominance components. In this case an f.m.-a.m. convertor has to be used.

F.M.-A.M. CONVERTOR

A selective circuit in the particular frequency range can be considered as a convertor of frequency modulation to amplitude modulation, and most frequency discriminators are actually only a combination of an f.m.-a.m. convertor and an amplitude detector.

But in the present case a simple selective circuit is unsuitable, for the convertor requires that two additional conditions be satisfied.

(1) Modulation converted to amplitude has to be carried not by the applied frequency but by a local and

$$A = 2A_x \sin \frac{x}{2} = 2A_x \cos \frac{\pi}{2} \frac{f_x}{f_0}$$
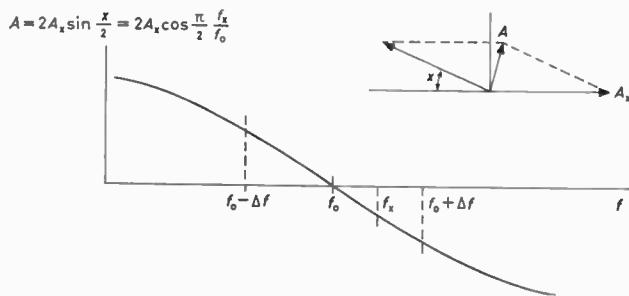
Fig. 5. Variation of amplitude as a function of frequency

constant frequency: the sampling frequency f_1 obtained, for example, as described previously.

(2) This amplitude modulation has to be of the 'suppressed sub-carrier' type as would be obtained from a balanced modulator.

The principle of an f.m.-a.m. convertor meeting these requirements is shown in Fig. 4.

It consists mainly of a delay line adjusted to produce a delay of half a period for the central frequency of the frequency-modulated wave. The result of adding the direct and the delayed signals is thus identically zero at this frequency.

But as soon as a frequency deviation appears phase opposition ceases to exist between the two direct and delayed signals, and their sum is a vector whose amplitude is proportional to the frequency deviation (at least, within the range of deviations under consideration). Fig. 5 shows the law of variation of the resulting vector amplitude A as a function of the instantaneous frequency f_x of the incoming signal. Reversal of polarity on passing through f_0 indicates the 180° phase change which takes place at this frequency. Thus the variations of incident frequency have been converted to correlative amplitude variations which have now to be transposed to the sampling frequency f_1.

This second operation is realized by two frequency transpositions. The first changes the frequency of the incoming signal around $f_0 + f_1$. It is the signal B.

There are then available two signals identically frequency-modulated, one B of constant amplitude, and the other A whose amplitude varies continually with the frequency deviation. All that is then required is to effect a second transposition of the latter by the former to obtain the signal carrying the amplitude modulation with suppressed carrier of the required frequency f_1 (see Appendix).

SECAM RECEIVER FOR SINGLE-GUN TUBE

It would be possible to design a SECAM receiver for single-gun tube by means of two f.m.-a.m. convertors as described above, each fed from an output of the sequential-simultaneous decoder of a conventional SECAM receiver, but it is simpler to use only one f.m.-a.m. convertor while effecting the sequential-simultaneous decoding on the a.m. signal.

The principle of the chromatic channel of such a receiver is shown in Fig. 6.

The modulation of the signal leaving the f.m.-a.m. convertor has to be de-emphasized. This is easily done by means of a selective circuit whose characteristic is the transposition about f_1 of the video de-emphasis of conventional SECAM receivers. Modulated signals R-Y and Y-B are thus alternately available at its output.

The circular sub-carrier signal can be expressed in terms of the two colour difference signals in the form[8]:

$$E_k' = K_R (R-Y) \sin \omega_1 t + K_B (B-Y) \sin (\omega_1 t + \phi) \quad \dots \dots (1)$$

where $K_R = -0.89$, $K_B = 0.74$, $\phi = 88°$.

Designating the signal delivered by the f.m.-a.m. convertor after de-emphasis as E_{DT}, it is readily shown that this quantity is alternately proportional to:

$$E_{TD}(t) + m E_{TD} (t - T + \phi/\omega_1) \quad \dots \dots (2)$$

and

$$E_{TD} (t + T + \phi/\omega_1) + p E_{TD}(t) \quad \dots \dots (3)$$

where T is the duration of a scanning line.

A suitable matrix operation on the signals obtained directly or through a 64μsec delay line and a switch operated at half line frequency will thus produce the required CCS sub-carrier signal.

The difference signal E_D' to be used for correcting the black-and-white channel is obtained by demodulating synchronously the CCS signal defined by equation (1) at an angle of $-27.5°$ with respect to the reference phase.

On the whole the complete SECAM sub-carrier to CCS sub-carrier convertor is no more complicated than the normal colour channel of a SECAM decoder and compares favourably with the facilities required for the NTSC signal.

Overall Picture Quality

Generation of the NTSC signal for single-gun tubes involves complete suppression of the incident colour sub-carrier components. This has to be done by low-pass filter, and it is well known that such a reduction of the luminance signal pass-band causes a substantial loss of resolution, even for reproduction tubes as at present available. This degradation will be even more apparent in the case of single-gun tubes.

The use in SECAM technique of different types of modulation for luminance and for chrominance permits eliminating the chrominance sub-carrier signal from the luminance signal without the need to introduce a low-pass

Fig. 6. Chromatic channel
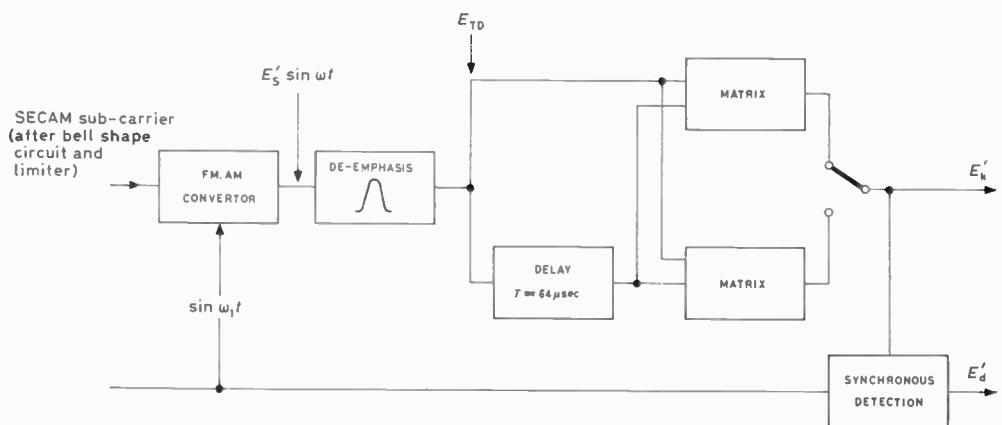
filter[9]. There results an appreciable improvement of picture resolution such as to ensure an intrinsic picture of higher quality than that obtained with NTSC signals.

Conclusion

Direct conversion of the SECAM colour signals to CCS suitable for application to a single-gun tube is possible and results in a simpler receiver than those which use the demodulation and remodulation technique.

In addition, the convertor system of Fig. 6 can be used for transcoding the SECAM signal to an NTSC signal with no intermediate demodulation of the colour signal, hence under conditions ensuring optimum quality.

APPENDIX

The signal to be converted may be written in the form:

$$E_{FM} = E_o \sin (\omega_o t + \phi(t))$$

where the amplitude E_o is assumed to be constant.

If the signal transmitted by frequency modulation is E_s', the instantaneous frequency of E_{FM} will be:

$$f_x = 1/2\pi \; (\omega_o + d\phi/dt) = F_o + E_s' \; \Delta F$$

The selective circuit which produces the f.m. to a.m. conversion is obtained by means of a delay line, the signal to be converted is added to the signal, the latter being delayed by a time $t_o = \pi/\omega_o$.

Then:

$$E_{FM}(t) + E_{FM}(t - t_o) = E_o \sin (\omega_o t + \phi(t)) + E_o \sin (\omega_o t - \pi + \phi(t - t_o))$$

$$= 2E_o \sin \frac{\phi(t) - \phi(t - t_o)}{2} \cos \left(\omega_o t + \frac{\phi(t) + \phi(t - t_o)}{2} \right)$$

A frequency change is effected on this signal, the hetero-dyne oscillation being:

$$\sin \; (\omega_o t + \omega_1 t + \phi(t + t_o/2) + \phi_o)$$

(ϕ_o being a constant phase).

The output signal is then:

$$E_T = E_o \sin \frac{\phi(t) - \phi(t - t_o)}{2} \sin \left[\omega_1 t + \phi(t - t_o/2) - \frac{\phi(t) + \phi(t - t_o)}{2} + \phi_o \right]$$

which is substantially equal to:

$$E_o = d\phi/dt \sin (\omega_1 t + \phi_o)$$

i.e.: apart from some fixed phase, proportional to: $E_s' \sin \omega_1 t$.

This expression is actually that for the signal which would be obtained from a balanced modulator fed with the sub-carrier frequency ω_1 and with the modulating signal E_s'.

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An Analogue Teaching Equipment

The increased use of analogue computation and simulation techniques in almost every field of research creates training problems. Early training—and practical training at that—is necessary for students from Sixth Form level onwards.

To meet this need a purpose-designed teaching equipment, known as the Analogue Tutor Mark 2, is now available from System Computers Ltd. of Newcastle-upon-Tyne.

This computer is an entirely self-contained, lightweight and portable unit with plug-in characteristics. It provides teaching authorities with the opportunity and facilities for introducing 'system investigation' and 'behaviour analysis' into biological, chemical, mechanical or electrical study subjects at low cost.

By virtue of its design the Tutor is easily read. Symbolic representation of the analogue circuits, engraved on the graphic patch panel, clearly demonstrates to students the basic principles of analogue computation and its effects within the circuit.

With the exception of the operational amplifiers, which are mounted in pairs on removable modules inside the console, the actual computing components including the feedback resistors and capacitors, input resistors, etc. are plugged directly into the face of the graphic patch panel. Thus they can be of any value, depending on the application and nature of study. This absence of limitation on components value selection has obvious advantages to the purchasing authority which cannot be met by fixed components systems, for example:

(1) The cost of the Tutor Mark 2 is initially less.
(2) Cheaper computing components can be used for initiation studies.
(3) Values can be varied to meet the requirements of small system analogue computation.

(4) Replacement costs are held to a minimum.

Six operational amplifiers are provided. Four may be used as 'summers' or 'integrators' and the remaining two as 'summers' only. Provision is also made for the addition of an electronic multiplier as a standard unit. Extras included are such computing refinements as function switches, diodes for discontinuities.

All metering instruments and control are arranged conveniently on the front of the console. These include:

(1) 8-ten turn helical potentiometers of better than 0.1 per cent resolution, $50k\Omega$. Two are earth-free with locking facilities.
(2) A centre zero null balance meter with three sensitivity ranges, $\pm 100V$, $\pm 10V$, $\pm 1V$.
(3) An external meter jack to accommodate use of a digital voltmeter.
(4) A computer mode control switch governing mode operations: reset; pot set; slave; compute; hold and repetitive.

Overload indication is given by a neon warning lamp at the centre of each amplifier symbol on the coloured graphic patch panel which also includes facilities for setting respective amplifier zeros.

To ensure that students arrive at a full understanding of analogue computer technology in the simplest way possible, the complete front panel assembly is made removable to show the construction throughout. The computer has an internal reference supply of $+100V$, 50mA, 0.25 per cent.

Being entirely portable, the unit can be removed by hand between classrooms or laboratories. In the case of larger scale studies, two or more Tutors can be interconnected to provide an increased computing capacity, one Tutor acting as the 'master'.

Improving the Performance of Cold-Cathode Counting Tubes

By F. Szlávik* and L. Garamszegi*

Some methods devised for improving the performance of gas-filled counter tubes at high counting rates are discussed. Circuits are described by which the rate of reliably counted pulses can be increased well over the frequency limit given in the manufacturer's specification. In this way, counting frequencies of 100kc/s and 2Mc/s can be achieved for 40 and 20 electrode tubes, respectively.

(Voir page 867 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 874)

IN recent years a number of papers have dealt with new applications and the development of refined versions of the cold-cathode, gas-filled, pulse counting tubes. This is further evidence of the fact that 'solid state' multistable devices are still far from taking over completely the job of the counting tubes. The specified volume of an up-to-date commercial decade with counting tube is about 3in³, its power consumption is less than 150mW, its cost is relatively low, performance reliability is 0·3 per cent per
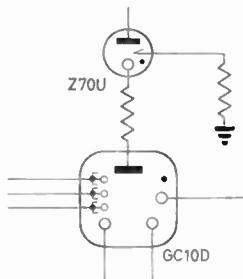
Fig. 1. Anode circuit of the counting tube with gas discharge tube

decade failure and one of its main attractions, in common with any gas-filled tube, is the direct visual indication of the number of counts by the counting element itself.

'Solid state' components are, of course, preferable in cases, when the storage and subsequent central automatic print-out of many data or the scaling off, or counting of events happening at frequencies, higher than the rated speed of the counting tubes are required. It is precisely that latter inadequacy of the counting tubes which has prompted the present efforts to improve the performance of most types of counting tubes by devising circuits for their reliable operation above the rated operation frequency.

Main Limitations

The frequency limit of counting stages, i.e. the maximum rate of periodical single input pulses which is resolved by the counting stage without appreciable error, is determined by the combination of several factors. The near optimum shape and duration of the stepping pulses is usually not difficult to provide for. Some special requirements in driving circuits will be discussed later. First, the limitations due to gas-filling, mechanical construction and d.c. setting will be considered.

Gas composition and pressure dependent recombination time usually affects the maximum counting rate of gas-filled tubes. Nevertheless, in present investigations it was found that, except for neon-filled types, this limitation is seldom close to the rated maximum frequency.

* *Central Research Institute for Physics, Budapest.*

Here it should be pointed out that the double pulse resolution time τ_d involving much lower repetition frequencies, has usually been found higher than the inverse of the rated frequency value given either by the manu-

Fig. 2. Conventional driving circuit for a 40 electrode tube
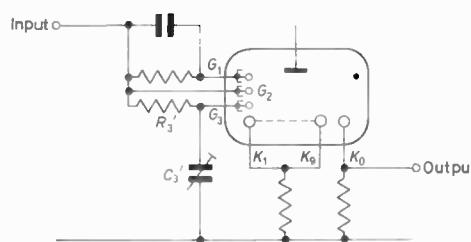
Fig. 3. Modified driving circuit for a 40 electrode tube

facturers or in the literature. In view of the more and more extensive nuclear techniques covering pulses distributed at random, it is thought advisable to specify for all counting circuits the rated resolving time in addition to the rated frequency.

The anode circuit of a counting tube is worth careful consideration. Owing to electrode and other stray capacitances the parallel combination of the anode resistor and the tube resistance is shunted usually by about 10pF. The resulting time constant causes the anode to follow but with some delay or, for stepping frequencies above a given value, only inadequately, the voltage variations of the driving electrodes. This may lead to failure in operation. The capacitance can be diminished only slightly even by the most careful mounting, while circuits designed with low anode load resistance in order to increase the frequency limit, are found to be unreliable. This is obvious, since it is the high anode load resistance which ensures that every time only one of the cathodes and guides should take part in sustaining the gas discharge.

Fig. 4(a). Waveform diagrams of the conventional driving circuit
(*See Fig. 2 for a 40 electrode tube*)
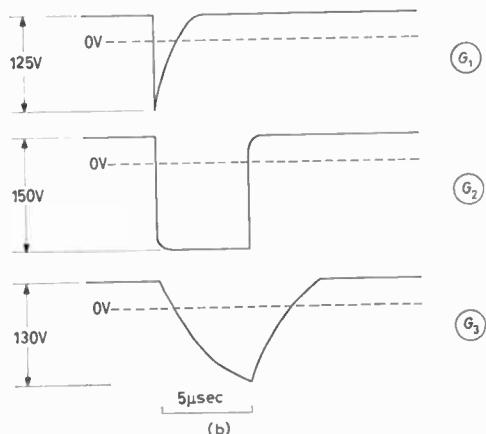
(b). Waveform diagrams of the fully modified circuit
(*See Fig. 5 for a 40 electrode tube*)
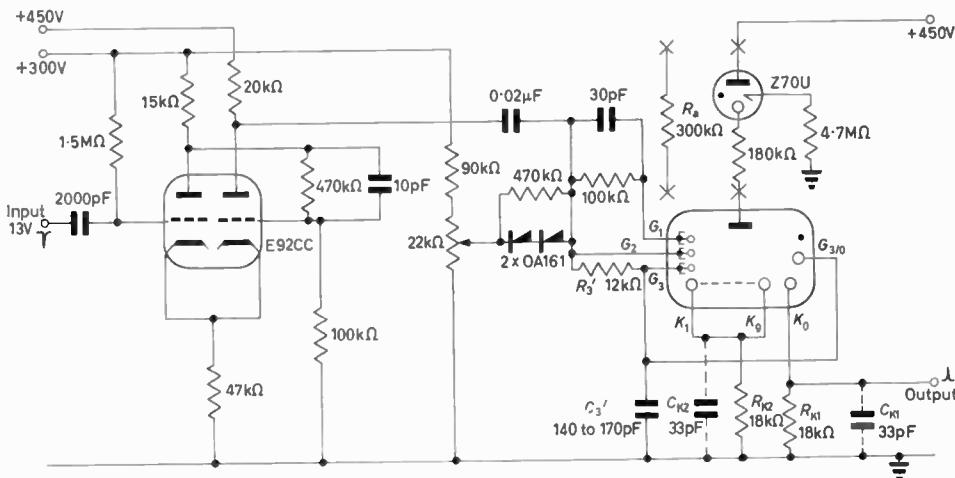
Fig. 5. Fully modified circuit for a 40 electrode tube

Circuit Developments

FORTY ELECTRODE TUBES

The tubes used in the experiments were ten of the type GC10D.

As a first step, some alterations were attempted on the anode circuit, since, as it follows from the above considerations that a high d.c. but low a.c. load resistance is likely to produce a beneficial effect. Under the given voltage and current conditions inserting a gas discharge tube into the anode circuit seemed the most promising (Fig. 1).

A miniaturized type (Z70U—Valvo) was used and its long-term stability was satisfactory. The grid of this cold cathode 'triode' was biased to a fixed value. This alteration alone, however, did not bring about an appreciable improvement, at least with this type of tube, for the circuit of the manufacturer's specification[4]. Thus, the driving circuit of the forty electrode tubes had to be considered (Fig. 2).

The counting tube is stepped as follows:

(1) The leading edge of input u_l triggers the guide G_1 nearest to the cathode K_n involved in the discharge.

(2) The charging of capacitor C_1 raises the potential at resistor R_1 so that the glow is transferred to the negative adjacent guide G_2.

(3) The charging of G_3 transfers the glow to the next adjacent most negative electrode, K_{n+1}, in the manner described above, and hereby the new stable state of the tube sets in.

Processes (3) and (4) take substantially more time than (1) and (2). This is easily understood by considering that the driving effect of pulse u_l having an amplitude of about four times U_b and steep leading and trailing edges, does not assert itself directly at the guides G_3. The circuit R_3 — C_3 thus, 'left alone' is somewhat slower in taking over and transferring the glow. This, however, can be remedied by applying the stepping pulse with a suitable delay to G_3 as well (Fig. 3).

In this way not only the transfer time from G_2 to G_3 has been shortened but also that from G_3 to K, since the trailing edge of the pulse u_l hastens the discharge of capacitor C'_3. The optimum transfer time was found to be less than the width of the input pulse, τ_l. Some numerical data are included in the waveform diagrams (Fig. 4).

The combined application of the two modifications (Fig. 5) resulted in an increase in the upper frequency limit to 110kc/s as compared with 30kc/s obtained with the circuit proposed by the manufacturer (Table 1).

A further improvement results from using additional capacitors (connected with dashed lines in Fig. 5). These have been introduced to stabilize discharge conditions which are less well defined as the stepping frequency increases. The capacitors keep the cathode positive

Table 1. Comparison of data of various circuit for 40 electrode tubes

MEASURED DATA THE CIRCUIT APPLIED	ANODE CURRENT (AVERAGE) (μA)	FREQUENCY LIMIT (MIN.) (KC/S)	RESOLVING TIME (MAX.) (KC/S)	DRIVING SIGNALS ON THE GUIDES	
				AMPLITUDE (V)	WIDTH (μSEC)
1. Circuit recommended by the manufacturer* ($R_{k_2} = 0, C_{k_1} = 0$ and $C_{k_2} = 0$†)	750	33	40	175	5
2. As in (1.), but $R_{k_2} = 18k\Omega$†	750	40	30	175	5
3. As in (2.), but $C_{k_1} = C_{k_2} = 33pF$†	750	45	24	175	5
4. Fully modified circuit (See Fig. 5)	800	115	10·7	150	3
5. As in (4.), but $C_{k_1} = C_{k_2} = 0$	800	108	12·5	150	3
6. As in (4.), but in the anode circuit only: $R_a = 300k\Omega$	800	112	12·8	150	3
7. As in (6.), but $C_{k_1} = C_{k_2} = 0$	800	105	13	150	3

* See reference 4 or Fig. 2.
† See Fig. 5.

past the rise time of u_i, increasing the effective magnitude of the latter, thus compensating for the finite recombination time of the ions close to the cathode. Furthermore they reduce the negative feedback from the resistors R_k in the cathode circuit during the transfer from G_3 to K where the transferred glow has to find a rest.

TWENTY ELECTRODE TUBES

EZ 10-B (Elesta) type tubes were chosen for the investigation, and for checking the design, 15 tubes were used. The main attraction of this type was that it was the only gas-filled counting tube available with a maximum rated frequency of 1Mc/s. Though this count rate could be attained for uniformly distributed pulses by reproducing the circuit suggested by the manufacturer[3], the double pulse resolution was not better than 1·6μsec.

The anode circuit was first changed as shown in Fig. 1. The result was an improvement of about 30 per cent in the resolving time. The electric inertia of the anode circuit was still the main obstacle to an increase in frequency, so the anode resistor was connected to the power supply across an inductance which delivered, acting as virtual load resistance of the driving tube, the negative driving pulses to the guides (Fig. 6).

The anode of the counting tube is thus 'forced' to follow up the variations in the guide potentials. The positive overshooting at the inductance, as the input pulse decays, results in a further shortening of the cycle.

As known, the hooked pins acting as guides and cathodes in the twenty electrode tubes are so turned that the discharge conditions of any cathode should lead to the pre-ionization of the clockwise adjacent guide. A negative pulse applied to the guides G transfers the glow to the pre-ionized one and as the pulse decays, the glow is transferred by a similar mechanism to the next clockwise adajcent cathode.

The analysis of the tube behaviour close to the 'frequency limit' has shown the spreading out of the discharging ion beam to be most responsible for the operation becoming unstable. The driving circuit therefore has been altered as follows.

The roles of the cathodes and guides have been reversed. This seemed necessary, since in the type EZ10-B tubes the guides are, unfortunately, not connected in accordance with their serial numbers being even or odd, but each set of five guides is set up in a half circle. Now, the cathodes have been connected in sequences of even or odd serial numbers and a bistable multivibrator has been introduced permitting a negative driving pulse to be passed only to the 'guides' K, one of which is clockwise adjacent to the 'cathode' G responsible at this time for the glow.

The circuit diagram devised with the two above alterations is shown in Fig. 7.

The triggering signal for the next decade is produced by a differential transformer. This signal of sufficiently high amplitude to trigger the blocking oscillator V_{3a} appears at the transformer output when the glow from the last electrode of the half circle G_A is transferred to the first electrode of the half circle G_B. The currents then flowing in the primary windings of the transformer mutually increase the effect of each other. The opposite coupling of the primary windings results in an appreciable neutralizing of the interfering negative pulses straying to all electrodes. The neutralizing effect is further increased by diodes MR_1 and MR_2. Waveforms are shown in Fig. 8.

For comparison, the measured values are summarized in Table 2.

Conclusions

The methods described above seem to be useful for investigating in each case whether the frequency limit of the counting tube chosen for application can be increased by relatively simple means to cover a higher range of frequency than that specified by the manufacturer.

Fig. 6. 'Forced' operation of the anode circuit
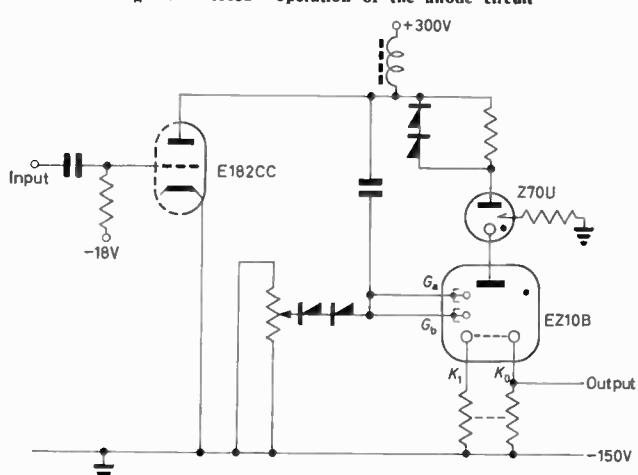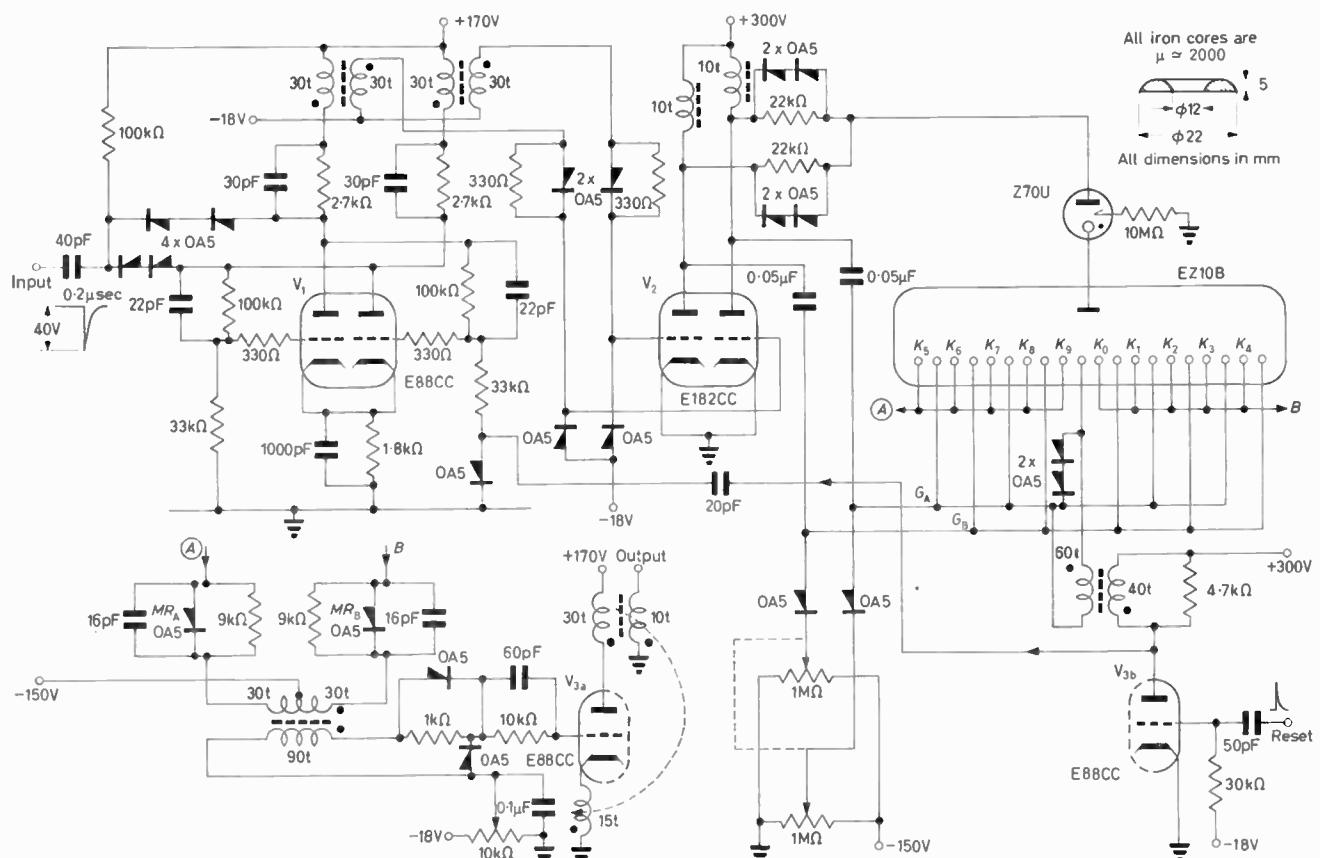
Fig. 7 (above). Fully modified circuit for a 20 electrode tube

Fig. 8. (left). Waveform diagrams of the fully modified circuit
(See Fig. 7 for a 20 electrode tube)

It would be useful to produce on a commercial scale twenty electrode tubes with guides connected in sequences of odd and even serial numbers. This would present but a minimum modification in the production process, yet could yield in return a 0·5μsec resolution time, while each cathode could still be applied as an individual electric lead out, the output circuit and the presetting of the tube would become, on the other hand, much simpler.

Acknowledgment

The authors are indebted to Head Department Leader, Mr. F. Szabó, for his stimulating interest in the present work.

Table 2. Comparison of data of various circuit for 20 electrode tubes

THE CIRCUIT APPLIED	MEASURED DATA				
	ANODE CURRENT (AVERAGE) (μA)	FREQUENCY LIMIT (MIN.) (KC/S)	RESOLVING TIME (MAX.) (KC/S)	DRIVING SIGNALS ON THE GUIDES	
				AMPLITUDE (V)	WIDTH (μSEC)
1. Circuit recommended by the manufacturer*	1500	1100	1·3	200	
2. Fully modified circuit (See Fig. 7)	1200	2000	0·65	75†	0·2†
3. As in (2.), but with unmodified anode circuit	1200	1400	1·1	150†	0·2†
4. As in (2), but without the alternative cathode drive	1200	1200	1·2	175	0·2

* See reference 3.   † Measured on points G_A and G_B (see Fig. 7)

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Harmonic Voltages in the Outputs of Controlled Rectifier Circuits

By R. Smith*, A.M.I.E.E.

In this article the harmonic voltages present in the output of controlled rectifier circuits are derived mathematically. Curves of the equivalent ripple related to 800c/s (weighted ripple) as used for telephone purposes, are also given. A table of ripple weighting factors is included.

(Voir page 867 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 874)

THE output waveform from a controlled rectifier, such as a thyristor, thyratron, or grid-controlled mercury-arc rectifier, is considerably distorted, and generally has to be smoothed before it can be used. To design the most economical smoothing filter, accurate information on the amplitude of the harmonic voltages in the output of the rectifier is required.

The harmonic voltages can best be derived mathematically, and this has been done for the rectifier connexions likely to use controlled rectifiers. In addition to this information, curves of the equivalent ripple related to 800c/s (weighted ripple) as used for telephone purposes, are also given.

In all the calculations, the effect of angle of overlap caused by supply reactance was neglected.

All the above types of rectifier give output waveforms which can be precisely forecast. A combination of transductor and rectifier gives similar waveforms, but, due to the essentially inductive nature of a transductor, the waveform has distortions introduced which cannot be so accurately predicted. Hence the curves given here are correct for semiconductor or electronic rectifiers, but are only approximate for transductor regulated rectifiers.

Unsmoothed Ripple Output

The unsmoothed ripple output from a controlled rectifier is generally larger than that from an uncontrolled rectifier and its value varies with the firing angle. There are two possible ways of expressing the ripple voltage: it can be referred to either the output voltage when α is zero (V_{DO}), or to the output voltage at a firing angle (V_{DO_α}). This article uses the former datum for two reasons: (a) If V_{DO_α} is used as a reference the ripple tends to an infinite value when V_{DO_α} approaches zero. (b) V_{DO} is always known or can be easily calculated, and is constant with any given set of components.

Two sets of ripple curves are given:

(a) Normal ripple.

(b) Ripple weighted to 800c/s.

The derivation of these curves is given in Appendices (1) and (2).

Normal Ripple Curves

These curves are derived from a Fourier analysis of the various waveforms, the actual computation being carried out by a 'ZEBRA' computer at Standard Telecomunication Laboratories. Details of the analysis and the formulae used in the computation are given in Appendix (1). To obtain the total ripple figures all harmonics up to the 24th were summed. Each graph gives the total ripple in the circuit considered and the first four significant harmonics.

Drawing higher harmonics on graphs of this size gives only a confusing picture.

Curves are given for:

Fig. 1. Single-phase full-wave connexion, with commutating (or free-wheeling) diode.

Fig. 2. Single-phase full-wave connexion, without commutating diode, on infinitely inductive load.

Fig. 3. Three-phase half-wave connexion, with commutating diode.

Fig. 4. Three-phase half-wave connexion, without commutating diode, on infinitely inductive load.

Fig. 5. Three-phase bridge connexion, using three controlled arms, with or without commutating diode.

Fig. 6. Three-phase connexion, using six controlled arms, with commutating diode.

Fig. 7. Three-phase bridge connexion, using six controlled arms, without commutating diode, on infinitely inductive load.

Figs. 6 and 7 also give the ripple and harmonics for


SYMBOLS

a_n	= Coefficient in Fourier series = twice the mean value of $f(\theta)\cos(n\theta)$ in the range $(-\pi, \pi)$
b_n	= Coefficient in Fourier series = twice the mean value of $f(\theta)\sin(n\theta)$ in the range $(-\pi, \pi)$
$E_{(pk)}$	= Peak voltage
E_n	= R.M.S. voltage of nth harmonic.
$E_{n(pk)}$	= Peak voltage of nth harmonic.
E_T	= Total r.m.s. ripple voltage
e	= Instantaneous voltage
m	= Number of pulses in the output of an uncontrolled rectifier in one cycle of the supply frequency (two for single-phase full-wave, three for three-phase half-wave, six for three-phase full-wave)
n	= Harmonic number, referred to supply frequency
V_{DO}	= Theoretical no load d.c. voltage $\alpha = 0$
V_{DO_α}	= Theoretical no load d.c. voltage at firing angle α
α	= Firing angle in degrees or radians—the delay in firing a controlled rectifier from the point of intersection of the commutating voltage waveforms
β	= Angle at which commutation occurs at $\alpha = 0$
θ	= Angle in degrees or radians
λ_n	= Weighting factor for n^{th} harmonic

Standard Telephones & Cables Ltd.
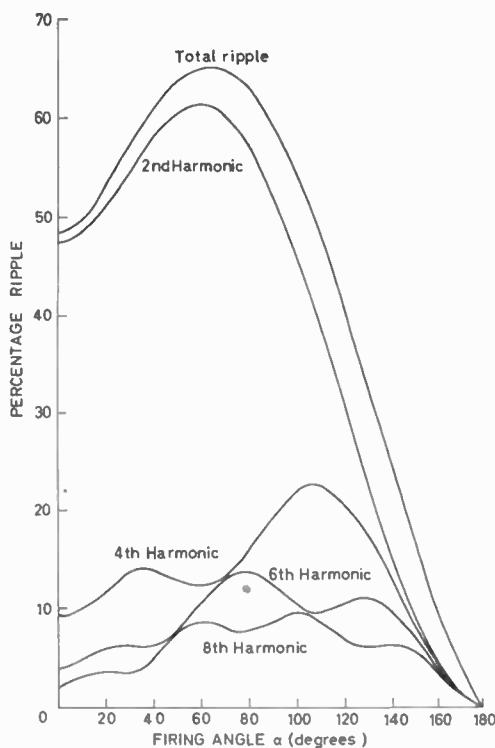
**Fig. 1. Single-phase full-wave ripple (with commutating diode)**

(1) *All ripple values are referred to V_{DO}, the output voltage when $\alpha = 0$*

(2) *Total ripple includes all harmonics up to the 24th*

(3) *These curves apply (a) to all single-phase full-wave rectifier connexions, with commutating diode, on resistive or inductive loads, (b) to all single-phase full-wave rectifier connexions, no commutating diode, on pure resistive loads, (c) to single-phase bridge connexions with only two controlled arms, no commutating diode on resistive or inductive loads*
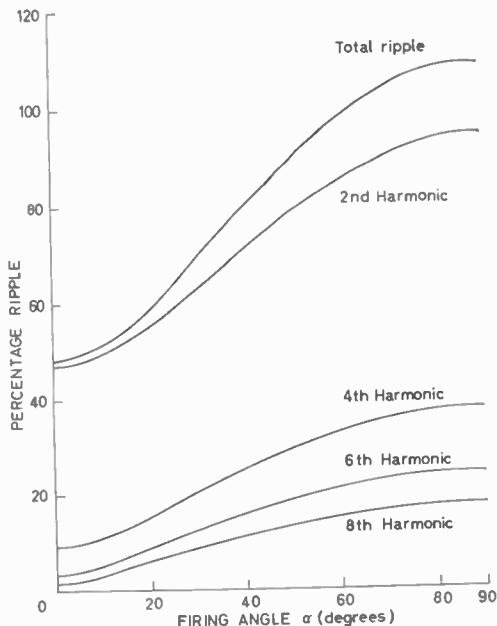
Fig. 2. Single-phase full-wave ripple (no commutating diode)

(1), (2) *As Fig. 1*

(3) *These curves apply only to (a) single-phase centre-tap controlled rectifier connexions, no commutating diode, on infinitely inductive loads, (b) single-phase bridge connexions with four controlled arms, no commutating diode, on infinitely inductive loads*

double-star with interphase transformer connexions. For these connexions α is measured from the output voltage commutation point, not from the commutation point of the individual three-phase half-wave rectifiers.
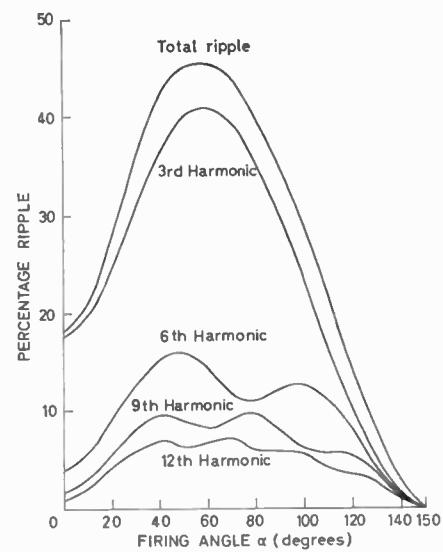
Fig. 3. Three-phase half-wave ripple (with commutating diode)

(1), (2) *As Fig. 1*

(3) *These curves apply to three-phase half-wave rectifier connexions, with commutating diode on resistive or inductive loads and to three-phase half-wave rectifier connexions, no commutating diode, on pure resistive loads*
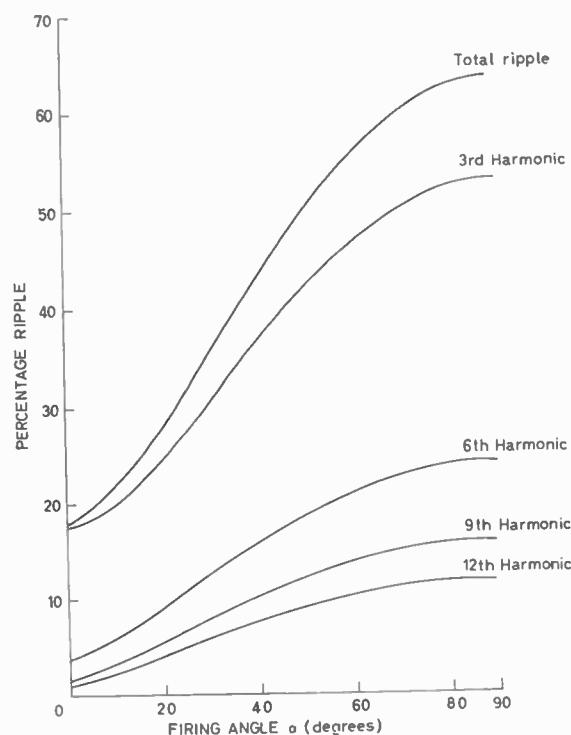
Fig. 4. Three-phase half-wave ripple (no commutating diode)

(1), (2) *As Fig. 1*

(3) *These curves apply only to three-phase half-wave controlled rectifier connexions, no commutating diode, in infinitely inductive loads*

Output waveforms and characteristics for the various circuits and the effect of the commutating diode have been published[2].

Weighted Ripple Curves

The combination of a telephone line, telephone receiver, and a human ear gives some frequencies greater prominence than others. Because of this, noise on the line will be more objectionable at certain frequencies. It
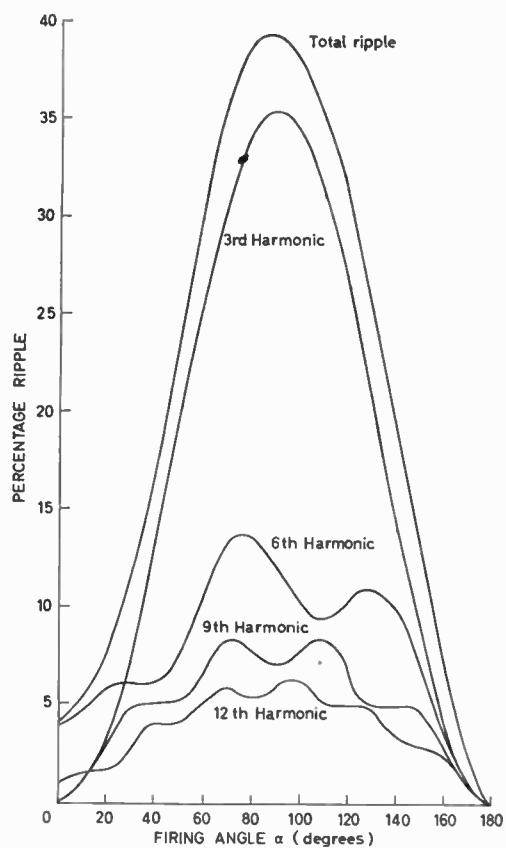
Fig. 5. Ripple from three-phase bridge with three controlled arms

(1), (2) As Fig. 1
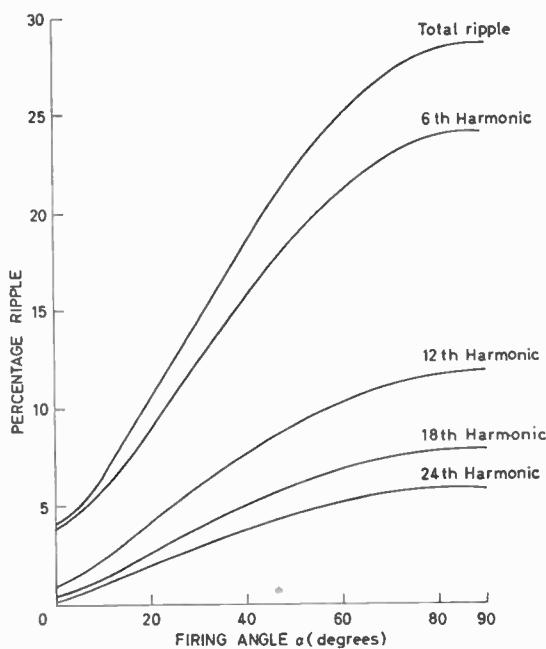
(3) These curves apply to three-phase bridge rectifiers, with three controlled arms with or without commutating diode, on resistive or inductive loads


Fig. 6. Three-phase bridge ripple (six controlled arms, with commutating diode)

(1), (2) As Fig. 1

(3) These curves apply to three-phase bridge rectifiers with six controlled arms and commutating diode, on resistive or inductive loads, and to three-phase bridge rectifiers with six controlled arms, no commutating diode, on pure resistive loads, and to double-star with I.P.T. with commutating diode on resistive or inductive loads, or without commutating diode on infinitely inductive loads

Fig. 7. Three-phase bridge ripple (six controlled arms, no commutating diode)

(1), (2) As Fig. 1

(3) These curves apply only to three-phase bridge rectifiers with six controlled arms, no commutating diode, on infinitely inductive loads, and to double-star with I.P.T., no commutating diode on infinitely inductive loads
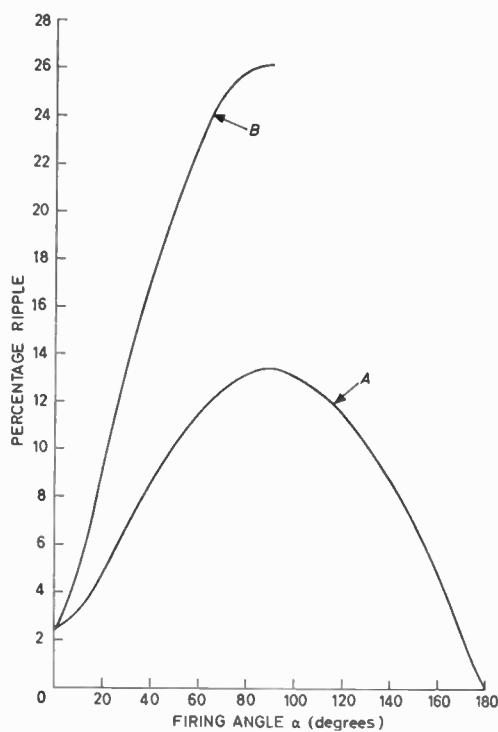
Fig. 8. Single-phase full-wave weighted ripple

(1) As Fig. 1

(2) Curves are weighted with respect to 800c/s, using weighting factors published by the C.C.I.F. and C.C.I.T. in 1952, see appendix II. These factors will be incorporated in the next edition of BS 1698

(3) The weighted ripple shown includes all harmonics up to the 24th

(4) Curve A applies to :

   (a) All single-phase full-wave rectifier connexions, with commutating diode, on resistive or inductive loads

   (b) all single-phase full-wave rectifier connexion, no commutating diode, on pure resistive loads

   (c) single-phase bridge connexions with only two controlled arms, no commutating diode, on resistive or inductive loads

   Curve B applies to :

   (a) Single-phase centre-tap rectifier connexions, no commutating diode, on infinitely inductive loads

   (b) single-phase bridge connexions with four controlled arms, no commutating diode, on infinitely inductive loads
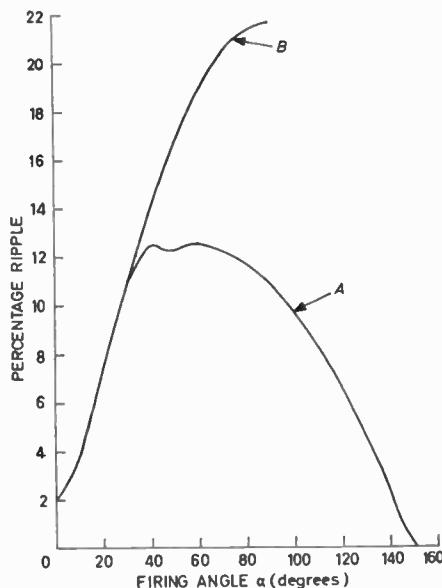
Fig. 9. Three-phase half-wave weighted ripple

(1), (2), (3) As Fig. 8

(4) Curve A applies to :

(a) Three-phase half-wave rectifier connexions, with commutating diode, on resistive or inductive loads

(b) three-phase half-wave rectifier connexions, no commutating diode, on pure resistive loads

Curve B applies to : three-phase half-wave rectifier connexions, no commutating diode, on infinitely inductive loads
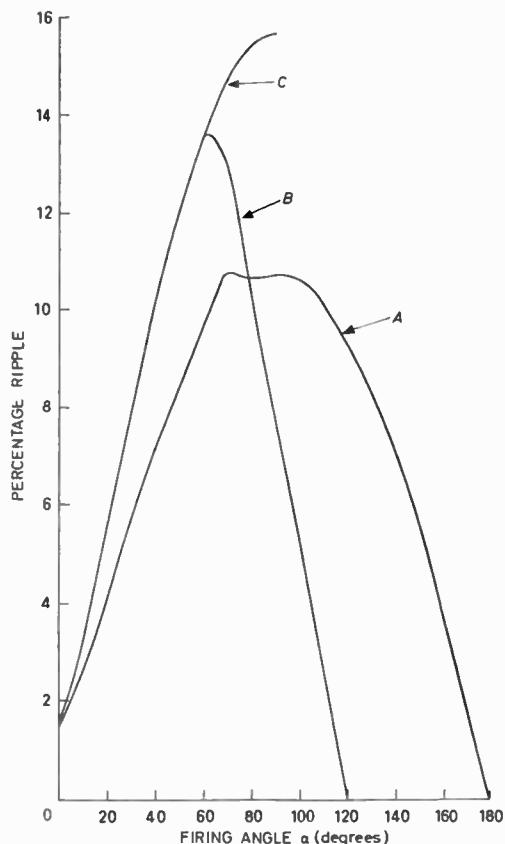
Fig. 10. Three-phase full-wave weighted ripple

(1), (2), (3) As Fig. 8

(4) Curve A applies to : three-phase bridge rectifiers with three controlled arms, with or without commutating diode, on resistive or inductive loads

Curve B applies to :

(a) Three-phase bridge rectifiers with six controlled arms and commutating diode, on resistive or inductive loads

(b) three-phase bridge rectifiers with six controlled arms, no commutating diode on pure resistive loads

(c) double-star with I.P.T. with commutating diode on resistive or inductive loads

(d) double-star with I.P.T. no commutating diode, on pure resistive loads

Curve C applies to :

(a) Three-phase bridge rectifiers with six controlled arms, no commutating diode, on infinitely inductive loads

(b) double-star with I.P.T. no commutating diode, on infinitely inductive loads

has been found experimentally, and agreed by an international convention[3], that noise is most objectionable at frequencies between 800 and 1 200c/s. Reference 3 includes a table of factors by which the various ripple harmonics occurring in the output of rectifiers should be multiplied to give an equivalent noise to that obtained with a pure 800c/s tone. This table is given in Appendix (2), unfortunately it does not agree with the curve given in Appendix E of BS 1698 : 1950. However, the British Standards Institution state that the next edition of BS 1698 will include the C.C.I.F. figures in the form of a table. A table will be used because the B.S.I. consider that this will allow more accurate values to be obtained than is possible by the use of a graph.

The weighted ripple curves, Figs. 8 to 10, given in this article have been calculated using the C.C.I.F. figures, for a supply frequency of 50c/s.

In practice the equivalent weighted ripple is measured by a ' psophometer ', a valve-voltmeter having a frequency response similar to that of an average ear using an average telephone. Because of this, weighted ripple values are often

TABLE 1

FREQUENCY (c/s)	WEIGHTING	
	RELATIVE VALUES	CORRECTING FACTOR (dB)
16·7	0·000056	−85
50	0·00071	−63
100	0·00891	−41
150	0·0355	−29
200	0·089	−21
250	0·178	−15
300	0·295	−10·6
400	0·484	− 6·3
500	0·661	− 3·6
600	0·794	− 2·0
700	0·902	− 0·9
800	1·0	0
900	1·072	+ 0·6
1000	1·122	+ 1·0
1050	1·109	+ 0·9
1100	1·072	+ 0·6
1200	1·0	0
1300	0·955	− 0·4
1350	0·928	− 0·65
1400	0·905	− 0·87
1500	0·861	− 1·3
1600	0·824	− 1·68
1650	0·807	− 1·86
1800	0·760	− 2·39
2000	0·708	− 3·0
2100	0·689	− 3·24
2200	0·670	− 3·4
2400	0·634	− 3·96
2600	0·598	− 4·46
2700	0·580	− 4·73
2800	0·562	− 5·01
3000	0·525	− 5·6
3300	0·444	− 7·05
3500	0·376	− 8·5
3600	0·335	− 9·5
4000	0·178	−15·0
5000	0·0159	−36·0

termed 'psophometric' ripple values. The word psophometer is derived from the Greek verb psophein—to make a noise.

Conclusions

Curves of raw ripple voltages referred to the theoretical open-circuit voltage with zero firing delay have been derived for the common controlled rectifier connexions. These curves are repeated after weighting to an equivalent 800c/s, assuming 50c/s supply, using weighting factors published by the C.C.I.F. and C.C.I.T. in 1952. They will allow accurate design of smoothing filters for circuits using semiconductor or electronic controlled rectifiers, or, with less accuracy, transductor regulated rectifiers.

APPENDIX

(a) Infinitely Inductive Loads

Ripple values for rectifier circuits not using commutating diodes and supplying infinitely inductive loads are published in various places, e.g. Rissik[1] pages 314 to 317.

The formula given in the above reference and used for the infinitely inductive load curves is:

$$E_n/V_{DO} = \frac{\cos \alpha}{(n^2 - 1)} \quad [2(1 + n^2 \tan^2 \alpha)]^{\frac{1}{2}} \quad \ldots \ldots (1)$$

This equation is correct for all the rectifier connexions considered, only the harmonics present actually vary, for example all even harmonics are present in the single-phase case, only third, sixth, ninth, etc., harmonics are present in the three-phase half-wave case.

(b) Rectifiers with Commutating Diodes

The effect of a commutating diode is to prevent the instantaneous voltage at the rectifier terminals ever reversing its polarity[2]. The output ripple is the same as for the infinitely inductive load case until a discontinuity in the output voltage waveform occurs, this happens when $\alpha = 0$ for the single-phase case, $\alpha = 30°$ for the three-phase half-wave case and at $\alpha = 60°$ for the three-phase full-wave connexions.

Fig. 11. Before discontinuity   Fig. 12. After discontinuity

Now the instantaneous output voltage,

$$e = V_{DO_\alpha} + \sum_{n=1}^{n=\infty} a_n \cos (n\theta) + \sum_{n=1}^{n=\infty} b_n \sin (n\theta)$$

V_{DO_α} is a constant for any α and has been defined for the various connexions[2].

Now by the Fourier theorem,

$$a_n = m/\pi \int_\alpha^{((2\pi/m)+\alpha)} f(\theta) \cos (n\theta) \, d\theta$$

$$b_n = m/\pi \int_\alpha^{((2\pi/n)+\alpha)} f(\theta) \sin (n\theta) \, d\theta$$

Since the waveform repeats itself m times per cycle of the fundamental, the integrations are only carried out over an angle of $2\pi/m$, and then multiplied by m.

Now, for the ripple components, only the magnitude of the harmonic terms are required.

The nth harmonic is:

$$a_n \cos (n\theta) + b_n \sin (n\theta)$$

and the magnitude of the nth harmonic is:

$$E_{n(pk)} = a_n + jb_n$$

$$= m/\pi \int_\alpha^{((2\pi/m)+\alpha)} f(\theta) \cos (n\theta) \, d\theta + (jm/\pi) \int_\alpha^{((2\pi/m)+\alpha)} f(\theta) \sin (n\theta) \, d\theta$$

$$= m/\pi \int_\alpha^{((2\pi/m)+\alpha)} f(\theta) \exp (jn\theta) \, d\theta \text{ where } f(\theta) = E \sin (\theta + \beta)$$

Integrating and simplifying as much as possible.

$$E_{n(pk)} =$$
$$\frac{2V_{DO_\alpha}\{[\cos(n\pi)\cos(n\beta)+\cos(n\alpha)\cos(\alpha+\beta)+n\sin(n\alpha)\sin(\alpha+\beta)]^2 + [\sin(n\alpha)\cos(\alpha+\beta)-n\cos(n\alpha)\sin(\alpha+\beta)-\cos(n\pi)\sin(n\beta)]^2\}^{\frac{1}{2}}}{(n^2-1)[1+\cos(\alpha+\beta)]}$$

$$\ldots \ldots \ldots (2)$$

Hence the root mean square value of the nth harmonic is:

$$E_n = E_{n(pk)}/\sqrt{2} =$$
$$\frac{V_{DO_\alpha} \sqrt{2}\{[\cos(n\pi)\cos(n\beta)+\cos(n\alpha)\cos(\alpha+\beta)+n\sin(n\alpha)\sin(\alpha+\beta)]^2 + [\sin(n\alpha)\cos(\alpha+\beta)-n\cos(n\alpha)\sin(\alpha+\beta)-\cos(n\pi)\sin(n\beta)]^2\}^{\frac{1}{2}}}{(n^2-1)[1+\cos(\alpha+\beta)]}$$

Taking a general case, before discontinuity the d.c. output voltage shape is as shown in Fig. 11 where the full lines show the output ripple. The harmonics in this waveform are calculated as described in section (a) of this appendix.

After discontinuity the d.c. output voltage is as shown in Fig. 12.

This condition can be defined as when
$$((2\pi/m) + \alpha) > (\pi - \beta)$$

For a single-phase full-wave connexion α varies between 0 and π

$$n = 2, 4, 6, 8, \text{ etc.}$$

$$\beta = 0$$

$$V_{DO_\alpha} = V_{DO} \left(\frac{1 + \cos \alpha}{2} \right)$$

Hence

$$E_n/V_{DO} = \frac{\{(1+\cos(n\alpha)\cos\alpha+n\sin(n\alpha)\sin\alpha)^2+(\sin(n\alpha)\cos\alpha-n\cos(n\alpha)\sin\alpha)^2\}^{\frac{1}{2}}}{\sqrt{2(n^2-1)}} \quad \ldots\ldots (3)$$

For three-phase half-wave connexion, α between 0 and $\pi/6$

$$E_n/V_{DO} = \frac{\cos\alpha\{2(1+n^2\tan^2\alpha)\}^{\frac{1}{2}}}{(n^2-1)} \quad \ldots\ldots\ldots (4)$$

and between $\alpha = \pi/6$ and $\alpha = 5\pi/6$

$$E_n/V_{DO} = \frac{\{[\cos(n\pi)+\cos(n\alpha)\cos\alpha+n\sin(n\alpha)\sin\alpha]^2+[\sin(n\alpha)\cos\alpha-n\cos(n\alpha)\sin\alpha]^2\}^{\frac{1}{2}}}{\sqrt{2}(n^2-1)} \quad \ldots\ldots (7)$$

$n = 3, 6, 9, 12,$ etc.

$\beta = \pi/6$

$V_{DO_\alpha} = V_{DO}/\sqrt{3}[\cos(\alpha+\pi/6)+1]$

Hence

$$E_n/V_{DO} = \frac{\sqrt{2}}{\sqrt{3}(n^2-1)}$$

$$\{[\cos(n\pi/6)+\cos(n\alpha)\cos(\alpha+\pi/6)+n\sin(n\alpha)\sin(\alpha+\pi/6)]^2 +$$
$$[\sin(n\alpha)\cos(\alpha+\pi/6)-n\cos(n\alpha)\sin(\alpha+\pi/6)-\cos(n\pi)\sin(n\pi/6)]^2\}^{\frac{1}{2}} \quad \ldots (5)$$

Three-phase bridge with three controlled arms. This is a special case, since, before discontinuity occurs a controlled pulse is followed by an uncontrolled pulse as shown in Fig. 13.

In this case $f(\theta)$ has three values.

When $0 < \theta < \alpha$ $f(\theta) = E_{(pk)}\sin(\theta+2\pi/3)$

$\alpha < \theta < \pi/3$ $f(\theta) = E_{(pk)}\sin(\theta+\pi/3)$

$\pi/3 < \theta < 2\pi/3$ $f(\theta) = E_{(pk)}\sin\theta$
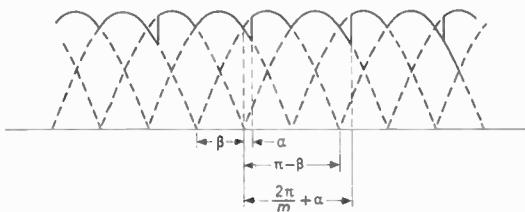
Fig. 13. Controlled pulse followed by uncontrolled pulse

Substituting these values of $f(\theta)$ in the integrands for a_n and b_n and solving for the special case when:

$\beta = \pi/3$

$n = 3, 6, 9, 12,$ etc.

$V_{DO_\alpha} = \frac{1}{2}V_{DO}(1+\cos\alpha)$ gives,

$$E_n/V_{DO} = [1+n^2\sin^2\alpha+\cos^2\alpha+2\cos(n\pi/3)$$
$$(n\sin\alpha\sin(n\alpha)+\cos\alpha\cos(n\alpha))]^{\frac{1}{2}} \quad \ldots\ldots (6)$$

Between $\alpha = \pi/3$ and $\alpha = \pi$, equation (2) is valid and

$\beta = \pi/3$

$n = 3, 6, 9, 12,$ etc.

$V_{DO_\alpha} = \frac{1}{2}V_{DO}(1+\cos\alpha)$

For three-phase bridge with six controlled arms.

Between $\alpha = 0$ and $\pi/3$

$$E_n/V_{DO} = \frac{\cos\alpha[2(1+n^2\tan^2\alpha)]^{\frac{1}{2}}}{(n^2-1)}$$

Between $\alpha = \pi/3$ and $2\pi/3$

$\beta = \pi/3$

$n = 6, 12, 18, 24,$ etc.

$V_{DO_\alpha} = V_{DO}[\cos((\pi/3)+\alpha)+1]$

$$E_n/V_{DO} = \frac{\sqrt{2}}{(n^2-1)}$$

$$\{[1+\cos(n\alpha)\cos(\alpha+\pi/3)+n\sin(n\alpha)\sin(\alpha+\pi/3)]^2 +$$
$$[\sin(n\alpha)\cos(\alpha+\pi/3)-n\cos(n\alpha)\sin(\alpha+\pi/3)^2\}^{\frac{1}{2}} \quad \ldots\ldots\ldots (8)$$

Total Ripple

To produce the total ripple curves in Figs. 1 to 7 the expression:

$$E_{T_\alpha} = E_{1_\alpha}^2 + E_{2_\alpha}^2 + E_{3_\alpha}^2 + \ldots\ldots$$

was used where harmonics up to the 24th were summed in each case. Harmonics above the 24th, in all cases were so small as to have negligible effect on the total.

It was assumed that for all connexions unbalance was negligible and hence $E_{1_\alpha} = 0$.

(2) RIPPLE WEIGHTING FACTORS

Table 1 gives the ripple weighting factors published[3] by the C.C.I.F. and C.C.I.T. in 1952.

The telephone harmonic factor as a percentage is defined as:

$$100/V_{DO}[\lambda_1^2E_1^2+\lambda_2^2E_2^2+\ldots\ldots\lambda_n^2E_n^2]^{\frac{1}{2}}$$

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2. ORR, E. R., SMITH, R., BUDEK, J. A. Thyristors in A.C. Power Circuits. S.T.C. Application Report MF/126X.
3. 'Directives' of the 'Comites Consulatifs Internationale Telephonique et Telegraphique' (C.C.I.F. and C.C.I.T.) (Geneva, 1952).
4. BS 1698:1950 Mercury-Arc Rectifier Equipments.

An Aerial Multi-Coupler

A new high signal level h.f. aerial multi-coupler to enable up to eight communications receivers to be fed simultaneously from a single aerial has been introduced by the Radio Communications Division of G.E.C. (Electronics) Ltd. Designed for use at broadcast-monitoring stations, the aerial multi-coupler covers the frequency range 1 to 30Mc/s.

The main feature of the equipment is that reception of weak wanted signals is not impaired by the presence of high level unwanted signals such as those produced by local transmitters, and in addition the noise factor of 10dB does not degrade the performance of the receiver. If required, aerial multi-couplers can be connected to obtain any multiple of eight outlets.

The equipment comprises three rack mounting units. These can be supplied either complete in a heavy duty cabinet, or as three individual panels to be mounted on standard 19in racks.

The three units are a low-pass input filter with switchable cut-off frequencies, a wideband linear amplifier, and a passive splitter unit.

The low-pass switched input filter allows the operator to confine reception to the minimum required frequency band so reducing the effects of interference caused by high level out-of-band signals.

The splitter unit provides by hybrid transformer techniques eight outlets at 75Ω. The mechanical construction of the unit is arranged so that four of the outlets can be extended by a cable and used remote from the main equipment.

Nominal input-output impedance of the unit is 75Ω. The amplifier operates from 100 to 120V at 60c/s, or 200 to 250V.

A Frequency Compensated Phase Delay Circuit

By A. De Sa*, Ph.D., and L. Molyneux*, Ph.D.

The circuit produces a phase delay that may be set at any value between 40 and 280°, but which changes little with frequency. In essence it is a frequency compensated time delay circuit. The circuit may also be used for frequency multiplication.

MOST phase delay circuits depend on frequency sensitive elements for their action so that the delay depends on the frequency of operation as well as on the setting of the delay control. This feature creates difficulties wherever stable phase delay is required in conditions of varying frequency.

Time delay circuits may be used in place of phase delay circuits but the phase delay for a particular time delay depends on the frequency of operation and a change in the latter will give an apparent change of phase if the time delay remains constant. The difficulty may be overcome if the delay time (t) is made a function of the periodic time (T). Phase delay (ϕ) may be expressed in terms of t and T in the form; $\phi = t/T \times 360°$, but if $t = KT$ then $\phi = K \times 360°$ and ϕ is independent of frequency. The electronic circuit must therefore establish, under manual control, the relationship between t and T. Varying the control varies K and therefore the phase delay.

reference potential occurs remains unchanged. For example, if the slider of the potentiometer is set in a central position the transition point of the comparator will occur half-way up the sawtooth and will be equivalent to a phase delay of 180°.

Changing the frequency changes the magnitude of the sawtooth at which the transition takes place, but not the proportion of its height. The setting of the potentiometer
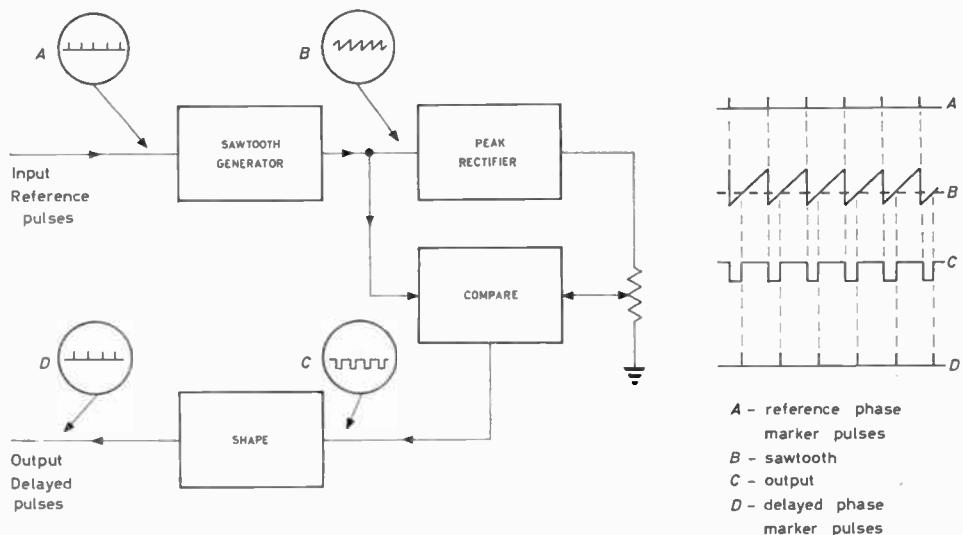
Fig. 1. Method of function of circuit

A – reference phase marker pulses
B – sawtooth
C – output
D – delayed phase marker pulses

Method

Suppose that the phase marker, with respect to which it is wished to define the phase delay, is indicated by a short pulse. This pulse is used to determine the instant of flyback of a sawtooth waveform (Fig. 1, waveform B).

A peak rectifier with a time-constant, long compared with the periodic time of the reference markers, is driven from the sawtooth waveform and a potentiometer connected across its output.

The sawtooth waveform is also connected to one input of a comparator, the other input of which is connected to the slider of the potentiometer. The output of the comparator changes suddenly when its two inputs are equal, otherwise it remains constant at either of two values depending on which of its inputs is the more positive. The transition between the two states marks the moment when the instantaneous value of the sawtooth becomes equal to the potential of the slider of the potentiometer. A change in frequency changes both the magnitude of the sawtooth and the potential across the reference potentiometer, but the proportion of the height at which equality with the

therefore controls the phase delay independently of the frequency of the reference markers. The output from the comparator is differentiated and the unwanted marker during fly-back (which has the opposite polarity to the marker during the linear rise of the sawtooth) is suppressed.

Frequency Multiplication

If the phase delay control is set to its mid-position the delayed markers occur midway between the reference markers, and the waveform of both combined is equivalent to a frequency multiplication of two. Further multiplication can be obtained by adding other comparators set to different phase delays. The circuit can therefore be used for frequency multiplication; moreover since the phase delays do not change with frequency, the circuit once set will provide multiplication of a varying frequency.

A Practical Circuit

If the circuit is to give a phase shift that is perfectly independent of frequency, the following conditions must be fulfilled:

(1) The sawtooth must be linear

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* The University of Newcastle upon Tyne.
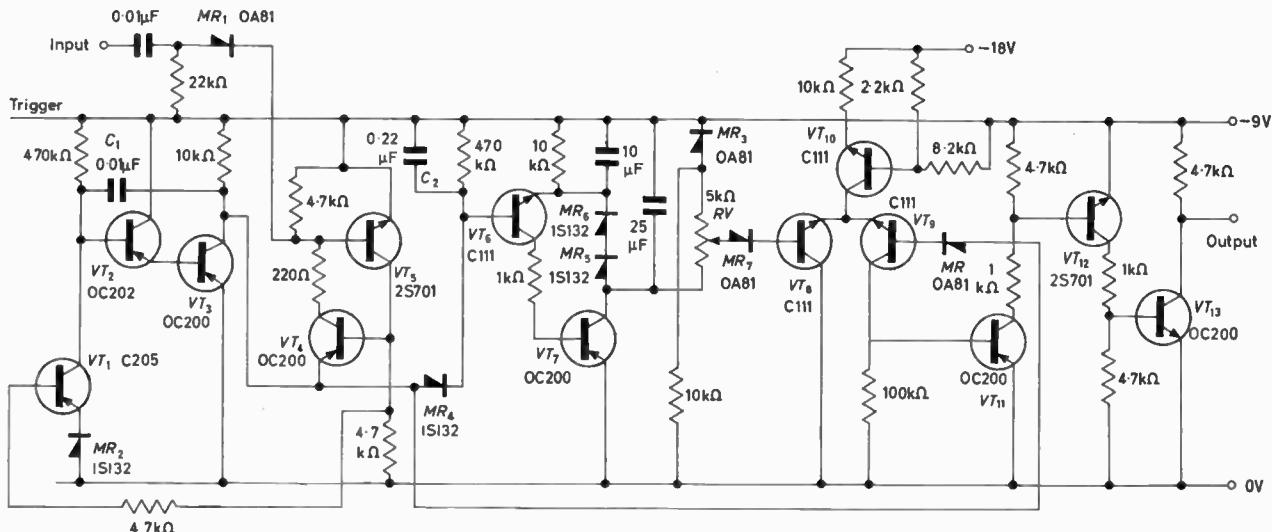
Fig. 2. The phase delay circuit

(2) The flyback must be instantaneous

(3) The peak rectifier must truly represent the magnitude of the sawtooth

(4) The comparator output must depend only on the relationship between its inputs and not on their actual magnitude.

The circuit given in Fig. 2 fulfills these conditions to a moderate degree being mainly deficient in that the peak rectifier does not truly represent the magnitude of the sawtooth and that the flyback is relatively slow (30μsec). However it is adequate for the purpose for which it was designed (phase and mark-space control of an air turbine spinner magnetometer) and serves as a simple example.

VT_2 and VT_3 form a Miller integrator, the sawtooth waveform appearing at the collector of VT_3. The flyback, which is due to the action of VT_4 and VT_5, is initiated by the input reference pulses. VT_1 diverts the changing current of C_1 from the base of VT_2 during the flyback period. VT_1 must have low hole storage if a short flyback time is to be attained. The sawtooth does not start at the negative supply rail, there is a difference of about 0·4V mainly due to the potential difference across VT_4 and VT_5 at the end of flyback. This is balanced by the potential across MR_3 which sets the negative end of the reference potentiometer to a similar value.

The sawtooth waveform is positive going and the positive peak charges C_2 via MR_4. VT_6 and VT_7 form a complementary emitter-follower which includes MR_5 and MR_6 in the feedback loop. The potential difference across these two diodes compensates for the potential difference across MR_4 and base-emitter diode of VT_6. Thus the potential across the reference potentiometer matches the magnitude of the sawtooth waveform.

VT_8 and VT_9 form the comparator, the output of which after amplification by VT_{11} and VT_{12}, is sufficient to drive VT_{13} into saturation. VT_{10} provides a constant current drain for the comparator. Shaping circuits are not shown and the output from the collector of VT_{13} resembles waveform C of Fig. 1.

Performance

The performance of the circuit is shown in Fig. 3. It will be seen that the circuit gives a low change of phase
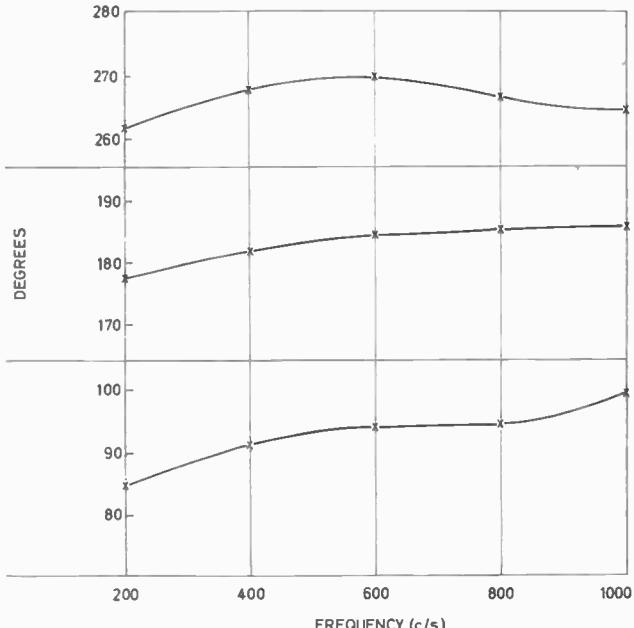
Fig. 3. Variation of three values φ with frequency

over a considerable range of frequency. The circuit will fail to give an output if the phase delay is set too near to 0° delay or too near to 360°. In either case the failure is due to imperfections of the comparator. A practical range for this circuit is 40 to 280° of delay, over a frequency range of 200 to 1 000c/s. The frequency range has a lower boundary which is determined by the maximum height that the sawtooth waveform can reach while remaining linear. With this circuit it is represented by a magnitude almost equal to the supply voltage and occurs at about 150c/s. There is not a well defined upper limit; but, as the frequency is increased the height of the sawtooth diminishes and the deficiencies of the comparator and peak rectifier become significant.

REFERENCE

1. DE SA, A., MOLYNEUX, L. A Transistor Voltage-to-Frequency Corrector. *Electronic Engng. 34*, 468 (1962).

Standing Waves on Transmission Lines

By W. S. Metcalf*, B.Sc.

An analysis of transmission lines is given and from the formulae derived a table is presented which gives the three v.s.w.r. relationships, voltage reflection coefficient, reflected power and transmission loss.

(Voir page 867 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 874)

TRANSMISSION lines are circuit elements with 'distributed constants' spread uniformly along the line. For the purpose of analysis a transmission line may be considered to be made of a large number of identical sections as in Fig. 1.

If a voltage $v\,e^{j\omega t}$ is applied to such a system and a current $i\,e^{j\omega t}$ flows, there will be a voltage drop and a decrease in current given by:

$$\delta v = -(r + j\omega l)\,i\delta x$$
$$\delta i = -(g + j\omega c)\,v\delta x$$

whence

$$d^2v/dx^2 = \gamma^2 v \quad \dots\dots\dots\dots\dots (1)$$

and

$$d^2i/dx^2 = \gamma^2 i \quad \dots\dots\dots\dots\dots (2)$$

where $\gamma = [(r + j\omega l)\,(g + j\omega c)]^{\frac{1}{2}} = \alpha + j\beta$

From equation (1) the voltage at any point along the line is

$$V_x = A\,e^{-\gamma x} + B\,e^{\gamma x} \quad \dots\dots\dots\dots (3)$$

A similar expression may be written for I_x. The relation between voltage and current at any point along an infinite line is constant

$$V_x/I_x = Z_0 = \left(\frac{r + j\omega l}{g + j\omega c}\right)^{\frac{1}{2}}$$

For a high quality line $r \ll j\omega l$ and $g \ll j\omega c$

$$\therefore Z_0 = (l/c)^{\frac{1}{2}} \text{ and } \alpha \text{ tends to } 0.$$

The propagation constant γ now becomes equal to $j\beta$

$$\beta = \omega(lc)^{\frac{1}{2}} = 2\pi/\lambda$$

where λ is the wavelength of the wave in the transmission line. For an open wire or coaxial line

$$\lambda = \frac{\lambda_{\text{free space}}}{(\epsilon_r)^{\frac{1}{2}}}$$

ϵ_r is the dielectric constant of the insulating medium between the transmission line conductors.

Standing Waves

From equation (3), for an ideal line

$$V_x = A\,e^{-j\beta x} + B\,e^{j\beta x}$$

where A and B are arbitrary constants and x is some distance measured from the end of the line.

Anticipating the result, it is known that the two parts of this equation refer to a forward and a reverse travelling wave on the line. If the line is open-circuit, the voltage at the end of the line ($x = 0$), due to the forward travelling wave, is $V_0\,e^{j\omega t}$, i.e. it is equal to the generator voltage. The forward travelling wave may be written

$$(V_x)_t = A\,e^{-j\beta x} = (V_0\,e^{j\omega t})_{x=0}$$

i.e.

$$A = V_0\,e^{j\omega t}.$$

---

* Cavendish Laboratory, Cambridge.

Now, because the line is open-circuited, the reverse wave may also be expressed in a similar fashion.

$$(V_x)_r = B\,e^{j\beta x} = (V_0\,e^{j\omega t})_{x=0}$$

i.e.

$$B = V_0\,e^{j\omega t}.$$

$$\therefore V_x = (V_x)_t + (V_x)_r = V_0\,e^{j(\omega t - \beta x)} + V_0\,e^{j(\omega t + \beta x)}$$

i.e. $V_x = V_0 \sin(\omega t - \beta x) + V_0 \sin(\omega t + \beta x)$

$$= 2V_0 \sin \omega t \,.\, \cos \beta x \quad \dots\dots\dots\dots (4)$$
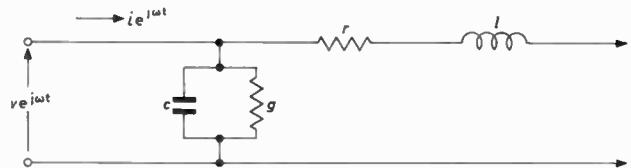
Fig. 1. Lumped section of a transmission line

$l = $ *series inductance per unit length*
$r = $ *series resistance per unit length*
$c = $ *shunt capacitance per unit length*
$g = $ *shunt conductance per unit length*
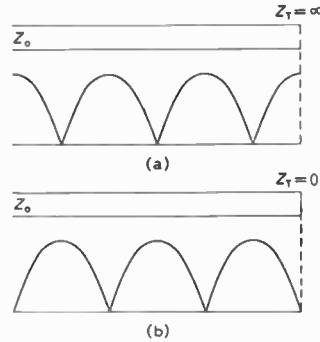
Fig. 2 (left). Voltage standing wave patterns on a transmission line

(a) *Open-circuit termination*
(b) *Short-circuit termination*
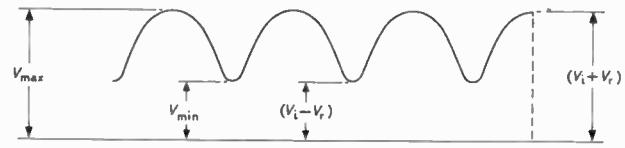
Fig. 3. Voltage standing wave ratio

Equation (4) represents a stationary or 'voltage standing wave' upon the line. This wave is a maximum ($2V_0$) when $\beta x = 0$, π, 2π etc. Substituting $\beta = 2\pi/\lambda$, a maximum occurs when $x = 0$, $\lambda/2$, λ etc. The voltage standing wave is a minimum for $x = \lambda/4$, $3\lambda/4$, $5\lambda/4$ etc.

The reflections from open- and short-circuited terminations are the special cases where the terminating impedance Z_T reaches the extremes of infinity and zero. More generally, perfect reflection does not occur at the end of a line but power is absorbed into the resistive part of Z_T. The impedance along such a line is not constant[1] but depends upon Z_T.

$$Z_x = Z_o \cdot \left(\frac{Z_T + j Z_o \tan \beta x}{Z_o + j Z_T \tan \beta x} \right) \quad \ldots \ldots (5)$$

This has maximum and minimum values as βx tends to zero and infinity depending on the ratio of Z_T/Z_o.

If $Z_T > Z_o$, Z_x is a maximum for $x = 0$, $\lambda/2$, λ etc. and a minimum for $x = \lambda/4$, $3\lambda/4$, $5\lambda/4$ etc.

I $Z_T < Z_o$, Z_x is a maximum for $x = \lambda/4$, $3\lambda/4$, $5\lambda/4$ etc. and a minimum for $x = 0$, $\lambda/2$, λ etc.

The voltage distribution along a line is the same as the impedance distribution for non-reactive loads (reactance at the terminating end leaves the shape of the distribution unchanged but shifts it along the line), i.e. it exhibits a standing wave pattern.
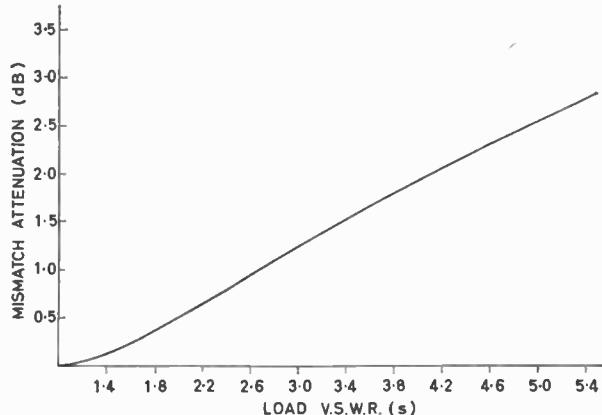
Fig. 4. Attenuation as a function of v.s.w.r.
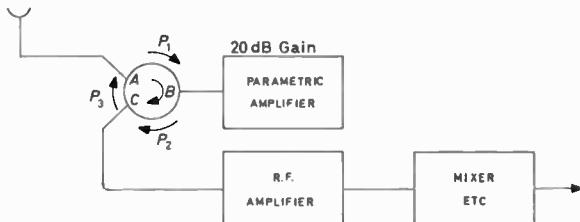
Fig. 5. Parametric amplifier system

Consider a forward travelling wave of peak amplitude V_i on a line and a reverse wave V_r. The forward wave may be thought of as two parts $V_i = (V_r) + (V_i - V_r)$, where $(V_i - V_r)$ represents that part of the wave conveying useful power. The resulting voltage standing wave pattern is shown in Fig. 3. The voltage standing wave ratio is defined as*

$$S = \frac{V_i + V_r}{V_i - V_r} = V_{max}/V_{min} \quad \ldots \ldots \ldots \ldots (6)$$

The v.s.w.r. is sometimes quoted in decibels,

$$(\text{v.s.w.r.})_{(dB)} = 20 \log S$$

The voltage reflection coefficient, ρ, is defined as

$$\rho = V_r/V_i$$

From equation (6)

$$\rho = \frac{S - 1}{S + 1}$$

The percentage of power reflected is

$$(P_r/P_i) \times 100 = (V_r/V_i)^2 \times 100$$

∴ Percentage power reflected = $100 \rho^2$.

---

* S is still occasionally defined as V_{min}/V_{max}, i.e. $0 \leq S \leq 1$; as with most conflicts of definition the American system is usually adopted, i.e. $1 \leq S \leq \infty$.

The ratio of incident power to reflected is $P_i/P_r = 1/\rho^2$ or expressed in decibels

$$(P_i/P_r)_{dB} = 20 \log (1/\rho)$$

The power loss due to reflection, or 'transmission loss' is

$$10 \log \frac{P_i}{P_i - P_r} = 10 \log \frac{1}{1 - \rho^2} \quad dB.$$

Using the above theory Table 1 was computed.

Transmission Line Systems

A mismatched transmission line has two main drawbacks:

(1) There is an attenuation introduced as a result of the v.s.w.r. This is not of great importance at low v.s.w.r.'s as shown in Fig. 4.

Fig. 6. A padding attenuator on a mismatched line
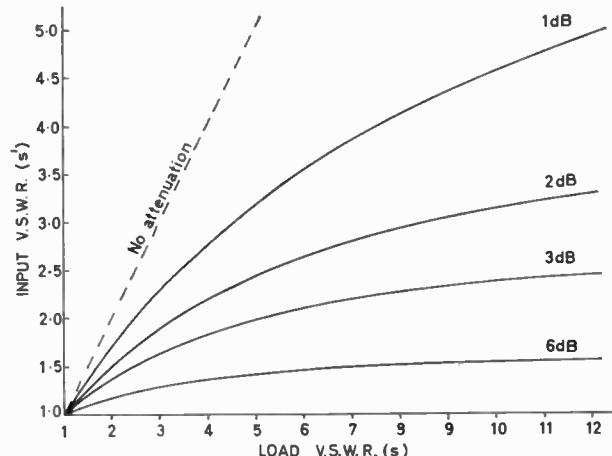
Fig. 7. Reduction of v.s.w.r. as a function of line attenuation

(2) The impedance at any point remote from the end of the transmission line is not constant but is given by equation (5). This can cause great trouble if the device on the remote end of the line is critically affected in its performance by a v.s.w.r.; such a device is a circulator.

A typical low noise parametric amplifier system could consist of an aerial, negative resistance amplifier and r.f. amplifier all connected via a circulator, Fig. 5.

The circulator transmits power in a clockwise direction with low loss (usually about 0·5dB between ports) but gives a high degree of isolation (30 to 40dB) in a reverse direction. Port B of the circulator will not be matched by its very nature and port A is likely to have a v.s.w.r. of 1·4 say; the following power relations may be set down from Table 1 and Fig. 5.

Across port A, $P_1 = 0\cdot0288\,P_3$ (v.s.w.r. = 1·4)

Across Port B, $P_2 = 100\,P_1$ (20dB gain)

∴ $P_2 = 2\cdot88\,P_3$

The condition for oscillation can now be satisfied if

$$P_3/P_2 \geqslant 1/2\cdot88 \text{ across port } C, \text{ or } P_3/P_2 \geqslant 0\cdot35$$

This means that the match or v.s.w.r. at port C must be better than 3·85, over the bandwidth of the parametric amplifier, to prevent oscillation. This does not seem too

V.S.W.R. <1	V.S.W.R. >1	V.S.W.R. (dB)	VOLTAGE REFLECTION COEFFICIENT	POWER REFLECTED (PER CENT)	RATIO OF INCIDENT TO REFLECTED POWER		TRANSMISSION LOSS (dB)
					(RATIO)	(dB)	
·05	20·00	26·02	·90	81·86	1·22	·87	7·41
·10	10·00	20·00	·82	66·94	1·49	1·74	4·81
·15	6·67	16·48	·74	54·63	1·83	2·63	3·43
·20	5·00	13·98	·67	44·44	2·25	3·52	2·55
·25	4·00	12·04	·60	36·00	2·78	4·44	1·94
·30	3·33	10·46	·54	28·99	3·45	5·38	1·49
·35	2·86	9·12	·48	23·18	4·31	6·35	1·15
·40	2·50	7·96	·43	18·37	5·44	7·36	·88
·45	2·22	6·94	·38	14·39	6·95	8·42	·67
·50	2·00	6·02	·33	11·11	9·00	9·54	·51
·52	1·92	5·68	·32	9·97	10·03	10·01	·46
·54	1·85	5·35	·30	8·92	11·21	10·50	·41
·56	1·79	5·04	·28	7·96	12·57	10·99	·36
·58	1·72	4·73	·27	7·07	14·15	11·51	·32
·60	1·67	4·44	·25	6·25	16·00	12·04	·28
·62	1·61	4·15	·23	5·50	18·17	12·59	·25
·64	1·56	3·88	·22	4·82	20·75	13·17	·21
·66	1·52	3·61	·20	4·20	23·84	13·77	·19
·68	1·47	3·35	·19	3·63	27·56	14·40	·16
·70	1·43	3·10	·18	3·11	32·11	15·07	·14
·72	1·39	2·85	·16	2·65	37·73	15·77	·12
·74	1·35	2·62	·15	2·23	44·79	16·51	·10
·76	1·32	2·38	·14	1·86	53·78	17·31	·08
·78	1·28	2·16	·12	1·53	65·46	18·16	·07
·80	1·25	1·94	·11	1·23	81·00	19·08	·05
·81	1·23	1·83	·10	1·10	90·75	19·58	·05
·82	1·22	1·72	·10	·98	102·23	20·10	·04
·83	1·20	1·62	·09	·86	115·88	20·64	·04
·84	1·19	1·51	·09	·76	132·25	21·21	·03
·85	1·18	1·41	·08	·66	152·11	21·82	·03
·86	1·16	1·31	·08	·57	176·51	22·47	·02
·87	1·15	1·21	·07	·48	206·92	23·16	·02
·88	1·14	1·11	·06	·41	245·44	23·90	·02
·89	1·12	1·01	·06	·34	295·21	24·70	·01
·90	1·11	·92	·05	·28	361·00	25·58	·01
·91	1·10	·82	·05	·22	450·38	26·54	·01
·92	1·09	·72	·04	·17	576·00	27·60	·01
·93	1·08	·63	·04	·13	760·18	28·81	·01
·94	1·06	·54	·03	·10	$10·45 \times 10^2$	30·19	·00
·95	1·05	·45	·03	·07	$15·21 \times 10^2$	31·82	·00
·96	1·04	·35	·02	·04	$24·01 \times 10^2$	33·80	·00
·97	1·03	·26	·02	·02	$43·12 \times 10^2$	36·35	·00
·98	1·02	·18	·01	·01	$98·01 \times 10^2$	39·91	·00
·99	1·01	·09	·01	·00	$396·01 \times 10^2$	45·98	·00

extreme until it is remembered that a tuned circuit, matched at its centre frequency, has a v.s.w.r. of 5·8 at the 3dB points.

The restriction of a high v.s.w.r. on a transmission line may be reduced by inserting a matched attenuator between the source and mismatched section, Fig. 6.

The ratio of incident and reflected voltage waves, at the termination, is determined by S

$$V_r/\beta V_i = \frac{S-1}{S+1} \dots \dots \dots (7)$$

Similarly, the incident and reflected waves at the input to the attenuator may be expressed in terms of S'

$$\beta V_r/V_i = \frac{S'-1}{S'+1} \dots \dots \dots (8)$$

Eliminating V_i and V_r from equations (7) and (8)

$$S' = \frac{S(1+\beta^2)+1-\beta^2}{S(1-\beta^2)+1+\beta^2}$$

So, typically if $\beta = \frac{1}{2}$ (6dB)

$$S' = \frac{5S+3}{3S+5}$$

This means that the overall v.s.w.r. of the system will be reduced in the ratio $(5S+3)/(3S+5)$ or, for example, a v.s.w.r. of 2 will be reduced to 1·18. This is a very convenient method of minimizing the effect of a mismatch providing the extra attenuation can be tolerated.

Fig. 7 gives details of the possible reduction in v,s.w.r. by different attenuations. This reduction must also be taken into account when measuring v.s.w.r. at a point remote from the termination as, for example, with a swept frequency feeding into a long cable.

Conclusions

Table 1 provides data on transmission lines collected in a form suitable for general use. The three v.s.w.r. relationships, while simple, are constantly in use by any v.h.f. designer. The table of transmission loss has a self evident purpose, similarly the voltage and power relationships: that listing the ratio of incident to reflected power is required when measuring v.s.w.r. by means of a reflectometer.

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1. JACKSON, WILLIS. High Frequency Transmission Lines, p. 81 (Methuen & Co. Ltd, 1958).

A Simple D.C. Voltmeter

By M. R. Osborne*

An instrument is described which will measure voltage up to 20V d.c. with an accuracy of ±0·2 per cent above ½V, ±1mV below. It is a comparator instrument in which the unknown is equated with a known voltage derived from a precision helical potentiometer and a silicon reference diode. The comparator has an automatic gain controlled characteristic obviating the need for a sensitivity control. This instrument fills the gap between the conventional moving-coil meter and the digital voltmeter in accuracy and cost.

A N instrument for the measurement of voltages up to 20V d.c. with ±0·2 per cent accuracy was required for the setting up of the f.m. recording and reproducing channels of a multi-channel data tape recorder. A moving-coil meter of this accuracy would not be rugged enough, and the use of the correction chart would make frequent use inconvenient. A digital voltmeter would be rugged and accurate enough, but the added convenience of a self-balancing meter hardly justifies its cost and complication in this application. Consequently a comparison method is used in which a voltage equal to the unknown is produced by a calibrated ratio from a reference voltage. The ratio device is a precision helical potentiometer which is operated by hand. The comparator is a mechanical chopper driven at mains frequency feeding a phase sensitive detector and indicator through a transistor a.c. amplifier. The amplifier is capacitively coupled to the chopper to suppress the pulse produced when both chopper contacts are open.

Description

A source of known voltage is applied to the helical potentiometer and a resistor of the same value in series with it; the position of these two in the circuit can be interchanged as shown in Fig. 1, allowing the wiper of the helical potentiometer to go from 0 to 1V or from 1 to 2V, depending on the position of the switch (S_2).

The polarity of the reference source can be reversed with switch S_3. As shown in Fig. 1 the reference is connected to one side of a chopper and the unknown to the other. The armature is connected through an amplifier to the phase sensitive detector. The chopper is operated at 50c/s. The null detector is equipped with a delayed a.g.c. system whereby the response is linear at full gain up to about 1/3 scale deflexion and then becomes logarithmic when the a.g.c. starts to act. The gain of the a.g.c. system can be made sufficient so that the meter does not exceed full scale for inputs of several volts even though near the null the sensitivity is approximately 2mV per division.

The Potentiometer Section

The detailed circuit of the potentiometer section is shown in Fig. 2. Three ranges are provided, 0 to 0·2V, 0 to 2V, 0 to 20V selected by S_1. Each of these is divided to 0 to 1, 1 to 2 by S_2; the polarity is reversed by S_3. For full accuracy over a wide temperature range it is essential that the temperature coefficients of the resistors in the potentiometer section be equal. For this reason the helical potentiometer, of which the temperature coefficient was not known, was shunted by a resistor of lower value to
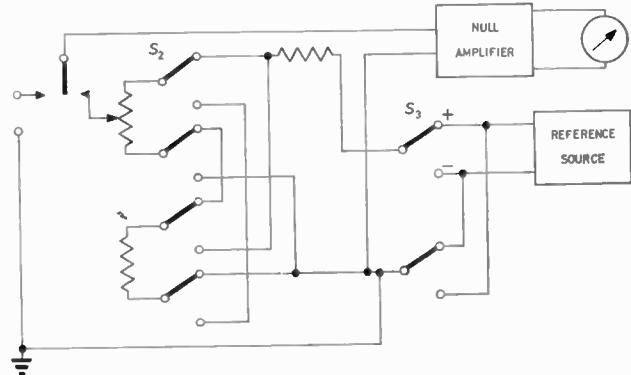
Fig. 1. Simplified circuit arrangement

Fig. 2. Potentiometer section
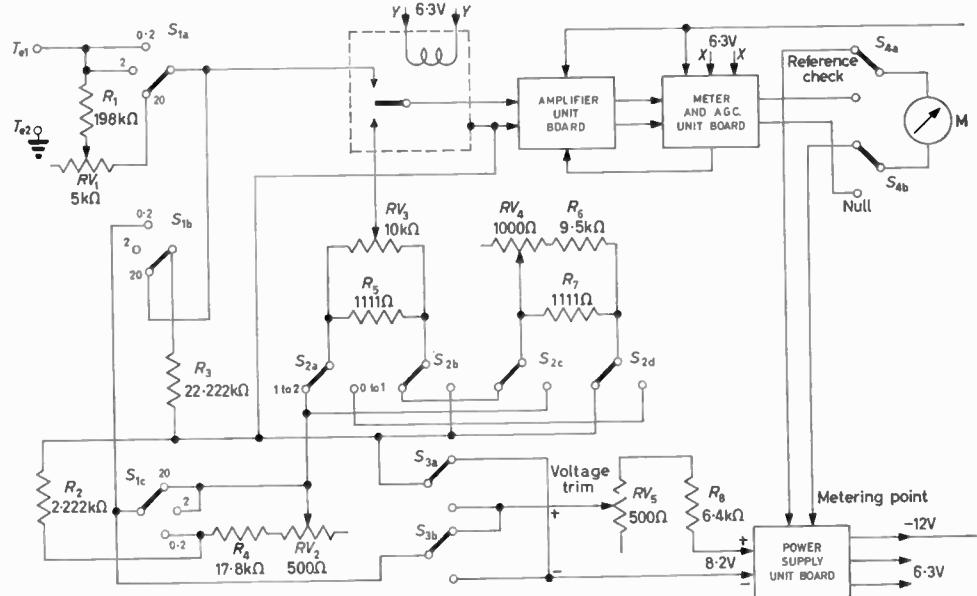
* *Australian Defence Scientific Service.*

swamp unbalanced temperature effects, and precision metal film resistors were used for their low temperature coefficient.

In the 0·2V position the potentiometer is supplied with 0·2V through series resistors R_4 and RV_2. The load on the reference source is maintained constant by inserting R_2 in this position. In the 2V position 2V is supplied to the potentiometer by series resistors R_8 and RV_5. The value of R_8 is selected to suit the voltage of the 1N430 reference diode used, which may range from 8 to 8·8V depending on the particular unit. Final adjustment is accomplished with RV_5. The 20V range is provided by the voltage
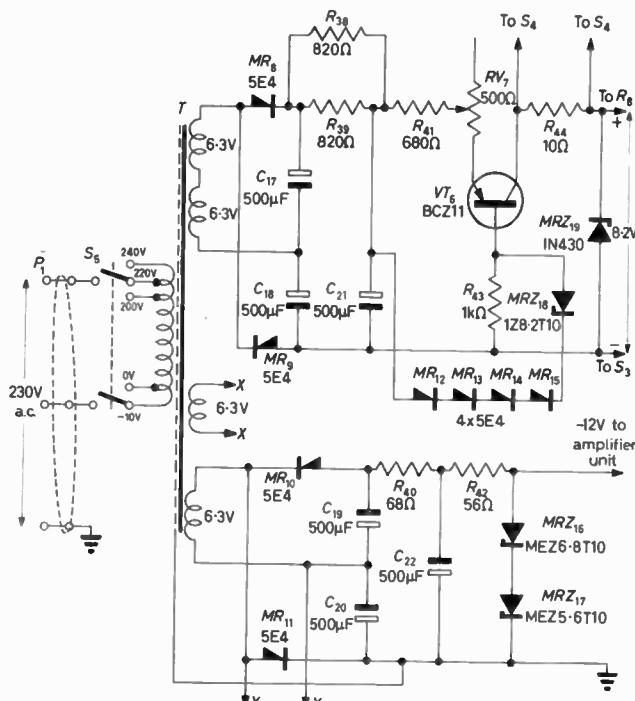
divider R_1. RV_1 and R_3; the potentiometer itself remains unaltered on this range. All critical components are equipped with trimming resistors which give a range of a few per cent for initial setting up of the instrument.

The Reference Source

The reference source (Fig. 3) is based on the International Rectifier type 1N430 reference diode. These units have a temperature coefficient of ±·002 per cent/°C over the range −55° to +100°C. To achieve this performance they must be provided with a current of 10mA ±2·5 per cent. The reference diode is therefore fed by a current regulator whose output can be adjusted to set the diode current to 10mA. The reference element in this regulator is an 8·2V breakdown diode, temperature compensated by four forward biased junctions in series. This simple regulator maintains the required output current over a mains voltage range of 200 to 260V and temperatures in excess of 50°C.

The Null Detector

The null amplifier and detector (Fig. 4) consists of a conventional transistor feedback amplifier feeding a phase sensitive detector of the diode ring modulator type. The gain controlling element is a junction diode, the current through which is varied to control its dynamic resistance. All of the gain controlling function is performed at the input of the amplifier to prevent overload of the amplifier itself.

High input impedance is achieved by the use of feedback around the two input stages and Bootstrapped bias. The voltage gain of this pair is 100. The third stage is a series feedback stage giving a g_m of 2·5mA/V. The a.g.c. amplifier is a shunt feedback stage which acts as a voltage source for the full wave a.g.c. rectifier. The output of the rectifier feeds the base of the control transistor which is a silicon type for low leakage. The threshold and gain of the a.g.c. system are determined by the emitter and base shunting resistors of this stage. Decoupling of the input stages is done with breakdown diodes to avoid the use of large electrolytic capacitors.

The reference source for the phase sensitive rectifier is

Fig. 3. Power supply

Fig. 4. Null amplifier
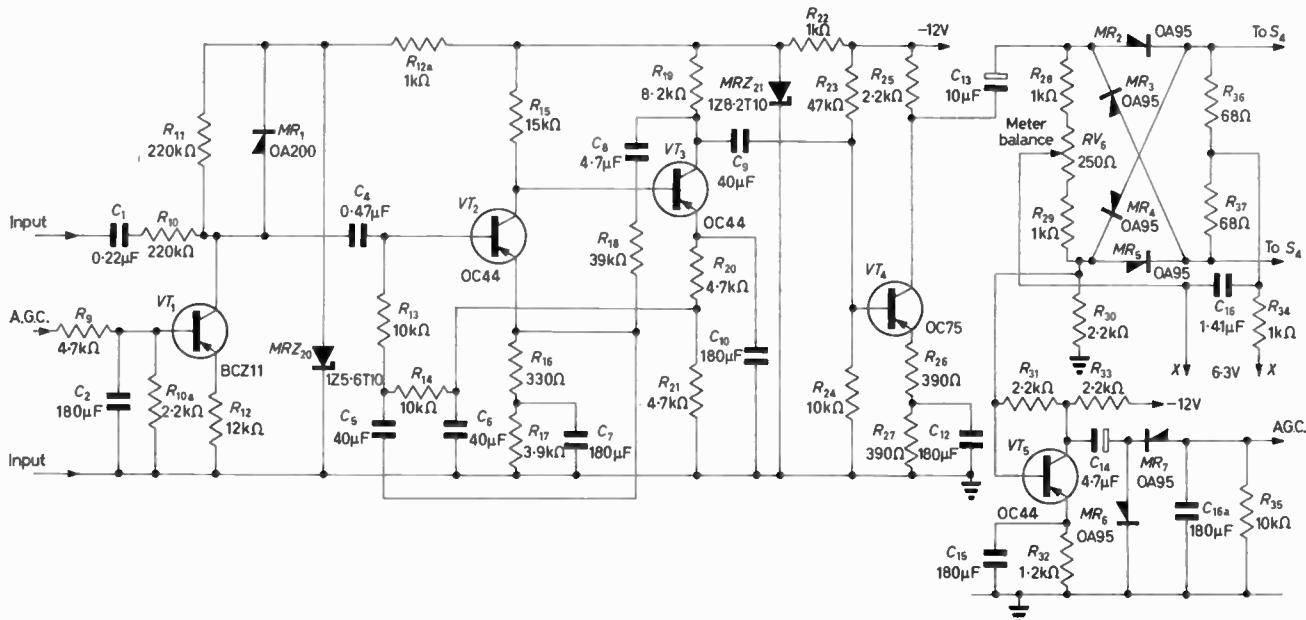
taken from a winding on the power transformer through a phase shift network. The phase shift network compensates for the difference in phase between the chopper excitation and the output, and phase shift in the amplifier. The output of the phase sensitive rectifier is fed to a centre zero $\pm100\mu$A meter.

Conclusion

This instrument has been in use for about a year including 4 000 miles of travel and two months in the tropics. No adjustments have been made and the instrument has been found to be still well within its ±0.2 per cent specification. The unit has found many uses outside the particular one for which it was designed, such as the calibra-

tion of moving-coil meters, and other jobs requiring an accuracy of an order better than the usual multi-meter. The accuracy of the instrument is largely limited by non-linearity of the helical potentiometer, though the temperature coefficient of the reference diode and the precision resistors do play a significant part. Below about 0·5V the accuracy is limited by the minimum readable unbalance, which is less than 1mV.

A New Storm Warning Radar

High power, long range, ease of installation and relatively low cost are four of the main features of 'Rainbow'—a new meteorological radar system developed by The Marconi Co. Ltd. This accurate and extremely reliable new system can track and pin point storms and rain producing clouds within an area of 125 000 square miles.

It has been designed to provide an inexpensive equipment suitable for use at airports, where it can provide first hand meteorological information to the air traffic control centre. Because of its simplicity and ease of installation, 'Rainbow' is also suited to meteorological surveillance in large areas where a number of equipments may be required.

The 70kW p.e.p. high power transmitter and the receiver are both housed in a 12ft high conical shaped aluminium tower on top of which is mounted a 6ft diameter parabolic dish aerial. This complete assembly is easily transported and installed, requiring only a simple 'mains' power supply. The display is connected by a single multi-core cable and may be sited up to 3 000ft away from the tower.

The complete tower installation can withstand wind speeds of up to 120 knots and the equipment will operate normally up to 80 knots. Special filters on all the air vents make the equipment panclimatic and proof against sand, dust and moisture.

The aerial has been designed to produce a narrow pencil beam enabling precision pin-pointing of storm centres both in azimuth and elevation and providing good definition on the display. All the drive and tilt mechanism is installed behind the aerial which rotates continuously in azimuth and may be

The aerial unit of the storm warning radar
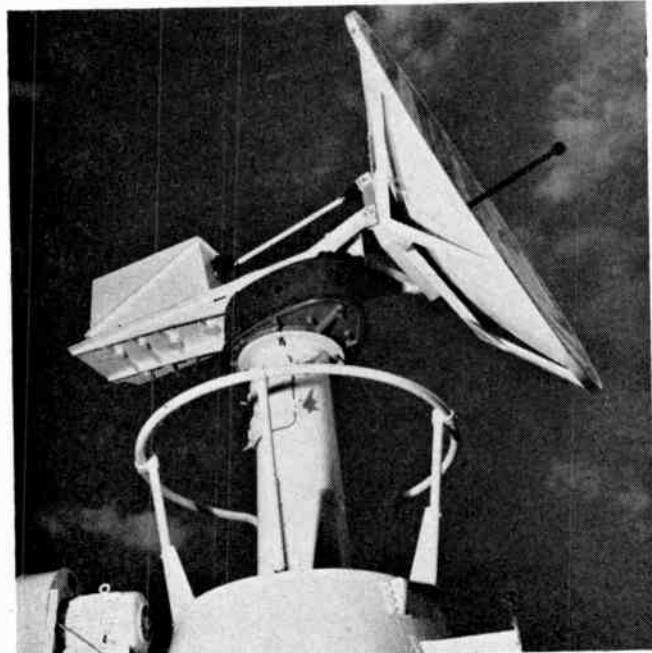
tilted in the vertical plane automatically or manually from the display position.

The display uses a 12in diameter high brightness tube that may be viewed in high ambient light conditions. Six different display ranges are available, from 12½ to 200 nautical miles. Fixed range rings are provided and a special feature is a variable range marker adjustable from 5 to 100 per cent of the displayed range. An electronic bearing indicator enables accurate, parallax free, bearings to be taken, although a mechanical cursor is also provided.

The equipment works in X-Band (9415 to 9475Mc/s) with a pulse length of 2μsec and a p.r.f. of approximately 360p/s. The aerial beamwidth (conical) is 1·4° and the rotation speed nominally 13rev/min.

An Automatic Message Handling and Switching System

The STC 8300 *ADX* is the latest system developed by STC to be used in communications systems for switching telegraph and data messages. Throughout the world there is at present a requirement to replace existing message switching systems in large organizations. These systems are largely manual in operation, although a few are automatic, using electro-mechanical switching devices that are inherently limited in their speed of operation and have heavy maintenance requirements.

Since 1956, STC has been engaged in the development of automatic electronic message switching systems and the 8300 *ADX* System, which is the U.K. version of the very successful ITT 7300 *ADX* System, represents an advanced combination of telecommunications techniques and stored-program computer technology.

Message switching systems collect telegraph messages from teleprinters and automatic message transmitters which transmit their information over telegraph channels. When the messages are received by the switching centre they are examined and sorted according to the destination to which they should be sent. This sorting process takes into consideration such aspects as degree of priority and security. The manual systems in use employ a large number of operators to handle these messages within the centre. Besides the delays and routing errors incurred, these systems are very expensive to operate.

The STC 8300 *ADX* System, however, carries out all these functions entirely automatically and at very high speed. As a result of introducing *ADX* Systems into major networks, operational efficiency has increased considerably and the actual savings in operating and maintenance costs have been high.

With the increasing demand for more communications facilities, the 8300 *ADX* System has applications not only in message switching but also in other systems where a number of communications inputs and outputs have to be handled. One example of this is the automation of meteorological centres for the collection of data and the automatic plotting of weather charts. Another application is the handling of communications within a large organization.

Since stored-program processors are used for the 8300 *ADX* System, the system can be modified easily to meet the ever-changing requirements of communications systems. One example of this is the increasing use of data transmission. The 8300 *ADX* System can handle considerably higher speeds than those normally encountered on telegraph circuits.

A Slow Switch

By K. F. Knott[*], B.Eng.

This article describes switching a transistor by a method which enables the switching time to be readily variable up to several seconds. Also described is a Schmitt trigger with readily variable hysteresis.

IN some types of adaptive control systems a test variation of one or more of the system parameters is made. Data is extracted from the effects of this variation to determine whether the performance of the system is improved. In some instances it is desirable that the test variation be gradual, and furthermore be realized by solid state circuits. A simple way in which a parameter may be varied is to switch an attenuator into that part of the system which determines the parameter, for example, the velocity feedback loop of a position control servo.
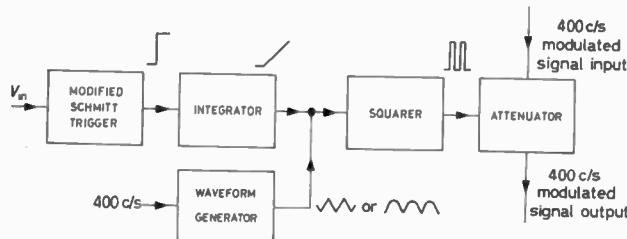
Fig. 1. Slow switch

This article describes a method of varying attenuation gradually between two predetermined values. The principle of operation may be applicable to many control systems but the particular circuit discused here was designed for the system outlined in the following paragraphs.

The system is essentially a position control system. The demand and output signals are impressed on a 400c/s carrier, this being the excitation frequency of the sensing devices. The error signal is amplified, demodulated and fed to the actuators. One of the parameters of the system is varied using an attenuator which is switched into the circuit on the demand of a trigger signal. The operation of the switch is summarized in the specification below:

Attenuator range	Zero to 2·7 in voltage
Switching time	4sec approximately
Attenuator switched in when	trigger voltage, $V_{in} = 10V$
Attenuator switched out when	trigger voltage, $V_{in} = 4V$

Principle of the Switch

The arrangement of the switch is shown in Fig. 1. The waveforms at different points in the circuit are also indicated.

The modified Schmitt trigger (Fig. 2) has two definite states, one for V_{in} equal to or greater than 10V, the other for V_{in} equal to or less than 4V. The state of the trigger for V_{in} between 4V and 10V is determined by the usual hysteresis effect. When V_{in} passes through a trigger level the Schmitt trigger produces a step function which is then integrated and added to a triangular or full-wave rectified sine wave. The sum so formed is amplified and squared. The output from the squarer is in the form of variable

* *Royal College of Advanced Technology, Salford.*

width rectangular pulses which are then applied to the attenuator.

The biasing of the circuit is such that for the two states of the Schmitt trigger the output pulses just have zero width or full width. In other words the output from the squarer sits at one of two d.c. levels depending on the trigger level but the time taken for the excursion betwen these two levels is governed by the integrator time-constant.
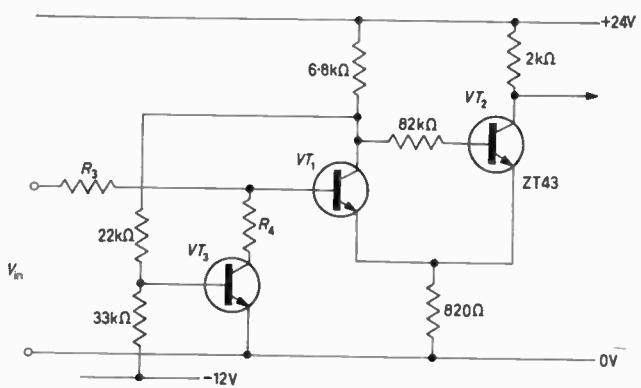
Fig. 2. Modified Schmitt trigger

A detailed explanation of the elements of the switch is given in the following paragraphs.

Modified Schmitt Trigger

The circuit diagram is given in Fig. 2. The circuit consists of a standard form of Schmitt trigger preceded by an attenuator.

Referring to Fig. 2, transistors VT_1, VT_2 and associated resistors comprise the Schmitt trigger which is designed to trigger with 4V on the base of VT_1 and to have negligible hysteresis. When V_{in} is less than 4V, VT_1 is off and its collector is at nearly +24V. Transistor VT_3 is conducting since its base will be positive and hence the attenuator consisting of R_3 and R_4 will be in the circuit. In order to trigger the circuit V_{in} will have to increase to satisfy the inequality:

$$V_{in} \geqslant \frac{R_3 + R_4}{R_4} \cdot 4V$$

When V_{in} satisfies this inequality VT_1 will be conducting and the base of VT_3 will be negative and hence R_4 will be out of circuit. If the current drawn through the base of VT_1 causes negligible voltage drop across R_3 the circuit will trigger again when V_{in} is decreased to satisfy the inequality:

$$V_{in} \leqslant 4V$$

If R_4 is replaced by a potentiometer the hysteresis may be varied readily over a wide range.

Complete Switch

METHOD (1) TRIANGULAR WAVEFORM

The circuit diagram is shown in Fig. 3. Referring to

Fig. 3. Complete switch with triangular waveform
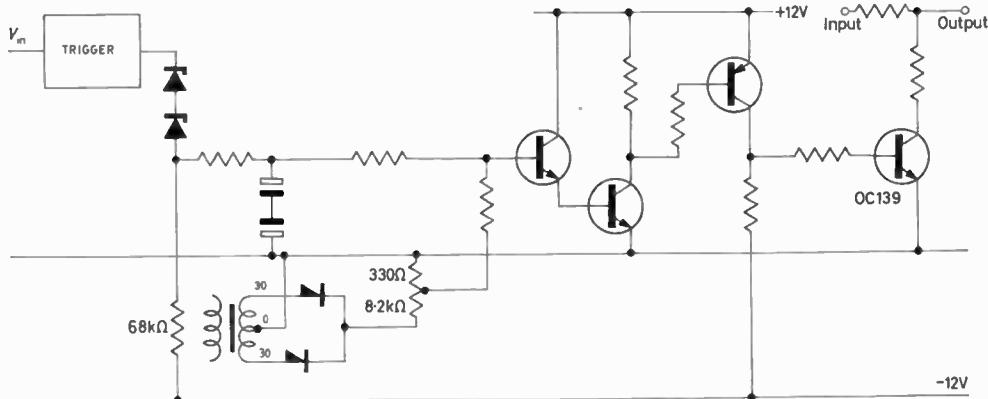
Fig. 4. Complete switch with rectified sine waveform

of the signal spent on the two values of attenuation. The average attenuation will thus vary with the pulse width.

Method (2) Rectified Sine Waveform

A further simplification in the circuit may be made if the triangular waveform is replaced by a full-wave rectified sine waveform. The circuit diagram is shown in Fig. 4. The switching period is not as well defined with this circuit however.

Distortion Introduced During Switching

Fig. 5 illustrates the distortion introduced by the two methods during the switching period. Owing to switching with the carrier the distortion bears a fixed phase relationship to the signal being attenuated. This distortion will be considerably reduced when the signal is demodulated and smoothed and may be quite acceptable. The use of the carrier frequency for switching would thus be justified. Distortionless switching can be achieved if the attenuator is switched at a much higher frequency than that of the signal it is handling. The output from the attenuator is then smoothed using a low-pass filter. The expected waveforms with high-speed switching are shown in Fig. 5(c).

Fig. 3, the triangular waveform is derived from the 400c/s energizing supply in the interests of circuit simplicity. The implications of switching the attenuator at the same frequency as that of the signal it is handling will be discussed later.

The 400c/s sinusoidal voltage is squared by two Zener diodes and then integrated to produce a triangular waveform.

The output from the Schmitt trigger is reduced in level by a 14V Zener diode and a potential divider to give the correct excursion of the pulse width. The reduced output from the Schmitt trigger is integrated by a simple RC network and added to the triangular waveform.

The sum of the integrator output and the triangular waveform is squared by a high gain amplifier and applied to the base of a switching transistor which is in one arm of the attenuator. In the steady state the base of this transistor is held to the $+12V$ or $-12V$ h.t. rails via a resistor.

The attenuator is a simple potential divider consisting of resistors R_1 and R_2 (Fig. 3). Assuming that the input impedance of the stage following the attenuator is sufficiently high, the steady state attenuation will be given by:

$$\text{Attenuation} = \frac{R_2}{R_1 + R_2} \text{ or Zero}$$

When the switch is triggered the attenuation will be switched between these two values by the rectangular pulses applied to the base of the switching transistor. The pulse width will determine the relative times per cycle

Results

The results obtained for the switching of a 2·7 to 1

Fig. 5. Distortion introduced during switching
(a) Method 1
(b) Method 2
(c) High frequency switching
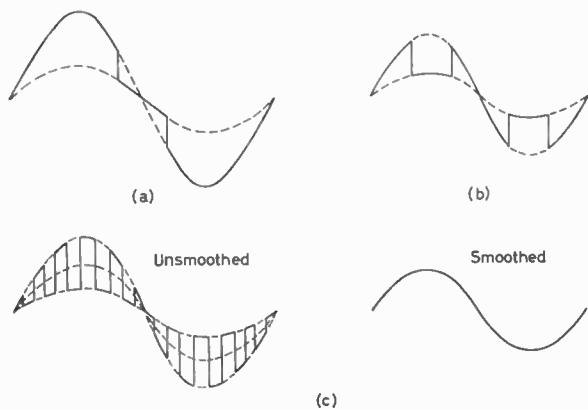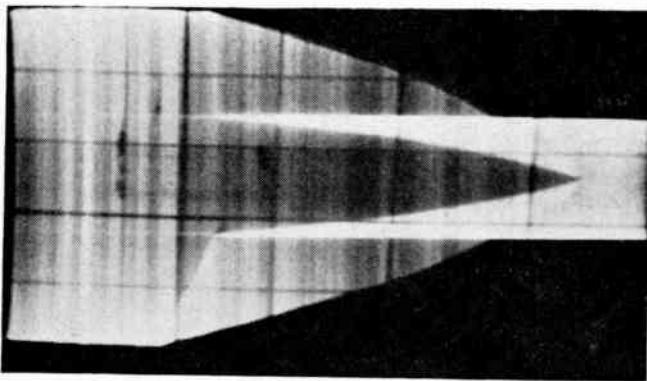
Fig. 6. Switch closing

Fig. 7. Switch opening

In both oscillograms the full width of the trace represents 6sec
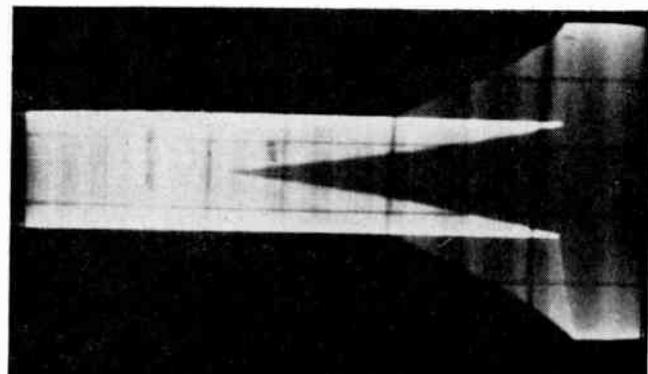
attenuator are illustrated by Figs. 6 and 7. These results were obtained with the circuit using the triangular wave-form.

Fig. 6 is an oscillogram showing the 400c/s signal output from the attenuator displayed on a slow time-base. At the beginning of the trace the attenuator is out of circuit. The Schmitt circuit is then triggered and the attenuator is gradually switched in. The distortion is indicated by the differences in brightness of the trace which occur during the switching period.

Fig. 7 shows an oscillogram of the attenuator being switched out of circuit. Similar results were obtained with the circuit using the rectified sinusoidal waveform.

Conclusions

Usually the emphasis in switching transistors is on the reduction of the switching time to as small a value as possible. However, for some applications it is required to increase the switching time to high values. The principle of the circuits described here provides switching times which depend on a simple time-constant and are essentially independent of the switching transistor characteristics. The switching time is thus readily varied over a wide range. The upper frequency limit of the signals the switch can handle will, of course, depend on the high speed switching characteristics of the transistor. In Method (1), the non-linearity of the change in signal level may be overcome, if required, by constant current charging of the integrating capacitor.

Numerical Machine Tool Control Systems in Germany

After a slow start, numerical machine tool control systems have progressed rapidly in Germany[1,2]. More new systems were on show at the recent German Machine Tool Exhibition in Hanover.

R & E Hopt KG exhibited their system 80050 which employs punched cards as input. The cards can be processed after production in existing punched card installations and existing equipment may be used in punching the cards in a straight binary code. An 80 column card with 12 lines is said to have a capacity of 20 operations, allowing for three axes movement and 12 bits for auxiliary functions.

In its simplest version the system has a maximum resolution of 2^{12}, which may be extended by five decades. The system is suitable for positioning and right angle movements. Absolute measurements operate on the photo-electric principle. For linear displacement the feed is switched to inching on approaching the preset nominal value. Silicon semiconductors are employed throughout in the comparator stages and card readers to ensure reliable working over a temperature range from 0° to 100°C. Small dimensions are obtained by use of micrologic circuits on printed circuit cards.

Brown, Boveri & Cie AG exhibited their input units AW 27, AW 40 and AW 60, developed for cam-operated machine tools. These units may also be added to already existing machines. A hand-punched endless plastic foil is used as programme medium, having 27, 40 or 60 channels or 'columns' respec-

tively. One operation is allocated to each channel and several operations may be programmed in one line. The transport mechanism feeds the programme through the reader line by line and the operations programmed in one line are read at the same time and not in the serial form inherent in punched tape inputs. The unit may be coupled with an electric type-writer to record the operation of the programme. The equipment was demonstrated on machine tools on several other stands, e.g. a drum turret lathe, a vertical automatic lathe, and a milling machine.

'Sinumerik' is a new versatile numerical control system in unit construction with plug-in function sub-assemblies, for which high accuracy is claimed. Positioning as well as dis-placement control is possible with either digital or analogue measurements. A digital incremental displacement principle is used for the simplest positioning. A counter is preset at the commencement of the operation and counted down by pulses. An inserted corrective logic eliminates the danger of over-running. Encoders with Hall generators and sensors offset by 90° supply the measurement feedback with appropriate direc-tional sign.

Where datum measurements are used, pulses arriving from the directional logic are fed into a bi-directional counter. For this application the shaft encoder may be replaced by a trans-ducer giving a digital output. The design allows operation in temperatures up to 65°C.

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1. Numerical Machine Tool Control. *Electronic Engng.* **36**, 492 (1964).
2. A Numerically Controlled Punching Press. *Electronic Engng.* **36**, 626 (1964).

LETTERS TO THE EDITOR

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

Simple Tuneable RC Null Networks

DEAR SIR,—Douce and Edwards[1] have been anticipated by White and Morgan[2]. In addition to varying one component for adjusting the null of an *RC* parallel-T they illustrated the circuit shown in Fig. 1 to provide the extremely broad tuning range of two decades.
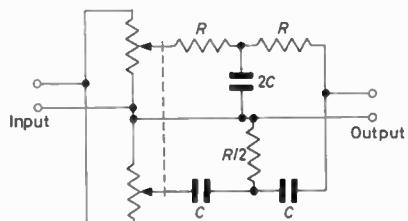
Fig. 1. Tuneable parallel-T with two decade range (White and Morgan) $\omega = 1/RC$
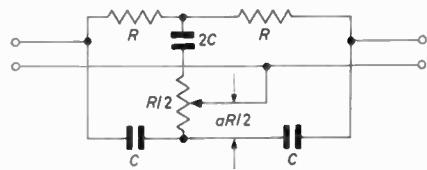
Fig. 2. Tuneable parallel-T (Andrevev)

$$\omega_0 = \frac{1}{RC\sqrt{1-\alpha^2}}$$
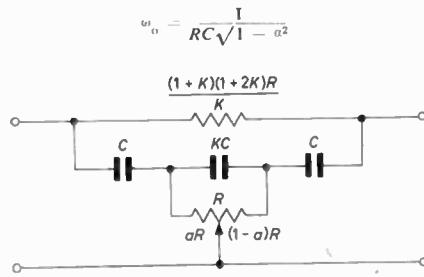
Fig. 3. Tuneable null circuit (Hall)

$$\omega_0 = \frac{1}{RC\sqrt{\alpha(1-\alpha)}}$$

Adrevev[3] has published a very useful adjustable null circuit based on the parallel-T in which the frequency is varied by a single potentiometer as shown in Fig. 2.

There are so few *RC* null circuits which are tuneable by varying a single potentiometer; one due to Hall[4] is shown in Fig. 3 for the sake of completeness.

It should be mentioned that all of these circuits show a variation in insertion loss, skirt selectivity and symmetry when the null is varied.

Yours faithfully,

T. C. PENN,
Physics Research Laboratory,
Texas Instruments Inc., U.S.A.

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1. DOUCE, J. L., EDWARDS, K. H. A Simple Null Filter with Variable Notch Frequency. *Electronic Engng.* 36, 437 (1964).
2. WHITE, C. F., MORGAN, K. A. The Dual-Input Parallel-T Network. *Proc. Nat. Electronics Conf.* 8, 588 (1952).
3. ANDREVEV. *Telecommunications (USSR)* No. 2, 1960, p. 195 (Pergamon Press translation).
4. HALL, H. P. RC Networks with Single Component Frequency Control. *IRE Trans. Circuit Theory*, p. 283 (September 1955).

Non-linear CR Circuit Using Silicon Carbide Varistors

DEAR SIR,—I have read the article published by W. G. P. Lamb in the October issue. The step response of a *CR* network consisting of a linear capacitance and a non-linear resistance has been calculated at two special values of the parameters α. See the author's Fig. 1. However it is not difficult to give a more general solution of that problem. The following solution is taken from my article published in the Swedish journal *Elteknik*, December 1958.
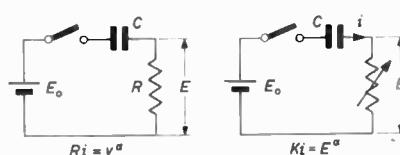
Fig. 1 (left). Basic circuit

Fig. 2 (right). Notation used
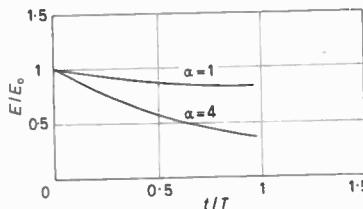
Fig. 3. Effect of α

$$E/E_0 =$$
$$\frac{1}{\left[\dfrac{t(\alpha-1)E_0^{\alpha-1}}{KC} + 1 \right]^{1/(\alpha-1)}} , \text{ for } \alpha \neq 1$$

$$\dots\dots\dots (1)$$

In equation (1) I have used the same notation as the authors except *K* which I have used instead of the author's *R*. See Fig. 2. The use of *R* is very misleading because *R* is not a resistance (*RC* is not a time-constant), except for the linear case α = 1. In the author's solution for α = 2.

$$E/E_0 = \frac{1}{(E_0 t/RC) + 1} \dots\dots\dots (2)$$

For an arbitrary value of α, *R* has the dimension volts$^\alpha$/amperes. Because of the misleading term *RC*, the authors are coming to the following wrong conclusion "equation (2), relevant to α = 2, gives a curve which is initially steeper than the exponential case".

The true time-constant *T* for the non-linear case can be defined by the equation

$$T = \frac{KC}{(\alpha-1)E_0^{(\alpha-1)}}, \text{ for } \alpha \neq 1$$

$$\dots\dots\dots (3)$$

The time-constant for the non-linear circuit is consequently dependent on the input voltage E_0. The rate between the initial step at a linear and a non-linear circuit at the same value of the time-constant *T*, is (α − 1) for α ≠ 1. At for instance α = 4, the linear circuit is three times initially steeper than the non-linear circuit. See Fig. 3.

Yours faithfully,

L. BRANDQVIST,
AB Scienta,
Gothenburg, Sweden.

The Author replies:

DEAR SIR,—Your correspondent appears to have compared the initial steepness of the decay curves for circuits of different α's but of equal time-constant *T* where this is defined by the relation

$$T = \frac{KC}{(\alpha-1)E_0^{\alpha-1}}$$

and as such correctly finds that as α is increased the initial decay is reduced in steepness.

In my article, however, I compared the initial rates of decay of circuits of different α but employing variable resistors having the same value of resistance when measured under conditions of 1V potential difference. Under these conditions, the higher values of α give a steeper initial decay when E_0 is greater than 1V. It is easy to see that this is so, since the resistance of the non-linear resistors decreases as E_0 is raised.

Yours faithfully,

W. G. P. LAMB,
St. John Cass College,
London E.C.3.

A Novel Integrator and its Applications in Null-Measurement Systems

SIR,—I regret that an error has appeared in my article in the November 1964 issue. The error is as follows:

Fig. 2.
The output *y* should read "to Fig. 6".
The output *x* should read "to Fig. 5".

Fig. 5.
The input from *x* should read "from *x* Fig. 2".

Fig. 6.
The input from *y* should read "from *y* Fig. 2".

Yours faithfully,

H. C. BERTOYA,
SIRA, Chislehurst, Kent.

Short News Items

The Low Temperature and Materials and Testing Groups of The Institute of Physics and The Physical Society are organizing a European Symposium on 'Electrical Conduction at Low Temperatures' to be held at the Royal Aeronautical Society, London, on 20 to 21 May 1965.

The Symposium will be concerned with the factors which limit the electrical conductivity of normal metals and superconductors.

Review papers will be given by Professor J. Friedel, Professor J. L. Olsen, Dr. A. C. Rose-Innes and Professor J. M. Ziman. Original contributions are invited and outlines not exceeding 300 words in length should be sent to Dr. P. Graneau, McFadzean Laboratory, 38 Wood Lane, London, W.12, to arrive not later than 15 February, 1965. Outlines should be prepared in the standard form described in leaflet ACB78 obtainable from the Institute and Society.

Advance registration for the Symposium will be necessary and further details and application forms will be available in March 1965 from the Administration Assistant, The Institute of Physics and The Physical Society, 47 Belgrave Square, London, S.W.1.

The Sixtieth Physics Exhibition, organized by the Société Française de Physique of 33 rue Croulebarbe, Paris 13, is to be held at the Parc des Expositions at Porte de Versailles. Paris on 14 to 19 December this year.

The Electronic Engineering Association has joined EUROSPACE as a Collective member following the formation of the E.E.A. Space Committee in July 1963.

EUROSPACE, the European Industrial Space Study Group is already collaborating closely with ELDO and ESRO.

The Third Annual Exhibition and Symposium on Industrial Electronic Components and Instruments is to be held at Cabot House, Bristol Technical College on 8 to 11 December.

Details of the Exhibition and the lecture programme are obtainable from Wireless-Electric Ltd, St. Thomas Street, Bristol, 1.

The "Silicon Gate-Controlled Switch or Gate Turn-off Thyristor" is the subject of a one day symposium to be held at the Enfield College of Technology on 2 December this year.

Details of the programme and registration are available from R. D.

Kitchener, Head of the Electrical Engineering Department, Enfield College of Technology, Middlesex.

E.M.I. Ltd has introduced a circuit for the protection of image intensifier tubes using multiple electrodes and high voltages.

It has been found that if, during operation, a spark occurs between one electrode and earth, this often results in the destruction of or damage to the tube, because of the sudden application of a voltage which the tube was not designed to withstand between the electrode to which the spark occurs and adjacent electrodes.

BINDING OF VOLUMES

Readers can have their copies of ELECTRONIC ENGINEERING bound. complete with index and with advertising pages removed, in a good quality red cloth covered case, lettered in gold on the spine, including packaging and return postage at a cost of £2 10s. per volume.

Home and overseas readers who require their issues for 1964 bound, are asked to comply with the following instructions:

Tie the issues together, enclose a remittance of £2 10s., with the senders name and address, and despatch carriage paid in a closed parcel to:

The Circulation Dept. (E.E. Binding), 28 Essex Street, Strand. London W.C.2. (Cheques or postal orders should be made payable to Morgan Brothers (Publishers) Ltd.).

Indexes:

Please note that the index for Volume 36 (1964) is bound in with this issue (December, 1964), and should be sent with copies for binding.

The following are also available from our Circulation Dept.:

Complete bound volumes for 1961. 1962, 1963 and 1964, price £4 4s. Reprinted volumes for the years 1940 to 1954 are also available, price £7 10s. per volume.

If readers wish to arrange for their copies to be bound locally, cases for permanent binding of volumes can be supplied at 7s. 6d. each. Postage: 9d.

If capacitors are connected between adjacent electrodes, the values of their capacitances preferably being inversely proportional to the voltage between adjacent electrodes, these difficulties would be much reduced. The values of the capacitors should preferably be such as to give time-constants of at least a millisecond in conjunction with the potential-supplying network.

Lancashire Dynamo Electronic Products Ltd (M.I. Group), is installing a modulated photo-electric equipment series MPU.2, at the Haulbowline Lighthouse, at the entrance to Carlingford Lough, Ireland. The order was placed by the Commissioners of Irish Lights, Dublin.

The equipment, to be used for fog detection and automatic control of a large Swedish supertyfon fog horn, will span the 1 300 yards from the north bank of the estuary to the lighthouse which rises from the sea bed to mark the approach channel, with a light beam.

Once the beam is broken by poor visibility an integral relay system automatically brings into operation the audible fog system apparatus. This operation ceases when visibility clears.

The equipment consists of a light projector unit and a light receiver unit—each fitted with a 6in lens, precision optical viewing sights with a ×4 magnification factor for ease of alignment, and a sensitivity control on the receiver unit for simple adjustment of the desired operating level.

The use of semiconductors and 'fail to safe' features ensure a high degree of reliability.

Further international discussion on colour television took place in London during October this year. Demonstrations were given by the BBC and the British Radio Equipment Manufacturers' Association to the E.B.U. Colour Television Group; this group includes representatives of the Broadcasting Oranizations. the Telecommunication Administrations and the Radio Industry in France, the German Federal Republic, Italy. the Netherlands. Switzerland and the United Kingdom. Representatives of broadcasting and other authorities interested in the problem in Eastern Europe also attended.

These demonstrations mark a further step towards choosing a common colour system for Europe, the final decision on which is to be made at a meeting of the C.C.I.R. (International Radio Consultative Committee) in Vienna next

March. The demonstrations compared colour pictures produced by the three systems, NTSC, SECAM and PAL, when transmitted over long international links, when recorded on video tape, and when transmitted direct to domestic receivers. They showed the effect of important improvements in colour television apparatus that have been made since the previous series of demonstrations in London last February.

Submarine Cables Ltd (owned jointly by Associated Electrical Industries Ltd and British Insulated Callenders Cables Ltd) has received a £500 000 contract from the Post Office for a repeatered submarine cable, providing 480 telephone circuits, to be installed between Bournemouth and Jersey in 1967.

The armoured 0·935in coaxial cable will be about 137 nautical miles in length and equipped with 18 transistorized submersible repeaters. The contract also includes special transistorized terminal-station equipment manufactured by the Telecommunications Division of Associated Electrical Industries Ltd, one of Submarine Cables Ltd's parent companies.

The transmitter at the Post Office Radio Station at Rugby which operates at a frequency of 16kc/s and is used for the radiation of time signals four times daily as well as for some 22 hours daily of morse transmissions, is to be modernized.

The present output stage consisting of three panels each containing 18 water-cooled valves is being replaced by three vapour-cooled valve stages, and the lower power stages by three wide-band amplifiers. This will enable the present output power of 210kW to be doubled and will also allow of teleprinter operation using 'frequency-shift' type keying in addition to the present hand-speed morse. The output circuit will also be arranged to be self-compensating for changes in the electrical performance of the aerial due to changing weather conditions.

The aerial system, supported on twelve 820ft masts was completely renewed a few years ago.

The Nineteenth Annual Frequency Control Symposium sponsored by the U.S. Army Electronics Laboratories, Fort Monmouth, New Jersey is to be held at the Shelburne Hotel, Atlantic City on 20 to 22 April 1965.

The Symposium, at which some 30 papers will be presented, will cover the past year's progress in research and development of quartz crystal devices and their application to frequency control selection, atomic and molecular resonance devices, piezo-electric resonators, masers, fundamental properties of quartz, and related matters.

Further information can be obtained on application to the Director U.S. Army Electronics Laboratories Hq, U.S.

Army Electronics Command (Attention: AMSEL-RD-PF) Fort Monmouth, New Jersey.

Welwyn Electric Ltd of Bedington, Northumberland has installed in its testing department new load testing chambers of its own design and construction.

These test chambers which have a capacity up to 46 000 components are designed to cope with a wide range of conditions and type of component. The test supplies may be 6V to 1500V a.c. or d.c., cyclic or continuous, and are available in any combination for connexion to any tray of components by means of a patch-board of plugs and sockets. A wide range of ambient conditions can be obtained from room temperature to 200°C.

In addition to the general purpose chambers, specially designed chambers are used for longer term reliability studies. Four chambers are in use, three of them for oxide film resistors with a total capacity of 9 750 separate components (85 000 000 resistor-hours per year). The other chamber has a capacity of 500 thin film integrated circuits, each containing 10 resistors, producing 44 000 000 resistor-hours per year.

The test house which is approved by the Ministry of Aviation has facilities for climatic and durability testing to the latest Defence Specifications. These facilities include, in addition to the usual climatic chambers, a mechanical testing department which is equipped with apparatus for vibration and acceleration tests. A 750lb thrust vibrator and its ancillary equipment is used to test components over a wide range of frequencies to simulate their use in dynamic equipment.

A computer is now used in the analysis of test data, thus greatly assisting the testing staff by shortening the time previously taken to collate masses of accumulated information.

Two transistorized repeaters have now been installed in the existing St. Margarets' Bay—La Panne telephone cable and will thus increase the capacity of the cable from 216 to 420 channels.

The cable becomes the first transistorized submarine cable system in the world.

The v.h.f. sound transmitters at the BBC's relay station at Hereford have now been brought into service thus completing another of the low-power combined television and v.h.f. sound relay stations being built by the BBC to extend and improve sound and television services. The television transmitter at Hereford was brought into operation in September this year.

The Hereford station will transmit the Midland Home Service on 94·1Mc/s, the Light Programme on 89·7Mc/s, and the Third/Network Three Programmes on 91·9Mc/s. Horizontal polarization will

be used which means that horizontal receiving aerials should be used.

These new v.h.f. transmitters will provide improved reception of the BBC's three sound programmes for some 40 000 people in the City of Hereford and its environs.

'Thin Film Electronics in the U.S.A.' is the title of Report 1002 issued by the Electrical Research Association to its members in June 1963 and which is now available for general release.

This report is based on the investigation carried out by the author, G. Siddall, into the work being done in American industrial laboratories and research institutions before a thin film research programme was started in the Electronics Department of the Electrical Research Associations at Leatherhead.

Copies of the report priced at 18s. plus 8d. postage, are obtainable from the Electrical Research Association, Cleeve Road, Leatherhead, Surrey.

Cableship 'Mercury' has now made the final splice of the new cable system between Singapore and Jesselton in Sabah (North Borneo), 860 nautical miles, and between Hong Kong and Jesselton, 1 050 nautical miles, by the beginning of next year.

This link is the first stage of the South East Asia Commonwealth Submarine Cable System known as SEACOM which is due for completion in 1966 and which will link into the existing Atlantic and Pacific Systems CANTAT and COMPAC.

Submarine Cables Ltd has supplied 1 400 nautical miles of submarine telephone cable and 25 submersible repeaters, costing £2½M in all.

Standard Telephones and Cables Ltd has supplied some 700 nautical miles of armoured shallow water cable and 70 repeaters together with 10 submersible equalizer units at a cost of £2½M.

The Royal Aircraft Establishment, Farnborough has placed an order valued at £25 000 with M.E.L. Equipment Co. Ltd, Crawley for the design and manufacture of a data handling system for the thermal balance testing of satellites.

The system will handle up to 200 channels of temperature or other measurements from a satellite placed in an environmental testing chamber in which conditions of space will be simulated. Such conditions include very low pressures and the extremes of temperature resulting from full solar radiation and complete shadow.

In the data handling system, the inputs will be connected to a sampling unit where they will be sequentially switched through a single channel of analogue-to-digital conversion and then punched out in a suitable computer code on paper tape. Provision will also be made for the connexion of certain of the inputs to off-limit detectors and pen recorders.

BOOK REVIEWS

Communication Satellites

By G. E. Mueller and E. R. Spangler. 280 pp. Med. 8vo. John Wiley & Sons. 1964 Price 75s.

THE standard of writing throughout this book is extremely high, and it is clear that a great deal of attention has been given to its balance and structure, not least to the association of diagrams and relevant text.

References are comprehensive, as is the 18-page index, the only criticism of which is that, if anything, it is too detailed. By sheer force of circumstance the subject matter of most aerospace books is almost entirely American in origin. 'Communication Satellites' does not follow this pattern completely—over a dozen references are made to European sources, notably to S. N. Barker on 'Group Synchronization of Binary Digital Systems' in two separate chapters.

One of the interesting sections of the book is devoted to 'Requirements' (Chapter 2). This is an extremely short—six page—chapter, but provides a useful summary of the position; and, in Table 1, a forecast of the growth of communication traffic to be expected by 1970. Comparison is made between the figures for 1960 and those predicted for the end of the ten-year period. It is thought that by this time the total overseas traffic of the U.S. will be carried by three satellites with a bandwidth per satellite of approximately 12Mc/s. Most of this bandwidth—10Mc/s—will be taken up by two 5Mc/s television circuits, although it is considered that the growth rate of television will be much less than for, say, telephony or telex time. The predictions for 1970 are 20 000 000 telephone messages, and 22 600 000 minutes of telex (actual 'units' in the table are given as thousands of messages and thousands of minutes respectively).

It is a little surprising to find that there is a chapter devoted to 'Active versus Passive Satellites', but its inclusion must be commended. Certain fundamental points are made and nine references are given.

The remainder of the book is built up to include the many techniques involved, finishing with a succinct chapter on 'Cost'.

R. E. YOUNG.

Matrix Analysis for Electrical Engineers

By Alexander von Weiss; translated by Egon Brenner. 172 pp. Med. 8vo. D. Van Nostrand. 1964. Price 62s.

TO the mathematician the concept of a matrix enables one to write a large number of equations as simply as a single equation. Thus $y = ax$ is one equation between two quantities or a matrix equation between two long columns of quantities x and y. He is thrilled because many matrices can be handled like numbers, but he is still concerned about the best way of handling the numerical properties of matrices.

The electronics engineer, no doubt, is in much the same position but expects matrices not only to simplify the presentation of linear relationships, but to indicate an orderly and rapid solution of his problems.

Dr. Weiss presents the mathematics in an admirable manner, lards it with plenty of examples, all carefully described. His selection of material is excellent and adequate because he does not hesitate to state results without proof when this is desirable (the names and notation of Weiss, however, are not those used by mathematicians in their proofs).

Moreover, he has a knack of suggesting that certain parts of his work should be supplemented by further reading, and impelling the reader to do just that.

The second half of the book on applications in electrical engineering is very interesting. The initial examples are just simple enough to illustrate adequately the analogues of matrix rules and lead inevitably to more complicated networks with a final chapter on three-phase problems.

It is considered that the book is nearly as good as the publishers claim, but that a useful addition would be references to sources of information on computation problems which are the natural result of using these methods.

G. S. KYNCH.

Telecommunications

By J. Brown and E. V. D. Glazier. 370 pp. Med. 8vo. Chapman & Hall. 1964. Price 45s.

IN the early days of electronics it was possible to write a book describing both theory and practice. With the rapid development of our knowledge this is no longer possible so that the principles and the techniques must be considered separately if, at least for pedagogic purposes, confusion is to be avoided. For example, most modern books on communications which come from the U.S.A. are concerned only with the basic theory leaving the detailed circuits to be dealt with in other, more appropriate, texts. All too often, however, British authors in this field have crammed in potted chapters on microwave theory, circuit design and so on, with the result that the communications principles are also somewhat perfunctorily treated.

The present text, which is the first of a proposed series on telecommunica-

tions, embodies a welcome change, as far as this country is concerned, from the latter approach. It is mainly concerned with the fundamental principles on which modern communications systems are based. The application of these principles is illustrated by a description of the main techniques used, but the theory is "not coloured by descriptions of the electronic circuitry required for its practical realisation". It is pleasing to report that the authors have produced a clearly-written, authoritative book well suited to the needs of degree-level (or equivalent) students of electrical communications and perhaps also as a reference book for practising engineers.

The fundamentals deal mainly with the time- and frequency-domain equivalents of a signal, the characteristics of communications channels, information theory and noise. In addition, there is a sizeable appendix dealing with transform methods. The techniques described include telegraphy, telephony, television, telemetry and radar.

Some minor criticisms could be made, but these are not serious enough to warrant mention here except, perhaps, for the fact that in some respects the book is not as up to date as it might have been. The title itself, for example, is rather out of fashion these days. Nevertheless this text is a welcome addition to the rather few good existing books on this topic.

W. A. GAMBLING.

Principles of Tunnel Diode Circuits

By Woo F. Chow. 387 pp. Med. 8vo. J. Wiley & Sons Ltd. 1964. Price 94s.

THE contents of this book have been assembled from the works of many authors whose particular contributions are acknowledged both in the text and in the extensive bibliography. Unlike some works on a new subject, the author has done much more than assemble knowledge; he has assimilated it and presented it in a unified and balanced manner.

Sufficient semiconductor physics is included in the early part of the book to enable the student to grasp the principles underlying the operation of the tunnel diode. For the applications engineer there is much useful information concerning the dependence of the basic parameters of the device upon temperature and nuclear radiation. Methods of parameter measurement and many of the special difficulties likely to be encountered are dealt with in detail.

The first half of the book contains chapters dealing with small and large signal equivalent circuits, noise, amplifiers, conversion, detection and oscillators, while the second half is devoted to digital applications. In a few instances the author makes a comparison between the tunnel diode circuit and the tran-

sistor equivalent, particularly when the tunnel diode circuit has some special virtue. One is left with the feeling that it has a special place in digital circuits as so much space is devoted to this subject. It would have been a help if some indication as to its suitability had been given in a conclusion to each section.

In his preface the author states that "the purpose of the book is to satisfy the need for a better understanding of tunnel diodes and to prepare the student or engineer for further study and application in this area". This he has achieved, but from the point of view of an application engineer some additional worked examples, using typical tunnel diode parameters, would have been a help. This would have enabled the reader to assess the relative merits of a particular circuit, when compared with its more conventional counterpart, without the labour of inserting practical values in the numerous design equations given in the book.

As a source of tunnel diode circuits and their detailed analysis this book undoubtedly fills a need; it may well become one of the standard reference works on the subject.

J. MERRETT.

Space Vehicle Electronics

By D. B. Newman. 397 pp. Med. 8vo. D. Van Nostrand. 1964. Price 122s.

THAT 'Space Vehicle Electronics' contains, at most, the equivalent of one chapter length of text* devoted to electronics as generally understood, will appear far less strange in America than in this country.

The book does, in fact, reflect the position in the U.S.A. where the tendency is for the engineer with an electronics background to be responsible for many more aspects of aerospace work than in the United Kingdom. Thus in the American context it could be claimed that 'Space Vehicle Electronics', in conjunction with its wide-ranging references, has to cover this huge field in order to satisfy the needs of the electronic engineer.

The spread of subjects can only be described as enormous. As a result, an alternative title of 'Aerospace Physics—Pure and Applied' does almost inevitably come to mind, and must be associated with the criticism that so many major points can only be touched upon because of lack of space.

This general approach tends to produce lack of clarity, particularly in diagrams. Examples of the latter, selected as being completely typical, are Figs. 5-19 and 9-3, 'C.W. Detection Ranges' and 'Vehicle Shadowing of Antenna Coverage' respectively. In the first, for instance, shaded areas A, B, C, and D, are shown on this graphical presentation with no reference made to them in the

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Reviewer's estimate.

text. There are other features which appear not to be explained, including the parameters of the 'augmenter'. Similarly Fig. 9-3, although much simpler than 5-19, has virtually no description of what it purports to show, and is not clear in itself as a drawing. (It is possible to give more than one interpretation to the components of this illustration, particularly to what can be deduced to be the 'vehicle' which is drawn as being hollow.)

Unfortunately such weaknesses can be found throughout the book, and are almost certainly due to the attempt which has been made to include so much material in a book of nine chapters. This concentration process is most uneven in its effect. It is not unfair to take 'Vehicle Design' as an example. This major subject is compressed into some 14 pages—the second shortest chapter in the book.

Nevertheless due regard must be paid to the amount of data and the references which have been gathered together in one volume. This is especially noticeable in 'Space Environment' (by far the longest chapter with over 110 pages). Table 2, for instance—extending up to 32 sections—is remarkably comprehensive and perhaps unique, and includes physical data and surface data for the planets, a section on asteroids, and data on the brighter discrete radio stars.

R. E. YOUNG.

Electronic Fundamentals and Applications

By J. D. Ryder. 693 pp. Demy 8vo. 3rd Edition. Sir Isaac Pitman. 1964. Price 55s.

The third edition has been almost completely re-written and presents an integrated treatment of solid-state devices and the older vacuum and gaseous forms of active electronic devices. Many of the fundamental phenomena underlying both types have been consolidated since each area is based on a knowledge of electron energies. charge movements, and conduction in solids, gases, or vacuums. In the study of the individual device forms, the theory of the transistor and other solid-state types generally receives prime consideration. whereas the vacuum devices are discussed less fully, due to similarities in the concepts and methods involved.

Much of the specialized vacuum and gas tube material has been condensed. The subject of amplifiers has been expanded by further study of pulse response and the chapter on modulation now contains an introduction to pulse forms of amplifiers and generators attaining prominence in the u.h.f. field are also included.

Cathode Processes in the Mercury Arc

By I. G. Kesaev. 345 pp. Med. 8vo. Consultants Bureau, New York. 1964. Price $17.50

This is claimed to be the only book in English devoted to cathode processes and the stability of the mercury arc.

On the basis of his investigations of cold arc phenomena, the author concludes that the stable state of the arc is actually a state of dynamic equilibrium between processes of decay and information of cells. This point of view leads to a new concept of discharge control.

Detailed information on the movement of the spot and its division under experi-

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mental conditions conclude this work, which is a major contribution to understanding of the cold arc phenomenon.

Semiconductor Thermoelectric Devices

By A. I. Burshteyn. 131 pp. Demy 8vo. Temple Press Books Ltd. 1964. Price 35s.

This book, translated from the Russian, is a graduate level text, designed to acquaint the applied physicist and engineer with the physical theory and limitations of thermoelectric devices based on semiconductors.

The introduction gives a brief survey of the history and practical applications of thermoelectricity, and is followed by a chapter on the general theory of steady-state heat flow through a conducting rod. The remaining three chapters are devoted to the principles of design of scientific types of systems, with exhaustive presentation of generalized analytical solutions, as well as simplified forms applicable to various conditions.

Storage Batteries

By G. Smith. 211 pp. Demy 8vo. Sir Isaac Pitman & Sons Ltd. 1964. Price 27s. 6d.

This book describes how vastly different duties are catered for by a wide range of batteries designed to have characteristics to suit operating conditions.

Battery design, performance, and behaviour in service are fully described, together with many illustrations, characteristic curves and explanatory tables.

A chapter is devoted to charging principles, methods, and equipment. Since reliability is service depends to a very large degree on correct charging procedures, further charging details and recommendations relative to particular applications appear in the appropriate chapters.

ELECTRONIC EQUIPMENT

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

(Voir page 861 pour la traduction en français; Deutsche Übersetzung Seite 868)

DATA LOGGER
Cambridge Instrument Co. Ltd, 13 Grosvenor Place, London, S.W.1
(Illustrated below)

The Cambridge-Allscott data-logging equipment, which has been developed in collaboration with James Scott (Electronic Engineering) Ltd, indicates and records in digital form the measured value of any number of channels in a sequential programme, together with the time of measurement and an identifying number. It also incorporates an adjustable alarm system that provides
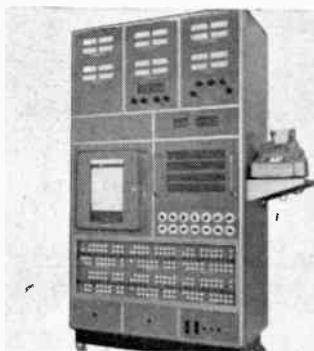
visible warning whenever the measured value of any channel or group of channels moves outside pre-set limits.

The equipment measures and records any variable that can be converted into an appropriate electrical potential or change of resistance, and mixed inputs can be accommodated by selecting the appropriate modular units and primary measuring elements. The channel inputs are connected in sequence to a Cambridge servo-rebalance measuring instrument by means of Allscott scanning units controlled by an Allscott automatic cycler unit. A binary shaft encoder, fitted to the measuring instrument, provides a digital output that is translated to decimal form and stored together with the channel number and time of measurement. This not only allows the measuring instrument to be switched to the next channel while the print-out is taking place, but also provides continuous digital read-out of the channel number and measured value. Channels are automatically scanned in accordance with the programme selected at the rate of one every three seconds, and manual control is also available allowing any channel to be selected by depressing the appropriate key-switch on an annunciator panel.

The equipment normally incorporates seven 'high' and seven 'low' adjustable alarm set points, providing 49 combinations of alarm limits. Additional alarm levels can be provided if required, and any set point can be linked to any single channel or group of channels by means of a patchboard selection unit. The measuring instrument incorporates a transmitting slidewire that provides the alarm level unit with a reference signal proportional to the measured value of the selected channel. If the measured value of a channel lies outside the alarm limits, control contacts are energized and the appropriate lamp on the alarm annunciator panel glows red, remaining on until manually cancelled. The digital read-out and print-out are displayed in red whenever the selected channel is in an alarm condition.

Provision is made to give digital records of channel number, measured value and time of measurement, but a strip chart recording mechanism can also be incorporated in the measuring instrument to provide an analogue record of the measured values of up to 100 channels.

EE 76 751 *for further details*

R.F. MICROVOLTMETER
Distributed by: Claude Lyons Ltd, 76 Old Hall Street, Liverpool 3
(Illustrated below)

The Millivac r.f. microvoltmeter type MV-28B has eight voltage ranges from 1mV to 3V full scale over the frequency range 10kc/s to 1 200Mc/s, the lowest readable voltage being 300μV. Accuracy is 3 per cent from 100kc/s to 60Mc/s. 5 per cent from 50kc/s to 200Mc/s and 10 per cent over the remainder of the range. Input impedance with the high-impedance probe tip is approximately 2 to 3pF and 75kΩ: a 50Ω BNC termination adaptor is supplied as standard, and accessories available include 100:1 capacitive voltage divider, N and G.R.

terminations and type N 'T' bridging adaptor.

The mirror-scaled meter is calibrated in both volts and dBm (relative to 1mW into 50Ω). Response is close to r.m.s. up to 30mV and peak above 30mV, calibrated in equivalent terms of r.m.s. sine wave. A useful point is that by the use of the 100:1 capacitive voltage divider the range of usefulness as an r.m.s.-responsive instrument can be increased to 3V.

EE 76 752 *for further details*
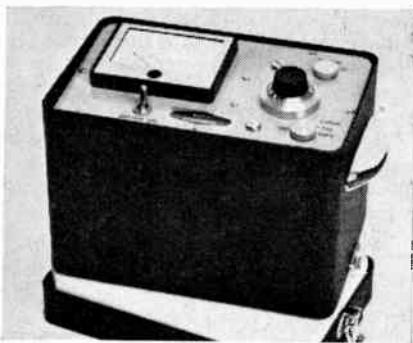
MOISTURE METER
Kappa Electronics Ltd, 159 Hammersmith Road, London, W.6
(Illustrated above)

The model AB66 transistor moisture meter is a portable battery-operated instrument designed to measure, from one side, the moisture content in sheet materials and materials or substances that possess a reasonably flat surface suitable for making physical contact with the 'sensing plate' that is mounted into the base of the equipment housing.

By this method the equipment measures the effect of the material upon the 'high frequency' electrostatic field radiated from the surface of the 'sensing plate', the effect being due to the electro-chemical properties of the material, of which, for many substances, the moisture content is a major factor.

Dependent upon the composition of the material and its moisture content, the instrument provides an indication that is relatable either to dielectric constant or in the case of highly conducting substances, impedance at high frequency. Good discrimination to changes in dielectric constant and impedance is afforded by a heterodyne detection circuit.

In operation the instrument is placed on to one surface of the material and the calibrated TUNE control adjusted to produce a maximum deflexion of the panel meter. The TUNE control setting is then read and converted into moisture content by reference to a chart or table that has previously been prepared with

the aid of samples of known moisture content.

Measurements on relatively thin sheet materials are conducted either by laying the material on to the flat surface of the base tray that is supplied with the instrument or, when more convenient, by placing it on a larger metal backed sheet of expanded polystyrene.

Dependent upon the selected type of 'sensing plate' and its depth of field penetration, relatively thick materials can be tested *in situ* provided that care is taken to ensure that the site conditions are repeatable.

EE 76 753 *for further details*

VALVE-VOLTMETER
K.L.B. Electric Ltd, 335 Whitehorse Road, Croydon, Surrey
(Illustrated below)

The PACO type V.70 valve-voltmeter is a low-priced, versatile instrument

which has seven d.c. ranges from 0 to 1 500V and seven a.c. ranges from 0 to 1 500V r.m.s., and 0 to 4 000V peak-to-peak plus a resistance range from 0 to 1 000MΩ and decibel range of −6dB to +66dB. The instrument is complete with a 3-way probe. Accessories available include a high frequency crystal probe and a high voltage probe which permits direct measurement of voltages up to 60kV.

EE 76 754 *for further details*

PRECISION GEARHEADS
Ancillary Developments Ltd, Blackwater Station Estate, Blackwater, Camberley, Surrey
(Illustrated above right)

A comprehensive new range of small precision gearheads and speed-reducers for high-grade electro-mechanical servo applications is shortly to go into production at Ancillary Developments Ltd. It includes what are believed to be the first British components of this type made in the 08 International Frame Size, plus four other sizes in common use. Although the entire range is designed to meet the relevant R.A.E. Specifications, these devices are not limited to aeronautical applications.

The materials used throughout are all corrosion-resistant and there is a choice of either plain or ball bearings to support the shafts. Adaptor plates can be

provided to match the gearhead or speed-reducer with varying sizes of servometer, and all units can be either spigot or clamp-mounted, as required.

Alternative forms of output shaft are also to be available, i.e., plain, pinion or splined and threaded.

Gear ratios up to 1000:1 are available in frame sizes 08, 10 and 11 and up to 5000:1 in sizes 15 and 18.

EE 76 755 *for further details*

SOLID STATE CONTROL METER
Kent Precision Electronics Ltd, Vale Road, Tonbridge, Kent
(Illustrated below)

Kent Precision Electronics Ltd has developed a solid-state calibrated control instrument for sensitive control, protection or alarm systems, with operation from a.c. or d.c. inputs over a wide range for solid-state switching of loads up to the kilowatt range, or alternatively, operation of an electromagnetic relay with isolated changeover contacts being made available for versatile operation. These instruments replace the conventional 'contact meter' type of control in all applications and since modern solid-state circuits are used throughout, an important feature is reliability of operation and long maintenance-free life.

The instrument has a linear scale calibrated in terms of the parameter to be controlled, and single or dual pointers which are set at the positions at which control is to be effected. The output power switch or relay contacts change state when the input parameter crosses these set points.

The instrument consists basically of a transistor low-drift d.c. amplifier which acts as a balance detector between a controlled reference voltage and the input parameter. The output of this amplifier operates a trigger circuit which in turn controls a thyristor output circuit


to switch the load, with either proportional or on-off control. When the controlled power is above the maximum available thyristor ratings or where isolated switching is required, a low power thyristor circuit is used to operate an electromagnetic relay or contactor with suitable ratings.

The instruments normally contain standard relays with 250V a.c. 5A contact ratings, with alternative ratings available on request. Standard thyristor outputs at ratings of up to 600V and 100A are available on specification.

Operation is from an auxiliary mains supply via an isolation transformer, with regulation for variations of ±20 per cent. The temperature coefficient is within 0·05 per cent/°C for all ranges and the instrument is suitable for operation in ambient temperatures of up to 85°C.

EE 76 756 *for further details*

STUDENTS' POTENTIOMETER
Croydon Precision Co., Hampton Road, Croydon, Surrey
(Illustrated above)

The students' potentiometer type P.1 is designed for use in technical colleges and grammar schools to supersede or follow on the simple banjo type of slide-wire potentiometer. The construction is robust and the accuracy is such that it can be used for serious work with a guarantee of repeatable results.

The voltage measuring range covered is 0 to 1·1V, the smallest subdivision being 0·001V. Two measuring dials are incorporated, the first dial having 10 steps of 0·1V and the second dial is in the form of a slidewire of 0·1V full scale sub-divided to 0·001V. The accuracy of measurement attainable is 0·1 per cent or 0·002V, whichever is the greater. A standardizing circuit is incorporated to enable the potentiometer to be standardized irrespective of the dial settings; and the potentiometer current is adjusted by means of a 'Cropico' constant contact resistance multi-turn rheostat. All necessary switches are incorporated and terminals are provided for connecting the standard cell, galvanometer, 1½V supply battery and the voltage to be measured.

One of the main features of this potentiometer is the graphic top panel which clearly indicates the circuit wiring and resistance networks. This has been clearly laid out to make it easy for the student to understand the function of the various controls, etc., and consequently also reduces the time necessary for the lecturer
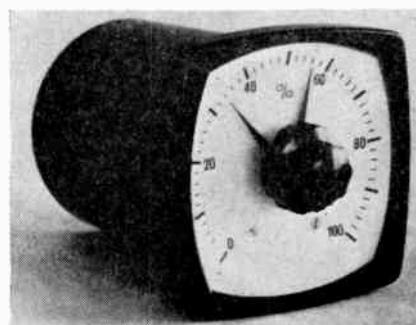
to explain the operation and function of the instrument.

A saturated acid Weston Standard Cell having a voltage of 1·01859V absolute at 20°C is provided as an accessory. This cell is contained in a rugged bakelite moulding with insulated terminals.

For the detector a portable moving-coil galvanometer with a sloping front panel is available. The sensitivity of 2μA per scale division is adequate whether the potentiometer is being standardized or used under short-circuit conditions.

EE 76 757 *for further details*

INERTIALLY DAMPED MOTORS
Bowmar Instrument Ltd, Sutherland Road, London, E.17
(Illustrated below)

New types of size 08 inertially damped motors for 400c/s operation are available from Bowmar. Reference voltages of 115V or 26V and control voltages to customer requirements are available. Type IM83-002 has a control winding of
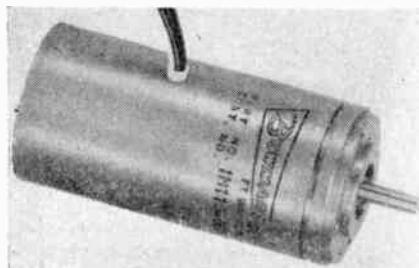
33V split phase, stall torque of 23·5g. cm, and a load speed of 6 500rev/min. Fly-wheel inertia is 4·6g. cm² with damping of 245dynes/cm/sec.

IM82-001 has a 33V centre-tapped control winding, stall torque of 21g. cm., and a no load speed of 5 500rev/min. Fly-wheel inertia is 1·6g. cm² with damping of 30dynes/cm./sec.

Both units have 12in lead out wires and 13 tooth 120 DP splined output shafts, and will operate in an ambient temperature of −65°C to +125°C. These motors can be supplied with gearheads with ratios up to 2 000 : 1. Size 11 units are also available with similar winding voltages and damping factors.

EE 76 758 *for further details*

PRINTED CIRCUIT DRILLING MACHINE
Drill Service (Horley) Ltd, 89 Albert Road, Horley, Surrey

In view of the recurring demand for unit drilling heads for use on special purpose printed circuit board machines, Drill Service now offer a ready made special purpose machine.

The method adopted is the well tried and familiar one of drilling stacks of boards from underneath while locating them from a template above them.

The boards to be drilled are stacked in a jig with a template on the top side. The boards and jig are then moved under

a stylus which locates in the template holes. The stylus is spring loaded and has a constant light pressure on the template to aid location. As soon as a location is made, the operator actuates the foot button to start the automatic drilling cycle.

On pressing the button, air pressure is applied to the drill feed mechanism and also to the locating head which is clamped firmly down on to the boards compressing them and locking them in the drilling position.

When the drilling cycle is completed the drill rapidly retracts and the clamping pressure is released for relocation.

The operator has both hands free for locating the template. A track on the template can increase the speed of the cycle since the lightly spring loaded stylus will follow the track from hole to hole.

The drilling head is the model 111 with the following features:—

> Infinitely variable speed up to 15 000rev/min.
>
> Infinitely variable feed rate.
>
> Both the feed and the speed are governed to ensure clean holes.

A drill cleaning timer arrangement allows the drill to retract and clear swarf at regular intervals to prevent clogging at high feed rates on large stacks of board.

The drill head is accurately aligned with the bush and stylus and stacks of 6 to 8 boards can be drilled very rapidly by unskilled operators using solid carbide drills without fear of accidental breakage.

This drilling unit is particularly suitable for use with extra length solid tungsten carbide circuit board drills. These drills will maintain their edge and size for extended periods, saving set-up time and unnecessary scrap of boards and components on assembly.

EE 76 759 *for further details*

HUMIDITY METER
The Wayne Kerr Laboratories Ltd, Sycamore Grove, New Malden, Surrey
(Illustrated below)

First in a new range of compact

measuring instruments by Wayne Kerr —to be known as the 'Midge' series— is humidity meter MB100. This self-contained unit requires no setting-up or calibration and is ready for immediate use at all times. It operates in conjunction with a low-cost transducer to provide direct readings of relative humidities from 20 to 100 per cent, in two ranges. Sealed packages can be so designed that each includes a transducer element with leads to a connector on the outer case, enabling the condition of equipment stored or shipped under humid conditions to be checked instantly without any unpacking.

Overall measurement accuracy, including the transducer, is 5 per cent. Transducers will withstand temperature variations between −40°C and +100°C and their performance is unimpaired by prolonged storage or use under extreme conditions of relative humidity. Essentially the measurement circuit is a transformer ratio-arm bridge operating at 2kc/s and balanced electronically by a high-gain feedback amplifier. Source, detector and battery are all housed in the case (4¼ × 2¼ × 6½in) and an output of 0 to 100μA is available for recorders, etc.

EE 76 760 *for further details*

METAL OXIDE SILICON TRANSISTOR
Mullard Ltd, Mullard House, Torrington Place, London, W.C.1

Announcing a new metal oxide silicon transistor (m.o.s.t.), Mullard Ltd state that produced in quantity it would be many times cheaper than any device of similar performance now available in Britain. Large scale production is planned; meanwhile the device can be supplied in sample quantities. High reliability is ensured by the use of well tried planar techniques.

The m.o.s.t., development type 95BFY, has an input resistance of a million megohms ($10^{12}\Omega$) — far higher than that of a conventional transistor or even a field effect transistor—and could, for example, be used as a high-to-low impedance convertor or a low-frequency amplifier in electronic instrumentation and control circuits, computers and tele-communications equipment. It is also suitable for use in logic circuits.

Since the m.o.s.t. is a majority carrier device it is unaffected by radiation and will therefore also find application in satellites and other vehicles sent into outer space.

Initial applications work shows that the device will operate as an oscillator at frequencies up to 150Mc/s, as a mixer giving a 20dB conversion gain (the substrate being used as the second signal terminal), or as an 'anode' modulated output stage for powers up to several hundred milliwatts at frequencies up to 100Mc/s. Its performance as a chopper is particularly impressive since the de-
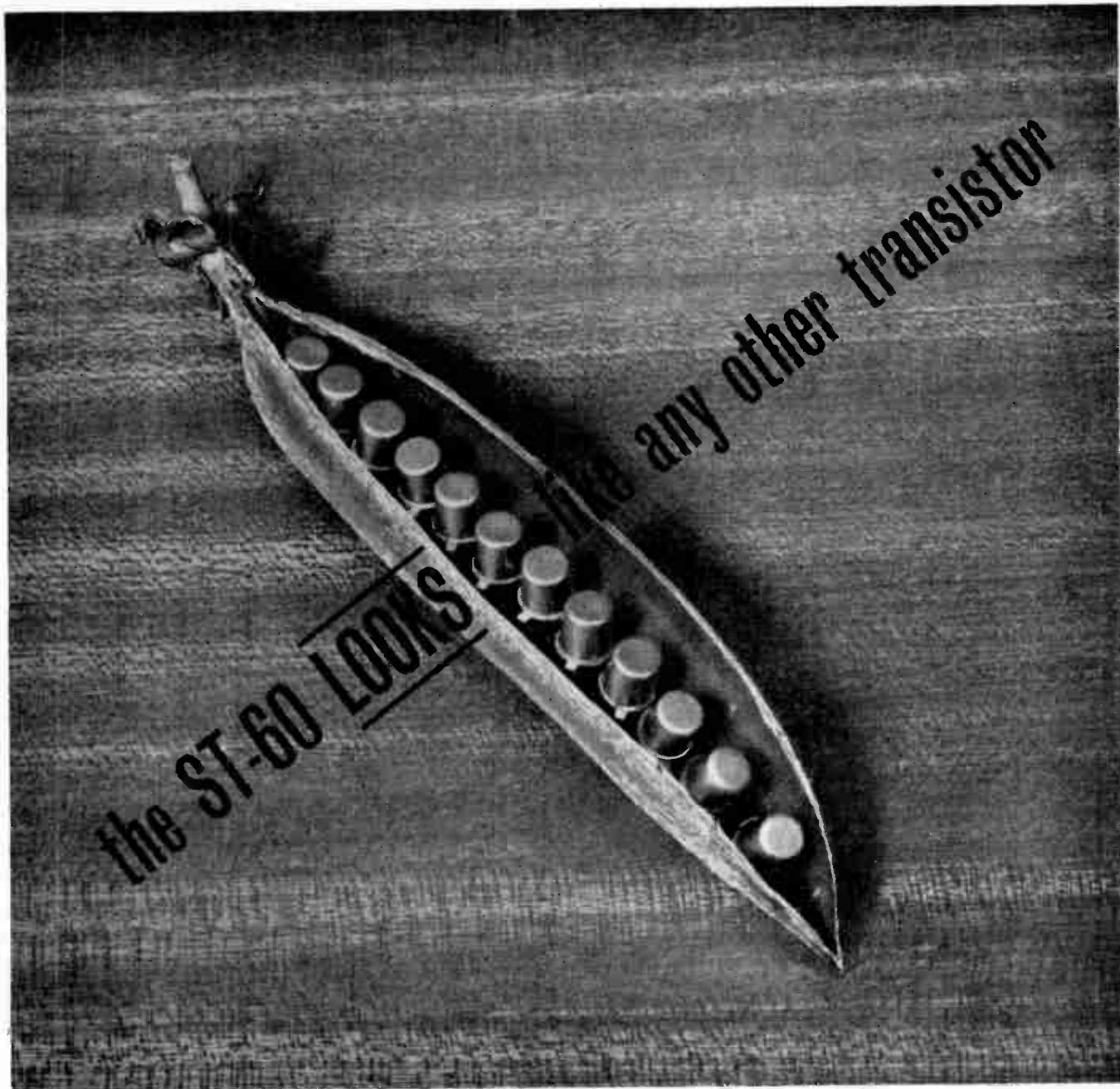
NEW FROM THORN
low cost

220-250v
MAINS OPERATED
SIGNAL LAMPHOLDERS

Thorn now offer a range of Signal Lampholders, for mains operation, at exceptionally low cost. The lampholders cover a wide variety of uses including: Control and alarm enunciators for remote control or visual supervision in switch gear, automated processes, electro-mechanical apparatus etc.

They will stand up well to rough handling and are stable in ambient temperatures up to 90°C. Seven distinctive lens colours are available in various styles to suit the majority of the lampholder bases. Proven economical in use, and unaffected by vibration or frequency of switching, the average life expectation of the lamps is in the region of 5-10,000 hours.

THORN SUPPLY THE LAMPS TOO

Three lamps, available in Red or Green Neon, to fit the lampholders with the exception of type 0216

1

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THORN SIGNAL BLINKER LAMP

A solid state device, unaffected by ambient temperature, giving instantaneous operation for critical installations where a flashing signal is vital. A.C. or D.C. Flash frequency can be regulated from 20 to 120 pulses per minute. Bayonet or screw cap versions available with lens colours of red, green, blue, amber, clear or opal.

This filament lamp is specially designed for lampholder type 0216

THE BEST COST LESS FROM THORN

a brief description of types available

1 Type 0216. Large lampholder base, three alternative lenses, available with black, white or polished nickel bezel.

2 Type 0171. Long lampholder base with two alternative lenses.

3 Type 0172. Short version of type 0171 with three alternative lenses.

4 Type 0195. Medium sized lampholder, three alternative lenses.

5 Type 0228. Small lampholder with integral bezel and choice of two plastic lenses.

6 Type 0176. Large, flush-mounting lampholder with either black or white base and one type of lens only.

7 Type 0239. Short style lampholder with one high moulded plastic lens.

8 Type 0231. New style press switch. Second circuit is completed when illuminated cap is pressed.

THORN ELECTRICAL INDUSTRIES LTD.,
SPECIAL PRODUCTS DIVISION,
105-109 Judd Street, London, W.C.1.

Please send me full details of the Thorn (Mains operated) Signal Lampholders.

Name...

Position..

Company ....

Address...

....

....

DTV/22

vice exhibits no 'off-set' voltage and an 'on' resistance of about 400Ω at chopping frequencies up to 100kc/s. Furthermore it exhibits no storage effect and thus when used in switching applications, switching time is solely dependent on the *RC* time-constant of the gate capacitance charging through the channel resistance.

The 95BFY is produced by diffusing two closely spaced n+ regions (source and drain) on a p-type silicon substrate. Between the source and drain, a thin layer of silicon oxide is grown on the substrate and an aluminium gate electrode evaporated on to the top side of the oxide layer.

When a positive potential is applied to the gate, a negative charge is induced at the substrate surface causing an n-type channel to form a conducting path between source and drain. If a potential is now applied between source and drain, a current flows which can be controlled by the potential applied to the gate.

EE 76 761 *for further details*

VIBRATION GENERATING EQUIPMENT
Derritron Electronic Vibrations Ltd,
24 Upper Brook Street, London, W.1
(Illustrated below)

This 6 000 lb thrust vibration system consists of a 30kVA power amplifier, designed for both random or sine testing, and a 6 000 lb thrust vibrator plus control equipment. The amplifier has a maximum continuous anode dissipation of 50kW, and has a full power frequency range of 15c/s to 5kc/s, and full current is available down to 5c/s. The latest

vapour cooled English Electric valves are employed, and hum and noise is at least 70dB down with reference to full output.

The 6 000 lb thrust vibrator (type number VP.600) employs a drive coil cooling system which provides an extremely stiff table and coil assembly. Distilled water cooling is normally employed, although for certain extreme climatic chamber conditions, oil cooling could be used. Brief specification for this vibrator is 6 000 lb vector thrust, frequency range approximately 5c/s to 3kc/s, and usable up to 5kc/s, with a first major disturbing resonance at 2·5kc/s. The low table weight of .70 lb allows an unloaded table acceleration of over 85g.

EE 76 762 *for further details*

VOLTAGE SENSING RELAYS
Newton Brothers (Derby) Ltd, Alfreton Road, Derby
(Illustrated above)

Newton Derby TVS voltage sensing relays have been designed to provide a means of switching at a pre-determined voltage level and can be used in under and over-voltage protection circuits where high accuracy is required.

The range covers sensing voltage between 10 to 220V d.c. sensitivity being ± 1 per cent and differential 1 per cent. The speed of response is 10 to 15msec but where it is required to prevent tripping by overloads under transient conditions a time delay from a few milliseconds up to 2 minutes can be provided.

These units are fitted with ten pin plugs and can be readily incorporated in control cubicles. Each is fitted with two normally closed and one normally open heavy duty contacts.

**EE 76 763** *for further details*

BRUSHLESS D.C. SERVO MOTORS
Vactric Control Equipment Ltd, Garth Road, Morden, Surrey
(Illustrated below)

This newly developed brushless d.c. servo motor in frame size 11 has been designed to meet the need for a unit requiring a much longer life than the conventional type of d.c. motor and it is particularly suitable for use where radio noise suppression is of importance. It is devoid of any brushes or commutator and the absence of arcing and brush wear account for the longer life. It is specially wound to accept waveforms from a separate electronic unit.

The motor has been constructed from corrosion resistant material throughout and is capable of operating in the ambient temperature range —65°C to +85°C at altitudes up to 60 000ft.

The electronic unit contains a multi-

vibrator which controls a series of flip-flops through suitable logic circuits, the flip-flops having power transistors which supply the motor windings. It is approximately two cubic inches in volume and can withstand the same environmental conditions as the motor. It can be supplied in encapsulated form in a size 11 cylindrical case with solder spills; altentatively, a plug-in circuit card can be provided for operation in less stringent circumstances.

EE 76 764 *for further details*

COMPONENT BRIDGE
K.L.B. Electric Ltd, 335 Whitehorse Road, Croydon, Surrey
(Illustrated below)

The M.300 component bridge is a versatile, low cost unit having four capaci-

tance ranges from 10pF to 1 000μF and four resistance ranges from 0·05Ω to 100MΩ, it also has facilities for capacitor leakage tests with an adjustable test voltage up to 500V d.c. Ratio tests in the ranges of 0·05 to 1 and 20 to 1 can be carried out between any two capacitors, inductors or resistors. This test is intended for use when components have to be adjusted against a standard.

EE 76 765 *for further details*

SWITCHABLE CIRCULATOR
The Marconi Co. Ltd, Chelmsford, Essex
(Illustrated on page 858)

The Marconi Company has developed a new, three-port switchable circulator in waveguide size No. 15, covering the frequency range 7·75 to 8·5Gc/s. High speed switching of the direction of circulation is achieved electromagnetically.

This new circulator has been designed for use with microwave communications equipment, and offers considerable improvements in reliability, speed of operation and space saving, over previous switching methods that employed mechanical devices. It is suitable for all two-way waveguide switching applications, a typical example being the branching networks connecting microwave transmitters and receivers with their aerials.

This switchable circulator is a version of the standard three-port junction circulator but the magnetic biasing field is supplied electromagnetically and is reversed to change the direction of circulation. The electromagnet and circulator
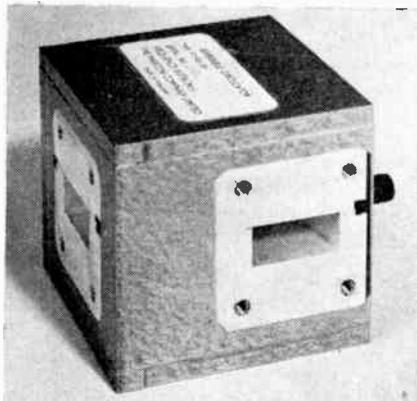
body are specially constructed to allow the magnetic field to be rapidly reversed by a normal transistor switching circuit. The switching time of the circulator is governed by the power supplied to the magnet. Switching times of less than 0·5msec may be achieved with a 25W supply.

EE 76 766 for further details

TELEVISION SIGNAL STRENGTH METER

Lab-Craft Ltd, Gainsborough Road, Woodford Bridge, Essex

(*Illustrated below*)

The type 215 v.h.f. signal strength meter is engineered to high standards using modern techniques. Frame grid valves are used in all amplifying stages and the stabilized h.t. reduces drift and gain variations to negligible proportions over the whole range of mains voltage.

The instrument is designed around a standard v.h.f. tuner working into a gain-controlled i.f. amplifier. The controlled current of one of the i.f. amplifier valves is read on the meter. The scale is calibrated directly in microvolts and millivolts giving clearly readable calibrations from fringe to swamp levels without the need for separate plug-in attenuators. Increasing signal levels produce a decrease in the current through the meter. This arrangement affords automatic protection to the meter against accidental overload.

The fine-tuner is set once for each channel and thereafter has a 'memory' adjusting automatically each time the turret is turned. This coupled with the non-sequential arrangement of coils in the turret makes the changeover from Band 1 to Band 3 instantaneous in most areas.

Highly sensitive with readings down

to 10μV, the model 215 can be used in extreme fringe locations where the need for an instrument is greatest. The aerial socket is located on the rear of the instrument. By using a short telescopic aerial relative assessment of aerial gain requirements can be made on any site.

EE 76 767 for further details

SELF-LATCHING CONNECTORS

Distributed by: Nutec Electronics Ltd, Mercantor House, East Street, Shoreham-by-Sea, Sussex

(*Illustrated below*)

Lemo Connectors, designed by Swiss Engineer Leon Mouttet, incorporate a patent latching system which obviates accidental disengagement by vibration or cable pulling of the plug and socket.

The contact surfaces are heavily gold-plated to ensure very low and constant contact resistance and only the highest quality insulating material is employed.
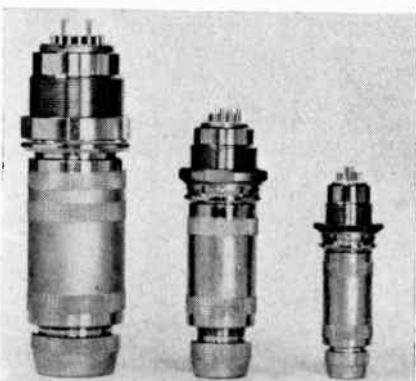
There are three fundamental types of Lemo connectors—coaxial, biaxial, triaxial high voltage or multi-pin. Each is available in seven sizes all of which can be combined in one assembly to give multiple variations.

In addition to the standard models and combinations, there are high voltage types: water tight and vacuum sealed connectors: special versions for high and low voltage, together with coaxial plugs and systems for television.

The various collett/clamp assemblies can also accommodate a wide range of cables.

EE 76 768 for further details

SUBMINIATURE TRANSDUCERS

Associated Engineering Ltd, Cawston, Rugby

(*Illustrated above right*)

A new type of subminiature inductive displacement transducer has been developed for displacement and vibration measurements in difficult environmental situations.

The transducer, which is made in the two sizes shown, may be used with carrier frequencies up to 300kc/s, so permitting the detection of vibration or displacement of any metallic material at frequencies from d.c. to 60kc/s. The two transducers have a working range of 0·010in, will withstand ambient temperatures up to 150°C with no appreciable change in sensitivity and when suitably

mounted can be subjected to accelerations up to 2 000g without malfunction.

Transducer size and weight, including printed circuit termination:

Large size: length 0·430in, diameter 0·250in, weight 0·7g.

Small size: length 0·400in, diameter 0·160in, weight 0·5g.

The transducers can be supplied with a threaded mounting ring as shown on the left in the photograph, if required. Otherwise, they are suplied as shown centre and right, when they may be mounted by clamping or by the use of a suitable cement or epoxy resin.

EE 76 769 for further details

SOLDERING IRONS

Distributed by: Lugton & Co. Ltd, 209-212 Tottenham Court Road, London, W.1

Imperial Ungar soldering irons, which are manufactured in Canada, are now available from Lugton & Co. Ltd.

The main feature of this soldering iron is its adaptability as all parts are rapidly interchangeable, so that the iron can be converted from say a 25W iron with a micro chisel bit to a 40W iron with a long taper chisel bit in a matter of minutes. A large selection of bits is available.

The heat cartridge sealed in stainless steel ensures longer life, eliminates oxydation and provides maximum heat transfer. Different colours allow two irons of different wattage to be distinguished on sight.

EE 76 770 for further details

WAVEFORM GENERATOR

Feedback Ltd, Crowborough, Sussex

(*Illustrated on page 859*)

The waveform generator type TWG 300 is a versatile instrument intended for investigations into control system performance.

The operation of the TWG 300 is based on an integration and switching circuit which produces triangular and square waves. The triangular waveform is shaped in a diode network to give a sine wave which is constant in amplitude and has a low harmonic content over the entire frequency range. A further network generates the clipped triangular wave producing alternate 'flats' and 'ramps' by removing the peaks from the triangular waveform. The clipping circuit is adjustable enabling the slope of the ramp to be varied but retaining the amplitude constant, so providing an adjustable ramp : rest ratio.

The frequency is from 0·001 to 1000c/s in six 10:1 switched ranges with continuous adjustment in each range. The calibration is in both frequency and period and there is a 20 per cent over-

lap at the ends of ranges ensuring ample continuity of cover. The scale accuracy is better than ±5 per cent. The hour to hour stability is typically better than 0·1 per cent. For frequency setting against counters or frequency standards, the discrimination is better than 0·1 per cent of the frequency. The selected waveform is available simultaneously on two outputs: either a constant level of 40V peak at a maximum current of 20mA, output impedance 100Ω or via a 2kΩ continuously adjustable attenuator calibrated 0 to 40V and a series switch giving ×1 - ×0·1—off conditions. Output amplitude is constant irrespective of the frequency setting. Outputs are taken between 'live' and 'common' terminals. The earth or 'ground' terminal, connected to the instrument frame, is normally linked to the 'common.' On removing the link external supplies may be connected to 'float' the instrument up to 250V from earth potential. Precautions must be taken to provide the external supply with a low impedance path for possible hum pick-up.

Monitor outputs, square and triangle have constant amplitude of 5V peak and a source impedance of 20kΩ. These outputs are synchronized with the main output waveform.

EE 76 771 *for further details*

TIME-BASE AND COUNTER
Specto Avionics Ltd, Feltham, Middlesex
(Illustrated above right)

A universal time measuring instrument, for the measurement of frequency, period, time and pulse width, has been developed by Specto Avionics Ltd. This instrument is designed for measurements that require extremely high accuracy and will be particularly useful in research establishments, universities and industry.

The instrument combines two standard Specto Avionics units, the deci-microsecond chronometer type 222 and the time-base unit type 233. The accuracy of these combined instruments is ±2 parts in 10^7 long term; ±1 part in 10^8 short term. In common with all instruments of the range, time-base type 233 uses silicon transistors, the latest printed circuit techniques and is of light weight

modular construction and can be rack mounted or used independently on the bench. The size of the time-base is 17¼ × 3¼ × 14½in, when racked with the deci-microsecond chronometer the combined height is 8¼in.

The brief Specification of the combined instruments is as follows:

Time and pulse width measurement—0·1sec to 10^6sec.

Inputs direct—±1V to ±5V peak pulse at 1kΩ load.

Inputs via attenuator—±5V to ±25V peak pulse at 1kΩ load.

Frequency—30c/s to 20Mc/s.

Period—1 event to 10^8 event minimum period 0·05μsec.

Ratio—f_1, 20Mc/s max. f_2, 15Mc/s max.

Input sensitivity—100mV to 5V r.m.s.

Input impedance—10kΩ.

Temperature range——20°C to +40°C.

EE 76 772 *for further details*

BRUSHLESS SYNCHROS
Elliott-Automation Ltd, Century Works, Lewisham, London, S.E.13
(Illustrated below)

A new brushless synchro for applications which require the use of high reliability components is announced by the Servo Components Division of Elliott-Automation.

The new Elliott Ellsyn brushless synchros are available as size 11 control transmitters and transformers. They possess many advantages over conventional components, but retain similar electrical and mechanical characteristics. Brushes and slip rings have been replaced by miniature injection transformers thereby eliminating brush contact problems and reducing stiction to approximately a quarter of the level found in conventional synchros. These new components are therefore particularly suitable for flight instrument applications.

The units are suitable for operation in temperatures up to +125°C with maximum errors of ±14 minutes of arc. Work is now proceeding to improve this performance and to permit operation in ambient temperatures up to +150°C.

By careful design it has been possible to keep the length of the control transformer the same as that of conventional versions. The control transmitter, which has to handle more power, is 0·5in longer than a standard synchro.

Both the 115V and 26V brushless synchros in the current Ellsyn range have been designed to be interchangeable with conventional units.

EE 76 773 *for further details*

IMAGE ORTHICONS
English Electric Valve Co. Ltd, Chelmsford, Essex
(Illustrated below)

The English Electric Valve Co. Ltd has announced a new development in the target material for its image orthicons.

This new material, known as ELCON, gives a new high of at least 750 hours in the guaranteed life of image orthicons. Reports on pre-production samples
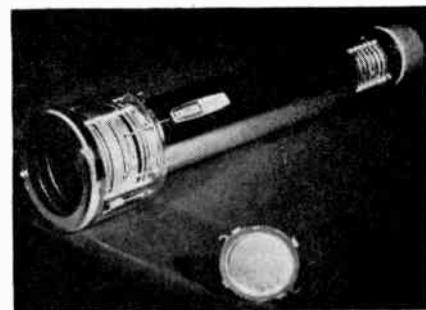
indicate operational lives as long as 3 000 to 5 000 hours.

The ELCON target also achieves complete lack of image retention, stability of contrast reproduction, sensitivity fall-off greatly reduced and immediate switch-on.

The ELCON target introduces a new principle of operation where electronic, rather than ionic, conduction is employed in transferring the charge pattern from the image side to the beam side. Ion migration is therefore virtually eliminated and in consequence there is no significant deterioration in performance with time. This means that sensitivity and stability of contrast reproduction (gamma) remains constant and there is complete freedom from 'sticking' for life.

EEV maintain that the new material has no hidden vices and that it provides a highly satisfactory answer to the problems experienced with any normal type of image orthicon.

The full EEV range of 3in image orthicons features the ELCON target and these are now in current production.

Sample quantities of the EEV 4½in tube, also incorporating this development, will be ready shortly and full scale production will commence early in 1965.

The name ELCON has been derived from the properties of the new target material, namely ELectronic CONducting, as opposed to the familiar ionic conductors normally used.

EE 76 774 *for further details*

MEETINGS THIS MONTH

THE INSTITUTION OF ELECTRONIC AND RADIO ENGINEERS

All meetings at the London School of Hygiene and Tropical Medicine, Keppel Street, Gower Street, London, W.C.1, unless otherwise stated.

Electro-acoustics Group
Date: 2 December   Time: 6 p.m.
Symposium: Special Loudspeaker Applications

Joint I.E.R.E.-I.E.E.
Medical Electronics Group
Date: 16 December   Time: 6 p.m.
Discussion: D. C. Defibrillation
Opened by: J. Spencer

South Wales Section
Date: 9 December   Time: 6.30 p.m.
Held at: The Welsh College of Advanced Technology, Cardiff
Lecture: The Industrial Applications of Digital Techniques
By: F. Shaw

West Midland Section
Date: 10 December   Time: 6 p.m.
Held at: University of Birmingham, Electrical Engineering Department, Birmingham
Lecture: Computers in Biochemical Research
By: W. T. Williams

Southern Section
Date: 15 December   Time: 6.30 p.m.
Held at: Brighton College of Technology, Brighton
Lecture: Application of Telemetry and Remote Control Systems to a Water Undertaking
By: G. Bishop
Date: 1 December   Time: 6.30 p.m.
Held at: The Lanchester Theatre of The University of Southampton, Southampton
Lecture: Modern Memory Systems
By: W. Renwick

Scottish Section
Date: 8 December   Time: 6 p.m.
Held at: Carlton Hotel, North Bridge, Edinburgh
Lecture: U.H.F. Aerials
(Joint meetings with the Scottish Centre, Electronics and Measurements Section, I.E.E.)
By: C. F. Whitbread
Date: 9 December   Time: 7 p.m.
Held at: The Department of Natural Philosophy, The University, Drummond Street, Edinburgh
Lecture: Transistor Audio Amplifiers
By: P. J. Baxandall
Date: 7 December   Time: 6 p.m.
Held at: The University of Strathclyde, Glasgow
Lecture: U.H.F. Aerials
(Joint meeting with the Scottish Centre, Electronics and Measurement Section, I.E.E.)
By: C. F. Whitbread
Date: 10 December   Time: 7 p.m.
Held at: The Institution of Engineers and Shipbuilders, 39 Elmbank Crescent, Glasgow
Lecture: Transistor Audio Amplifiers
By: P. J. Baxandall

Yorkshire Section
Date: 2 December   Time: 6.30 p.m.
Held at: Leeds University, Electrical Engineering Department, Leeds
Lecture: Computer Input/Output Equipment
By: M. P. Circuit

East Midland Section
Date: 9 December   Time: 6.30 p.m.
Held at: University of Leicester, Leicester
Lecture: Survey of Radio Astronomy
By: J. Heywood

Merseyside Section
Date: 16 December   Time: 7.30 p.m
Held at: The Walker Art Gallery, Liverpool
Lecture: Medical Electronics
By: W. H. Wilkes

North Western Section
Date: 10 December   Time: 6 p.m.
Held at: Renold. Building, Manchester College of Science and Technology, Manchester
Lecture: Applied Electronics in Medicine
By: J. F. Cordell

South Midland Section
Date: 17 December   Time: 7 p.m.
Held at: The Winter Gardens, Malvern
Lecture: Small High Fidelity Loudspeaker Systems
By: K. F. Russell

North Eastern Section
Date: 9 December   Time: 7 p.m.
Held at: The Institute of Mining and Mechanical Engineers, Westgate Road, Newcastle upon Tyne
Lecture: Instrumentation for Ultrasonic Diagnosis and Surgery
By: D. Gordon

THE INSTITUTION OF ELECTRICAL ENGINEERS

Unless otherwise stated, all meetings will be held at Savoy Place, commencing at 5.30 p.m.

Electronics Division
Date: 2 December
Lecture: The U.K. 3 Satellite
By: H. J. Sketch
Date: 7 December   Time: 10 a.m. and 2.30 a.m.
Colloquim: Programmed Test Equipment and Techniques
Date: 9 December
Lecture: Telemetry—The Present Position and Future Trends
By: R. E. Young
Date: 16 December   Time: 6 p.m.
Discussion: D.C. Defibrillation
(Joint meeting with the I.E.R.E. Medical and Biological Group at the London School of Hygiene and Tropical Medicine)
Opened by: J. Spencer
Date: 17 December   Time: 10 a.m., 2.30 p.m. and 5.30 p.m.
Colloquium: Modern Techniques for Microwave Circuit Synthesis

Power Division
Date: 2 December   Time: 3.30 p.m. and 5 p.m.
Lecture: Survey of Basic Stray Losses in Squirrel Cage Induction Motors
By: K. K. Schwarz
Lecture: Stray Losses in Squirrel Cage Induction Motors; The Validity of the Reverse Rotation Test Method
By: B. J. Chalmers and A. C. Williamson
Lecture: Measurement of Stray Load Losses in Squirrel Cage Induction Motors
By: B. M. Bird
Date: 10 December
Lecture: Resistive Earthing of Electrical Distribution Systems
By: D. Jones
Date: 16 December
Lecture: Some Economic Aspects of Extra-High-Voltage Underground Cables
By: H. K. Beale
Lecture: Gas Pressurized Lapped Polythene Dielectric for Extra-High-Voltage Power Cable Systems
By: J. A. M. Gibbons, P. R. Howard and D. J. Skipper
Lecture: Impulse Strength Measurements on Compressed Gas Insulation for Extra-High-Voltage Power Cables
By: D. J. Skipper and P. I. McNeall

Science and General Division
Date: 1 December
Lecture: Optimal Control of a Boiler
(Joint meeting with the I.Mech.E. Automatic Control Group)
By: H. Nicholson
Date: 7 December
Lecture: Non-Linear Circuit Theory
By: J. C. West
Date: 9 December
Discussion: The Electrometer Amplifier: its Design and Applications
Opened by: G. I. Hitchcox and G. R. Taylor
Date: 15 December
Discussion: Transistors—The First Encounter
Opened by: V. H. Attree

THE RADAR AND ELECTRONICS ASSOCIATION

Date: 10 December   Time: 7 p.m.
Held at: The Royal Society of Arts. John Adam Street, Adelphi, London, W.C.2.
Lecture: Pay Television Systems
By: K. A. Russell

THE TELEVISION SOCIETY

Date: 4 December   Time: 7 p.m.
Held at: The Conference Suite, I.T.A., 70 Brompton Road, London, S.W.3
Lecture: Television University
By: Weltman
Date: 10 December
Discussion: Television Servicing—Is it Satisfactory?

PUBLICATIONS RECEIVED

INDUSTRIAL ELECTRONIC COMPONENTS by A. C. Farnell Ltd are described in a recent catalogue which includes a supplementary transistor and thermionic valve catalogue and price list and also Hunts Capacitor stock list and price list. Copies of this information are available on request to the Director, A. C. Farnall Ltd, Hereford House, North Court, Vicar Lane, Leeds 2.

THE ELECTRONIC ENGINEERING ASSOCIATION has produced a ' Guide for the Quality of Copper Clad Epoxide Bonded Glass Fabric Laminate, suitable for Precious Metal Plating ', the latest addition to the EEA Printed Wiring Guides. This has been prepared in an attempt to cover a subject on which little or no positive information has been previously published, and is intended for the guidance of both manufacturers and users. Copies of the Guide are obtainable, free of charge, from the Information Officer, Electronic Engineering Association, 61 Green Street, London, W.1.

MULLARD SEMICONDUCTOR DESIGNERS' GUIDE is now available from the company. The Guide contains quick-find charts for transistors which list the devices under the main headings of collector voltage, total dissipation and cut-off frequency. Full dimensional drawings and details of the international encapsulation outlines to which the devices comply are given at the back of the booklet. Request for copies should be made on company headed notepaper to the following address: Technical Office, Industrial Markets Division, Mullard Limited, Mullard House, Torrington Place, London, W.C.1.

PHOENIX QUICK-FIX TERMINALS is the subject of a catalogue which is available from D.J. Equipment (Hersham) Ltd, 39 Primrose Road, Queens Road, Hersham, Walton-on-Thames, Surrey. The catalogue has a comprehensive price list, illustrated details of accessories, typical applications and technical information.

MAGNETS FOR REED SWITCHES describes a considerable programme of research and development work which has been carried out by Marrison and Catherall Ltd, Forge Lane, Killamarsh, nr. Sheffield, into the design of permanent magnets for use with reed switches. This has been summarized in a booklet which lists 19 different types of magnet and gives their operating distances with a range of switches of differing size and sensitivity. The booklet with price list is available from the company on request.

SIMA REGISTER OF MEDICAL INSTRUMENT MANUFACTURERS. One of the aims of the SIMA Medical Instruments Group is to provide channels of communication between members and Government departments, customers and users of medical instruments. To assist in this, a Register of Information has been prepared to give details of the general activities and the equipment and services in the medical field, of certain member companies of this group. This register contains information from 48 companies and further details of specific medical instruments may be obtained on request to Scientific Manufacturers' Association. Sima House, 20 Peel Street, London, W.8.

ÉQUIPEMENT ÉLECTRONIQUE

Une description basée sur des renseignements fournis par les fabricants de nouveaux organes, accessoires et instruments d'essai

Traduction des pages 854 à 859

APPAREIL DE CONCENTRATION DE DONNÉES

Cambridge Instrument Co. Ltd.
13 Grosvenor Place, London, S.W.1
(Illustration à la page 854)

L'équipement de concentration de données Cambridge-Allscrott, qui a été mis au point en collaboration avec la société James Scott (Electronic Engineering) Ltd, indique et enregistre sous forme numérique la valeur mesurée de n'importe quel nombre de canaux dans un programme séquentiel, ainsi que le temps de mesure et un numéro d'identification. Il comporte également un système d'alarme réglable qui donne l'alerte de manière visible lorsque la valeur mesurée de n'importe quelle voie ou groupe de voies se déplace au dehors des limites préréglées.

L'équipement mesure et enregistre toute variable pouvant être convertie en un potentiel électrique approprié ou en un changement de résistance. Des entrées mixtes peuvent être réalisées à l'aide d'éléments modulaires appropriés et d'éléments de mesure primaires. Les entrées de canaux sont reliées en séquence à un instrument de mesure de rééquilibrage d'asservissement Cambridge, au moyen d'éléments d'exploration Allscott contrôlés par un appareil cyclique automatique Allscott. Un encodeur à arbre binaire, monté sur l'instrument de mesure, fournit une sortie numérique traduite sous forme décimale et emmagasinée avec le numéro de la voie et le moment de mesure. Ce dispositif permet non seulement de commuter l'instrument de mesure sur la voie suivante, pendant que l'impression s'effectue, mais il assure également une lecture numérique du numéro de la voie et de la valeur mesurés. Les voies sont explorées automatiquement suivant un programme choisi à la vitesse d'un balayage toutes les trois secondes. La commande manuelle peut également être effectuée, en permettant le choix de la voie voulue par pression du bouton approprié sur un panneau annonciateur.

L'équipement comprend normalement sept points d'alerte "élevés" et sept points d'alerte "bas," qui sont tous réglables et assurent 48 combinaisons de limites d'alerte. Des niveaux d'alerte supplémentaires peuvent être fournis sur demande, et n'importe quel point réglé peut être relié à n'importe quelle voie unique ou groupe de voies au moyen d'un élément de sélection à répartiteur. L'instrument de mesure comporte un fil à contact glissant de transmission qui donne à l'élément de niveau d'alerte un signal de référence proportional à la valeur mesurée de la voie choisie. Si la valeur mesurée d'une voie se trouve en dehors des limites d'alerte, les contacts de commande sont excités et la lampe appropriée sur le panneau annonciateur d'alarme s'allume en rouge, demeurant allumée jusqu'à ce qu'elle soit éteinte à la main. La lecture et l'impression numérique sont affichées en roug lorsque la voie choisie se trouve et état d'alarme.

On peut, enfin, obtenir un enregistrement numérique du numéro de la voie, de la valeur mesurée et du moment de mesure, mais un mécanisme d'enregistrement à bande peut être également incorporé à l'instrument de mesure et fournir un enregistrement analogique des valeurs mesurées jusqu'à 100 voies.

EE 76 751 *pour plus amples renseignements*

MICROVOLTMÈTRE HF

Distributed by: Claude Lyons Ltd,
76 Old Hall Street, Liverpool 3
(Illustration à la page 854)

Le microvoltmètre HF Millivac typ MV-28B comprend huit gammes de tensions allant de 1 mV à 3 V sur la totalité de l'échelle dans la gamme de fréquences de 10 kHz à 1200 MHz, la tension la plus basse pouvant être lue étant de 300μV. La précision est de 3% de 100 kHz à 60 MHz, 5% de 50 kHz à 200 MHz et 10% pour le reste de la gamme. L'impédance d'entrée avec le bout de la sonde à haute impédance est d'environ 2 à 3 pF et 75 kΩ. Un adaptateur de terminaison BNC de 50 Ω est fourni comme accessoire normal, les autres accessoires

comprenant un diviseur de tension capacitive 100:1, des terminaisons N et GR et un adaptateur de shuntage type N 'T'.

L'instrument de mesure à échelle à miroir est étalonné en volts et en dBm (par rapport é 1 mW dans 50 Ω). La réponse est proche de la valeur efficace jusqu'à 30 mV et de la pointe jusqu'à 30 mV, et elle est étalonnée en termes équivalents à des ondes sinusoïdales efficaces. Il est à noter, en outre, qu'en employant le diviseur de tension capacitif 100:1, on peut porter jusqu'à 3 V son utilité en tant qu'instrument de réponse efficace.

EE 76 752 *pour plus amples renseignements*

ENREGISTREUR D'HUMIDITÉ

Kappa Electronics Ltd, 159 Hammersmith Road, London, W.6
(Illustration à la page 854)

L'enregistreur d'humidité à transistors, modèle AB66, est un instrument portatif fonctionnant sur batterie et conçu pour mesurer d'un seul côté la teneur en humidité de matériaux en feuilles ainsi que de matérieux ou de substances ayant une surface relativement plate permettant d'effectuer un contact physique avec la plaque sensible montée sur la base du logement de l'équipement.

Grâce à cette méthode, l'appareil mesure l'effet du matériau sur le champ électrostatique à haute fréquence rayonné à partir de la surface de la plaque sensible, l'effet étant dû aux propriétés électrochimiques du matériau dont, pour plusieurs substances, la teneur en humilité est un facteur majeur.

Selon la composition du matériau et sa teneur en humidité, l'instrument fournit, une indication pouvant être reliée soit à la constante diélectrique soit à l'impédance. Cette indication est fournie par un circuit de détection hétérodyne, la mesure étant indiquée sur un cadran fonctionnant à la main et étalonné en unités arbitraires.

Pour l'utilisation, l'instrument est placé sur l'une des surfaces du matériau et la commande d'accord étalonné est

réglée de manière à produire une déviation maxima de l'enregistreur du panneau. Le réglage de la commande d'accord est ensuite lu et converti en teneur d'humanité en se référant à un diagramme ou à une table préparés auparavant à l'aide d'échantillons de teneur en humidité connus.

La mesure sur des matériaux en feuilles relativement minces s'effectue soit en posant le matériau sur la surface plate du bac de base fourni avec l'instrument soit, si cela est plus pratique, en le plaçant sur une feuille plus large à dos métallique en polystyrène dilaté.

Selon le type de plaque sensible choisi et selon sa profondeur de pénétration, des matériaux relativement épais peuvent être contrôlés sur place à condition que l'on prenne soin de s'assurer que ces conditions puissent être répétées.

EE 76 753 *pour plus amples renseignements*

VOLTMÈTRE ÉLECTRONIQUE
K.L.B. Electric Ltd, 335 Whitehorse Road, Croydon, Surrey
(Illustration à la page 855)

Le voltmètre électronique PACO, type V.70, est un instrument à bas prix et d'une grande souplesse d'emploi, comportant sept gammes de tension continue de 0 à 1500 V efficaces et de 0 à 4000 V crête à crête, ainsi qu'une gamme de résistances d 0 à 1000 MΩ et une gamme de décibels de −6 dB à +66 dB. Il est fourni complet avec sonde à 3 directions. Les accessoires prévus comprennent une sonde en cristal à haute fréquence et une sonde à haute tension pour la mesure directe de tensions allant jusqu'à 60 kV.

EE 76 754 *pour plus amples renseignements*

TÊTES D'ENGRENAGE DE PRÉCISION
Ancillary Developments Ltd, Blackwater Station Estate, Blackwater, Camberley, Surrey
(Illustration à la page 855)

Une nouvelle gamme complète de petites têtes d'engrenage et de réducteurs de vitesse de précision pour les applications d'asservissement électromécanique de haute qualité sera bientôt produite par la société Ancillary Developments Ltd. Elle comprend les premiers composants britanniques de ce genre réalisés dans la dimension internationale 08, ainsi que quatre autres dimensions d'usage commun. Bien que toute la gamme ait été conçue conformément aux spécifications du Royal Aeronautical Establishment, ces dispositifs ne sont pas limités aux applications aéronautiques. Tous les matériaux utilisés sont résistants à la corrosion et on a prévu un choix de paliers simples ou à billes pour supporter les arbres. Des plaques d'adapteurs peuvent être fournies pour s'adapter à la tête d'engrenage ou au réducteur de vitesse avec différentes dimensions de servomètre, et tous les éléments peuvent être montés sur ergot ou bride, selon les besoins.

Différentes formes d'arbre de sortie sont également prévues.

Des rapports d'engrenage jusqu'à 1000 : 1 peuvent être obtenus dans les formats 08, 10 et 11 jusqu'à 5000 : 1 dans les formats 15 et 18.

EE 76 755 *pour plus amples renseignements*

CONTRÔLEUR CONSTITUÉ DE CORPS SOLIDES
Kent Precision Electronics Ltd, Vale Road, Tonbridge, Kent
(Illustration à la page 855)

La société Kent Precision Electronics Ltd, a réalisé un contrôleur étalonné constitué de corps solides, pour le contrôle de la sensibilité et les systèmes de protection ou d'alarme, fonctionnant sur entrées de courant alternatif ou de courant continu dans une gamme étendue, pour la commutation de charge atteignant plusieurs kilowatts ou, en variante, fonctionnant à partir d'un relais électromagnétique avec contacts de permutation isolés, afin d'en faciliter l'utilisation. Cet instrument remplace le type classique de contrôle par instrument à contact dans toutes les applications. Etant entièrement constitué de corps solides, l'appareil se distingue par la sûreté de son fonctionnement et sa longue durée d'emploi sans entretien.

Il comporte une échelle linéaire, étalonnée en fonction du paramètre devant être contrôlé, ainsi que des aiguilles simples ou doubles calées aux positions auxquelles le contrôle devra s'effectuer. Les contacts des commutateurs ou relais de puissance de sortie changent d'état lorsque le paramètre d'entrée traverse ces points.

L'instrument se compose essentiellement d'un amplificateur à courant continu transistorisé et à faible dérive dont le rôle est celui d'un détecteur d'équilibre entre une tension de référence contrôlée et le paramètre d'entrée. La sortie de cet amplificateur amorce un circuit de déclenchement qui commande à son tour un circuit de sortie à thyristors afin de commuter la charge, soit avec contrôle proportionnel soit avec contrôle arrêt-marche. Lorsque la puissance contrôlée est au-dessus de l'indice nominal maximum du thyristor ou lorsqu'il y a lieu d'effectuer une commutation isolée, un circuit à thyristors de faible puissance est utilisé pour mettre en oeuvre un relais ou contacteur électromagnétique à l'indice voulu.

L'instrument comprend normalement des relais classiques à puissance de contact nominale de 250 V c.a., 5 A. D'autres indices peuvent être prévus sur demande. Des sorties normales de thyristor atteignant jusqu'à 600 V et 100 A peuvent être prévues sur spécification.

La mise en oeuvre de l'appareil se fait à partir d'une alimentation secteur auxiliaire, au moyen d'un transformateur d'isolement avec réglage de variations de ±20 %. Le coefficient de température est de 0,05 %/° C pour toutes les gammes et l'appareil peut être utilisé

dans des temperatures ambiantes atteignant 85° C.

EE 76 756 *pour plus amples renseignements*

POTENTIOMÈTRES POUR ÉTUDIANTS
Croydon Precision Co., Hampton Road, Croydon, Surrey
(Illustration à la page 855)

Le potentiomètre pour étudiants, type P.1, a été prévu pour l'emploi dans les collèges techniques et lycées, afin d'y remplacer ou servir de complément au potentiomètre à fil de contact glissant du type simple dit à "mandoline." Il est de construction fort robuste et sa précision est telle qu'il peut être utilisé pour les travaux les plus importants avec pleine garantie de résultats pouvant se répéter.

Sa gamme de mesure de tension s'étend de 0 à 1,1 V, la plus petite subdivision étant de 0,001 V. L'appareil comprend deux cadrans de mesure dont le premier comporte 10 plots de 0,1 V, le second ayant la forme d'un fil de contact glissant à plots de 0,1 V sur la totalité de l'échelle, subdivisée jusqu'à 0,001 V. La précision de mesure pouvant être atteinte est de 0,1 % au 0,002 V, selon celle qui est la plus élevée. Un circuit de standardisation est incorporé à l'appareil afin de lui permettre d'être standardisé quels que soient les réglages du cadran. Le courant du potentiomètre est réglé au moyen d'un rhéostat à plusieurs tours de résistance et à contact constant "Cropico." Tous les commutateurs nécessaires sont incorporés et des bornes sont prévues pour relier la cellule normale, le galvanomètre, la batterie d'alimentation de 1,5 V et la tension devant être mesurée.

Ce potentiomètre se caractérise en particulier par le panneau supérieur graphique qui indique clairement le bobinage de circuit et les réseaux de résistance. Ces derniers ont été montés de façon parfaitement distincte afin de permettre à l'étudiant de comprendre aisément la fonction des diverses commandes. Le temps nécessaire au professeur pour expliquer le fonctionnement et le rôle de l'instrument se trouve par conséquent considérablement réduit.

Une pile standard Weston à acide saturé, d'une tension absolue de 1,01859 V à 20° C, est fournie comme accessoire. Cette pile est renfermée dans un solide moulage en bakélite avec borne isolée.

On a prévu pour le détecteur un galvanomètre portatif à cadre mobile avec panneau frontal en pente. La sensibilité de 2 μA par division d'échelle est pleinement suffisante, que le potentiomètre soit normalisé ou qu'il soit utilisé dans des conditions de court-circuit.

EE 76 757 *pour plus amples renseignements*

MOTEURS À AMORTISSEMENT D'INERTIE
Bowmar Instrument Ltd, Sutherland Road, London, E.17
(Illustration à la page 856)

La société Bowmar Instrument Ltd, a

réalisé de nouveaux types de moteurs à amortissement d'inertie, format 08, pouvant être utilisée sur 400 Hz. Des tensions de référence de 115 V ou 26 V et des tensions de contrôle suivant la demande du client sont prévues pour ces moteurs. Le moteur à répulsion type IM83-002 est à enroulement auxiliaire de contrôle de 33 V; ses caractéristiques de décrochage sont de 23,5 g/cm et sa vitesse de charge est de 6500 tours/min. L'inertie de volant est de 4,6 g/cm² avec amortissement de 245 dynes/cm/sec.

Le moteur type IM82-001 comporte un bobinage de commande à prise centrale, un couple de décrochage de 21 g/cm, et une vitesse sans charge de 5500 tours/min. L'inertie de volant est de 1,6 g/cm² avec amortissement de 300 dynes/cm/sec.

Les deux types d'appareils comprennent des fils métalliques de sortie de 30 cm et des arbres cannelés de 120 DP à 13 dents. Ils peuvent être utilisés dans une température ambiante de −65° C à +125° C. En outre, ces moteurs peuvent être fournis avec des têtes d'engrenage à rapport maxima de 2000:1. Il existe également des moteurs du format 11 ayant les mêmes tensions de bobinage et les mêmes facteurs d'amortissement.

EE 76 758 pour plus amples renseignements

PERCEUSE RIGIDE À CIRCUIT IMPRIMÉ
Drill Service (Horley) Ltd. 89 Albert Road, Horley, Surrey

Répondant à la demande constante d'éléments de perçage pouvant être employés sur les machines à circuit imprimé spéciales, la société Drill Service offre maintenant une perceuse spéciale complète.

Le fonctionnement de la perceuse est basé sur la méthode classique et éprouvée qui consiste à percer des plaquettes superposées par en-dessous tout en les localisant à partir d'un gabarit placé au-dessus.

Les plaquettes devant être percées sont superposées dans un bâti de montage avec un gabarit sur la partie supérieure. Les plaquettes et le bâti sont ensuite déplacés sous un style logé dans les trous du gabarit. Il s'agit d'un style à ressort soumis à une pression légère constante sur le gabarit afin de faciliter la localisation. Aussitôt cette dernière effectuée, l'opérateur met en action le bouton de pied afin de mettre en oeuvre le cycle automatique de perçage.

Lorsqu'on presse le bouton, la pression d'air est appliquée au mécanisme d'alimentation du processus de perçage ainsi qu'à la tête de localisation. Cette dernière est fermement fixée sur les plaquettes qu'elle comprime et verrouille dans la position de perçage.

Lorsque le cycle de perçage est achevé le mécanisme de perçage se retire rapidement et la pression de fixation se relâche pour permettre une nouvelle localisation. Les deux mains de l'opérateur sont

libres afin qu'il puisse localiser le gabarit. Une piste sur le gabarit peut accroître la vitesse du cycle car le style à ressort léger suit la piste d'un trou à l'autre.

La tête de perçage est du modèle 111 et ses caractéristiques sont les suivantes:

Vitesse variable à l'infini jusqu'à 15000 tours/min.

Vitesse d'alimentation à variation infinie.

Tant l'alimentation que la vitesse sont contrôlées pour assurer la réalisation de trous nets.

Un dispositif à minuterie pour le nettoyage de la perceuse permet à cette dernière de se retirer et de dégager les copeaux à intervalles réguliers afin d'empêcher l'encrassement aux vitesses d'avance élevées sur les gros empilements de plaquettes.

La tête de perçage est alignée avec précision sur le style et le manchon, de sorte que des empilements de 6 à 8 plaquettes peuvent être percés très rapidement par des opérateurs non qualifiés utilisant des outils de perçage au carbure solide, sans le moindre risque de casse accidentelle.

Cette nouvelle perceuse est particulièrement indiquée pour l'emploi avec des outils de perçage pour plaquettes de circuit au carbure de tungstène solide à rallonge. Ces outils de perçage peuvent conserver leur dents de fraisage et leur dimensions pendant de longues périodes, économisant ainsi le temps de montage et la destruction inutile de plaquettes et de composants au cours de l'assemblage.

EE 76 759 pour plus amples renseignements

HUMIDIMÈTRE
The Wayne Kerr Laboratories Ltd. Sycamore Grove, New Malden, Surrey
(Illustration à la page 856)

L'humidimètre MB100 constitue le premier d'une nouvelle gamme d'instruments de mesure compacts réalisés par Wayne Kerr et qui porteront le nom d'instruments de la série "Midge." Cet appareil autonome n'exige aucun montage ou étalonnage et il est prêt à l'emploi immédiat à n'importe quel moment. Il fonctionne et liaison avec un transducteur à bas prix et fournit des indications directes d'humidité relative de 20 à 100 %, en deux gammes. Des ensembles scellés peuvent être réalisés de manière à ce que chacun comprenne un élément transducteur avec câbles reliés à un connecteur sur le coffret extérieur, ce qui permet la vérification instantanée et sans déballage de l'état de matériels emmagasinés ou expédiés dans des conditions de forte humidité.

Le précision de mesure totale, transducteur y compris, est de 5 %. Les transducteurs peuvent résister à des variations de température de −40° C à +100° C et la qualité de leur comportement n'est nullement affectée par l'emmagasinage prolongé ou l'utilisation dans des conditions d'humidité relative extrêmes. Le circuit de mesure consiste

essentiellement en un pont à bras de rapport transformateur fonctionnant à 2 kHz et équilibré électroniquement à l'aide d'un amplificateur à réaction à gain élevé. La source, le détecteur et la batterie sont logés dans le coffret (mesurant 12 cm × 7 cm × 15,7 cm) et une sortie de 0 à 100 µA est prévue pour les enregistreurs, etc.

EE 76 760 pour plus amples renseignements

TRANSISTOR AU SILICIUM À OXYDE MÉTALLIQUE
Mullard Ltd, Mullard House, Torrington Place, London, W.C.1

En annonçant la réalisation d'un nouveau transistor au silicium à oxyde métallique, la société Mullard Ltd, déclare que la production en série de ce composant le rendrait considérablement meilleur marché que n'importe quel autre composant analogue actuellement en vente en Grande-Bretagne. La production en grande série est donc prévue. Entretemps, le nouveau composant peut être fourni en quantitées limitées. Une grande fiabilité est assurée par l'emploi de techniques planaires éprouvées.

Le transistor au silicium à oxyde métallique, type 95BFY, a une résistance d'entrée d'un million de mégohms ($10^{12}\Omega$), c'est à dire une résistance bien plus élevée que celle d'un transistor classique ou même celle d'un transistor à effet de champ. Il pourrait, par exemple, être utilisé comme convertisseur d'impédance élevée en impédance réduite ou comme amplificateur à basse fréquence pour instruments électroniques et circuits de contrôle, ainsi que pour les calculatrices et le matériel de télécommunication. Il se prête également à l'emploi dans les circuits logiques. Etant donné que le transistor au silicium à oxyde métallique est un dispositif porteur majoritaire, il est insensible aux radiations et il peut donc trouver des applications dans les satellites et autres véhicules spatiaux.

Les premiers travaux d'application montrent que le composant peut être utilisé comme oscillateur à des fréquences atteignant 150 MHz, comme mélangeur donnant un gain de conversion de 20 dB (la couche sous-jacente étant utilisée comme borne de second signal) ou comme étage de sortie modulée par "anode" pour des puissances maxima de plusieurs centaines de milliwatts à des fréquences atteignant 100 MHz. En tant que relais modulateur, son comportement est particulièrement remarquable car il ne présente aucune tension "arrêt-marche" et une résistance "marche" d'environ 400 Ω à des fréquences d'interruption pouvant atteindre 100 kHz. De plus, il ne présente aucun affet d'emmagasinage de sorte que lorsqu'il est utilisé pour les applications de commutation, la durée de commutation ne dépend que de la constante de temps RC de la capacité de porte passant par la résistance des voies.

Le transistor 95BFY est produit par la diffusion de deux régions n+ à espacement réduit (source et perte) sur une couche sous-jacente de silicium du type p. Entre la source et la perte, une couche mince d'oxyde de silicium revêt la couche sous-jacente et une électrode de porte en aluminium s'évapore sur la partie supérieure de la couche d'oxyde.

Lorsqu'un potentiel positif est appliqué à la porte, une charge négative est induite à la surface de la couche sous-jacente, ce qui provoque la formation par une voie de type n d'un chemin conducteur entre la source et la perte. Si l'on applique alors un potentiel entre la source et la perte, il s'en suivra un flux de courant pouvant être commandé par le potentiel appliqué à la porte.

EE 76 761 pour plus amples renseignements

GÉNÉRATEUR DE VIBRATIONS
Derritron Electronic Vibrations Ltd, 24 Upper Brook Street, London, W.1
(Illustration à la page 857)

Ce générateur de vibrations d'une poussée de 2721,5 kg se compose d'un amplificateur de puissance de 30 kVA, conçu pour le contrôle aléatoire et sinusoïdal, et d'un vibrateur à poussée de 2721,5 kg et d'un appareillage de commande. La dissipation anodique continue maxima de l'amplificateur est de 50 kW, sa gamme de fréquences à pleine puissance est de 15 Hz à 5 kHz et il peut fournir un courant total jusqu'à 5 Hz. L'équipement comprend des tubes mous English Electric du tout dernier modèle, le bourdonnement et le bruit étant inférieurs d'au moins 70 dB par rapport à la puissance de sortie totale.

Le vibrateur à poussée de 2721,5 kg (type No. VP.600) utilise un système de refroidissement à bobine d'entraînement qui fournit un assemblage de bobine et de table extrêmement rigide. Le refroidissement s'effectue normalement à l'eau distillée, bien qu'on puisse utiliser le refroidissement à l'huile pour certaines conditions extrêmes de chambre climatique. Les caractéristiques essentielles de ce vibrateur sont une poussée vectorielle de 2721,5 kg, une gamme de fréquences d'environ 5 Hz à 3 kHz, utilisable jusqu'à 5 kHz, dont la première résonance perturbatrice importante est de 2,5 kHz. Le poids réduit de la table, soit 32 kg, permet une accélération de table non chargée dépassant 85 g.

EE 76 762 pour plus amples renseignements

RELAIS SENSIBLES À LA TENSION
Newton Brothers (Derby) Ltd, Alfreton Road, Derby
(Illustration à la page 857)

Les relais sensibles à la tension Newton Derby TVS ont été conçus pour permettre la commutation à un niveau de tension prédéterminé et ils peuvent être utilisés dans les circuits de protection de surtension et de sous-tension qui exigent une grande précision.

La gamme de perception des tensions varie entre 10 et 220 V c.c., la sensibilité étant de ±1% et le différentiel de 1%. La vitesse de réponse est de 10 à 15 m sec, mais lorsqu'il y a lieu d'empêcher la disjonction par des surcharges dans les conditions de phénomènes transitoires, un retard de temps de quelques millisecondes à 2 minutes peut être prévu.

Ces relais sont munis de fiches à dix broches et peuvent être aisément incorporés à des cabines de contrôle. Chaque relais est doté de 2 contacts de puissance normalement fermés et d'un contact de puissance normalement ouvert.

EE 76 763 pour plus amples renseignements

MOTEURS ASSERVIS À COURANT CONTINU SANS BALAIS
Vactric Control Equipment Ltd, Garth Road, Morden, Surrey
(Illustration à la page 857)

Le nouveau moteur d'asservissement à courant continu et sans balais que vient de réaliser la société Vactric Control Equipment Ltd dans le format 11 a été conçu pour répondre à la demande d'un élément d'un plus longue durée de vie que le moteur d'asservissement à courant continu de type classique. Il est particulièrement indiqué pour la suppression des bruits radiophoniques. Il ne comporte ni balais ni commutateur et sa longue durée est essentiellement due à l'absence d'arc et d'usure des balais. Il a été spécialement bobiné pour la réception de formes d'ondes à partir d'un appareil électronique séparé.

Le nouveau moteur a été entièrement construit avec des matériaux résistant à la corrosion et il peut être utilisé dans une gamme de températures ambiantes de —65° C à +85° C, à des altitudes maxima de 18288 mètres.

L'appareillage électronique contient un multivibrateur qui commande une série de basculeurs à travers des circuits logiques appropriés, les basculeurs comportant des transistors de puissance qui alimentent les enroulements du moteur. Le volume de l'élément électronique est d'environ 2 pouces cubes et il peut résister aux même conditions d'environnement que le moteur. Il peut être fourni sous forme encapsulée dans un coffret cylindrique format 11 avec cheville de soudage. En variante, une plaquette à circuit à fiches peut être fournie pour l'utilisation dans des conditions moins astreignantes.

EE 76 764 pour plus amples renseignements

PONT DE COMPOSANTS
K.L.B. Electric Ltd, 335 Whitehorse Road, Croydon, Surrey
(Illustration à la page 857)

Le pont de composants M.300 est un élément d'une grande souplesse d'utilisation et à bas prix comportant quatre gammes de capacité de 10 pF à 1000 mF et quatre gammes de résistance de 0,05 Ω à 100 MΩ. Il permet également

de contrôler les fuites de condensateur avec une tension de contrôle réglable jusqu'à 500 Vc.c. Les contrôles de rapport dans les gammes de 0,05 à 1 et de 20 à 1 peuvent être effectués entre n'importe quels deux condensateurs, inducteurs ou résistances. Ce contrôle s'effectue lorsque des composants doivent être réglés par rapport à un étalon.

EE 76 765 pour plus amples renseignements

CIRCULATEUR COMMUTABLE
The Marconi Co. Ltd, Chelmsford, Essex
(Illustration à la page 858)

La société Marconi a mis au point un nouveau circulateur commutable à trois directions du format de guide d'ondes No 15, couvrant la gamme de fréquences de 7,75 à 8,5 GHz. La commutation à grande vitesse du sens de la circulation s'effectue de manière électromagnétique.

Le nouveau circulateur a été conçu pour être utilisé avec le matériel de communication microondes. Il présente de nombreuses améliorations en ce qui concerne la fiabilité, la vitesse de fonctionnement et l'économie d'espace par rapport au dispositif mécanique de commutation précédemment employé. Il convient à toutes les applications de commutation de guide d'ondes à deux directions, dont un exemple caractéristique serait les réseaux de branchement reliant les émetteurs à microondes et les récepteurs à leurs antennes.

Le circulateur commutable est une version du circulateur standard à jonction à trois directions mais le champ de polarisation magnétique est fourni électromagnétiquement et peut être inversé pour changer le sens de la circulation. L'électro-aimant et le circulateur ont été spécialement construits pour permettre l'inversion rapide du champ magnétique au moyen d'un circuit de commutation à transistors normal. Le temps de commutation du circulateur dépend de la puissance fournie à l'aimant. Des temps de commutation inférieurs à 0.5 msec peuvent être réalisés avec une alimentation de 25 W.

EE 76 766 pour plus amples renseignements

INSTRUMENT POUR MESURER L'INTENSITÉ DES SIGNAUX DE TÉLÉVISION
Lab-Craft Ltd, Gainsborough Road, Woodford Bridge, Essex
(Illustration à la page 858)

Le contrôleur d'intensité des signaux V.H.F., type 215, est un appareil de haute qualité utilisant les techniques de construction les plus modernes. Des lampes à circuit de cadre sont utilisées dans tous les étages amplificateurs et la haute tension stabilisée réduit la dérive et les variations de gain à des proportions négligeables sur toute la gamme de tensions secteur.

L'appareil est conçu sur la base d'un bloc d'accord très haute fréquence standard, relié à un amplificateur m.f. à gain

contrôlé. Le courant contrôlé de l'une des lampes d'amplification m.f. est indiqué sur l'instrument de mesure. L'échelle est étalonnée directement en microvolts et millivolts, donnant ainsi des étalonnages clairement lisibles des niveaux de grognement de réaction au niveau de "submersion" sans qu'il y ait lieu d'utiliser des atténuateurs à fiches séparés. L'augmentation des niveaux de signaux produit un décroissement dans le courant à travers l'appareil de mesure. Ce dispositif assure la protection automatique de l'instrument de mesure contre les surcharges accidentelles.

Le bloc d'accord précis est réglé d'avance pour chaque voie et il comporte une mémoire qui effectue le réglage automatique chaque fois que l'on tourne le barillet. Le montage non séquentiel des bobines dans le barillet rend instantané le changement de la bande 1 à la bande 3 dans la plupart des zones.

Le modèle 215 qui est d'une haute sensibilité et dont les indications vont jusqu'à 10 mV, peut être utilisé dans les emplacements exigeant l'emploi d'un pareil instrument. La douille de l'antenne est située à l'arrière de l'instrument. Grâce à une courte antenne télescopique, on peut estimer de manière relative et en n'importe quel lieu les conditions de gain de l'antenne.

EE 76 767 *pour plus amples renseignements*

CONNECTEUR À AUTOVERROUILLAGE
Distributeurs: Nutec Electronics Ltd, Mercantor House, East Street, Shoreham-by-Sea, Sussex

(Illustration à la page 858)

Les connecteurs Lemo, conçus par l'ingénieur suisse Léon Mouttet, comportent un système de verrouillage breveté qui obvie au décrochage accidentel dû aux vibrations ou au tirage par câble de la fiche et de la douille.

Les surfaces de contact sont dorées de manière à assurer une résistance de contact constante et très réduite. Des matériaux isolants de la plus haute qualité sont employés.

Il existe trois types fondamentaux de connecteurs Lemo, à savoir le type coaxial, biaxial et triaxial à haute tension ou à plusieurs broches. Chacun de ces types peut être fourni en plusieurs formats qui peuvent tous être combinés en un seul assemblage pour donner des variations multiples.

En dehors des modèles et des combinaisons normales, il existe des types à haute tension: des connecteurs scellés au vide et étanches; des versions spéciales pour haute et basse tension, ainsi que des fiches et systèmes coaxiaux pour la télévision.

Les divers assemblages de serrage peuvent recevoir une gamme étendue de câbles.

EE 76 768 *pour plus amples renseignements*

TRANSDUCTEURS SUBMINIATURE
Associated Engineering Ltd, Cawston, Rugby

(Illustration à la page 858)

Un nouveau type de transducteur subminiature à déplacement inductif a été réalisé pour la mesure du déplacement et des vibrations dans des conditions d'environnement défavorables.

Le nouveau transducteur qui est fabriqué dans les deux formats que l'on voit dans notre gravure, peut être utilisé à des fréquences porteuses maxima de 300 kHz, permettant ainsi la détection de vibrations ou du déplacement de n'importe quel objet métallique à des fréquences allant du courant continu à 60 kHz. Les deux transducteurs ont une gamme de régime de 1/100° de pouce et peuvent résister à des températures ambiantes atteignant 150° C sans changement appréciable de sensibilité. Lorsqu'ils sont montés de façon appropriée, ils peuvent être soumis à des accélérations de 2000 g sans le moindre défaut de fonctionnement.

Le poids et les dimensions des transducteurs, terminaisons de circuit imprimé compris, sont comme suit:

Grand format: 10,9 mm de longueur; 6,35 mm de diamètre; poids: 7 g.

Petit format: 10,2 mm de longueur; 4 min de diamètre; poids: 5 g.

Les transducteurs peuvent être fournis sur commande avec un anneau de montage fileté tel qu'on le voit à gauche de notre gravure. Autrement, ils sont fournis tel qu'on le voit au centre et à droite, et ils peuvent être montés par serrage ou par l'emploi d'un cément ou d'une résine d'époxyde appropriée.

EE 76 769 *pour plus amples renseignements*

FERS À SOUDER
Distributeurs: Lugton & Co. Ltd, 209-212 Tottenham Court Road, London, W.1

Les fers à souder Imperial Ungar, qui sont fabriqués au Canada, sont maintenant vendus par la société Lugton & Co. Ltd.

La principale caractéristique de ce fer à souder est son adaptabilité, car toutes ses parties sont rapidement interchangeables, de sorte qu'il peut être converti en quelques minutes, par exemple, d'un fer de 25 W à barre miniature en un fer à barre allongée de 40 W. Un choix étendu de barres est prévu.

L'étui à chaleur scellé en acier inoxydable assure une plus longue durée, élimine l'oxydation et fournit un transfert de chaleur maximum. Les différentes couleurs permettent de distinguer facilement deux fers à souder d'une tension différente.

EE 76 770 *pour plus amples renseignements*

GÉNÉRATEUR DE FORMES D'ONDES
Feedback Ltd, Crowborough, Sussex
(Illustration à la page 859)

Le générateur de formes d'ondes, type TWG 300, est un instrument d'une grande souplesse d'emploi conçu pour les travaux de recherche dans le domaine du comportement des systèmes de contrôle.

Le fonctionnement du TWG 300 est basé sur un circuit de commutation et d'intégration produisant des ondes triangulaires et carrées. La forme d'onde triangulaires est façonnée dans un réseau à diodes de manière à former une onde sinusoïdale constante en amplitude et à faible teneur en harmoniques sur toute la gamme de fréquences. Un autre réseau produit l'onde triangulaire tronquée qui donne à tour de rôle des parties "plates" et des "rampes" lorsqu'on supprime les crêtes de la forme d'onde triangulaire. Le circuit d'ébarbage est réglable permettant ainsi de varier la pente de la rampe, tout en maintenant l'amplitude constante, et de réaliser une rampe réglable.

La fréquence s'étend de 0,001 à 1000 Hz en six gammes à commutation de 10:1, avec réglage continu dans chaque gamme. Les gammes sont étalonnées en fréquence et en période et il y a un chevauchement de 20% aux extrémités des gammes ce qui assure une continuité suffisante de couverture. La précision de l'échelle est supérieure à $\pm 5\%$. La stabilité d'heure en heure est typiquement supérieure à 0,1%. Pour le réglage de fréquence par rapport à des compteurs ou à des étalons de fréquence, la discrimination est supérieure à 0,1 % de la fréquence. La forme d'onde choisie est fournie simultanément sur deux sorties: soit à un niveau constant de 40 V de crête à un courant maximum de 20 mA, impédance de sortie de 10 Ω, soit au moyen d'un atténuateur réglable de manière continue de 2 kΩ étalonné de 0 à 40 V et un commutateur série donnant des conditions de X 1-X 0,1—arrêt. L'amplitude de sortie est constante quel que soit le réglage de la fréquence. Les sorties sont prises entre les bornes "directe" et "commune". La borne de terre ou de masse, reliée au châssis de l'instrument, est normalement accouplée à la borne "commune". Lorsqu'on supprime la liaison, des alimentations extérieures peuvent être reliées de manière à "flotter" l'instrument jusqu'à 250 V à partir du potentiel de terre. Il faut s'assurer que l'alimentation extérieure soit pourvue d'un faible chemin d'impédance en cas de captage éventuel de bourdonnement.

Les sorties de contrôle, carrées et triangulaires, ont une amplitude constante de 5 V de crête et une impédance de source de 20 kΩ. Ces sorties sont synchronisées avec la forme d'onde de sortie principale.

EE 76 771 *pour plus amples renseignements*

BASE DE TEMPS ET COMPTEUR
Specto Avionics Ltd, Feltham, Middlesex
(Illustration à la page 859)

La société Specto Avionics Ltd, a réalisé un instrument de mesure universel pour mesurer la fréquence, la période et

la largeur d'impulsion. Cet instrument a été conçu pour les mesures exigeant une très grande précision et il sera particulièrement utile aux établissements de recherche, aux universités et à l'industrie.

L'instrument comporte deux éléments Spectro Avionics standard: le chronomètre de déci-microsecondes, type 222, et la base de temps, type 233. La précision de ces éléments combinés est de ± 2 parties dans 10^7 à long terme; ± 1 partie dans 10^8 à court terme. De même que tous les instruments de la gamme, la base de temps, type 233, est à transistors au silicium et utilise les plus récentes méthodes de circuits imprimés; elle est de construction modulaire légère, pouvant être montée sur bâti ou utilisée indépendamment sur banc d'essai. La base de temps mesure 55 cm $\times$ 8,89 cm $\times$ 35,56 cm. La hauteur totale de la base de temps et du chronomètre à décimicrosecondes est de 22,2 cm.

Les caractéristiques essentielles des instruments combinés sont les suivantes:

Mesure de temps et de largeur d'impulsion—0,1 sec à 10^8 sec.

Entrées directes—± 1 V à ± 5 V impulsion de crête pour une charge de 1 kΩ.

Entrées par l'atténuateur—± 5 V à ± 25 V impulsion de crête pour une charge de 1 kΩ.

Fréquences—30 Hz à 20 MHz.

Périodes—1 événement à 10^8 événements; période minimum: 0,05 μsec.

Rapport—f_1, 20 MHz max. f_2, 15 MHz max.

Sensibilité d'entrée—100 mV à 5 V efficaces.

Impédance d'entrée—10 kΩ.

Gamme de températures—$-20°$ C à $+40°$ C.

EE 76 772 *pour plus amples renseignements*

SYNCHROS SANS BALAIS
Elliott-Automation Ltd, Century Works, Lewisham, London, S.E.13
(Illustration à la page 859)

La Division des Composants d'asservissement de la société Elliott-Automation vient de mettre au point un nouveau synchro sans balais en vue des applications nécessitant des composants d'une grande sécurité d'emploi.

Les nouveaux synchros Elliott-Ellsynn sont fournis sous forme d'émetteurs et de transformateurs de commande, format II. Ils présentent de nombreux avantages par rapport aux composants classiques tout en gardant des caractéristiques électriques et mécaniques analogues. Les balais et les bagues collectrices ont été remplacées par des transformateurs à injection miniature, ce qui supprime les problèmes de contact des balais et réduit le collage à environ un quart du niveau que l'on trouve dans les synchros classiques. Ces nouveaux composants sont donc particulièrement indiqués pour les instruments de vol.

Ils peuvent être utilisés dans des températures atteignant $+125°$ C avec erreurs maxima de ± 14 minutes d'arc. Les travaux en cours visent à améliorer ce rendement et à permettre le fonctionnement dans des températures ambiantes maxima de $+150°$ C.

Grâce à une conception judicieuse, on a pu maintenir le transformateur de contrôle à la même longueur que celui des versions classiques. L'émetteur de contrôle, qui reçoit plus de puissance, est plus long de 1,25 cm que le synchro standard.

Tant le synchro sans balais à 115 V que celui à 26 V dans la gamme actuelle Ellsyn sont interchangeables avec les éléments classiques.

EE 76 773 *pour plus amples renseignements*

ORTHICONOSCOPE
English Electric Valve Co. Ltd, Chelmsford, Essex
(Illustration à la page 859)

La English Electric Valve Co. Ltd, vient d'annoncer une nouvelle réalisation dans le matériau de cible de ses orthiconoscopes.

Ce nouveau matériau, portant le nom de ELCON, assure une nouvelle durée d'au moins 750 heures aux orthiconoscopes. Les premiers résultats obtenus sur les échantillons de pré-production indiquent des durées de vie atteignant 3000 à 5000 heures.

Le matériau ELCON réalise une absence complète de rétention d'image, une grande stabilité de reproduction de contrastes, une perte de sensibilité fortement réduite et la commutation immédiate.

L'ELCON introduit un nouveau principe de fonctionnement utilisant la conduction électronique plutôt qu'ionique, pour le transfert de la mire du côté de l'image au côté du faisceau. La migration des ions est donc pratiquement éliminée et, par conséquent, il n'y a aucune détérioration importante du comportement dans le temps. La sensibilité et la stabilité de la reproduction des contrastes (gamma) demeure constante et il n'y a aucun risque de "collage" pendant toute la durée du matériau.

La English Electric Valve Co. Ltd, déclare que le nouveau matériau n'a aucun défaut caché et qu'il répond de manière très satisfaisante aux problèmes rencontrés avec n'importe quel type normal d'orthiconoscope.

La gamme complète des orthiconoscopes EEV de 7,5 cm comprend l'ELCON qui est actuellement produit en série.

Les premiers lots du tube EEV de 11,25 cm, qui comporte également ce matériau, seront prêts prochainement et la production en grande série commencera au début de 1965.

Le nom ELCON provient des propriétés du nouveau matériau c'est à dire "ELectronic CONducting," par rapport au conducteur ionique normalement utilisé.

EE 76 774 *pour plus amples renseignements*

Résumés des Principaux Articles

Un instrument linéaire transistorisé pour la mesure de la densité à quatre décades par E. Gordy, P. Hasenpusch et G. F. Sieber

Résumé de l'article aux pages 808 à 812

On a réalisé un circuit linéaire de mesure de la densité constitué de corps solides et dont la gamme de fonctionnement de densité optique va de 0 à 4. Ce circuit est à contre-réaction à tension de dynode combinée avec la correction de la linéarité par diodes polarisées. La linéarité du signal de sortie par rapport à la densité optique est de $\pm 0,5\%$. Une version au germanium et à transistors au silicium est décrite.

Un système expérimental d'allumage électronique par W. Alexander et P. Sample

Résumé de l'article aux pages 813 à 816

Il s'agit ici d'une étude expérimentale des avantages et des désavantages d'un système d'allumage électronique pour moteur de voiture automobile. Dans la méthode utilisée, un condensateur est déchargé dans le primaire d'un transformateur à l'aide d'un redresseur piloté au silicium utilisé pour la commutation. Il est démontré ainsi qu'un rendement accru peut être obtenu du moteur.

La démodulation d'un signal modulé en fréquence par deux tonalités   par G. May

*Résumé de l'article
aux pages 816 à 819*

Le processus de modulation et le spectre diffus des fréquences latérales qui apparaissent par suite de la modulation en fréquence d'une porteuse avec deux tonalités ont été examinés dans plusieurs textes. En revanche, il est étonnant de constater combien peu d'attention a été apportée au processus inverse de démodulation, indiquant la manière dont le signal de modulation de fréquence est transformé en une onde modulée en amplitude dans un discriminateur typique.

Le but de cet article est de montrer la façon dont ces fréquences latérales, en particulier les fréquences combinées, sont utilisées au cours de la démodulation.

Spectromètre électronique pour fréquences acoustiques et sub-acoustiques   par D. Llanwyn Jones

*Résumé de l'article
aux pages 820 à 823*

L'auteur développe la théorie d'un spectromètre simple pouvant être utilisé aux fréquences acoustiques et sub-acoustiques et indique la sélectivité du circuit et les caractéristiques de stabilité. Il indique également un circuit pratique qu'il analyse par rapport au résultats de l'analyse théorique.

Décodage de signaux SECAM pour tubes à canon unique   par P. Cassagne et M. Melchior

*Résumé de l'article
aux pages 824 à 827*

Cet article traite de l'emploi du signal SECAM dans les récepteurs à tubes à canon unique. Il décrit un nouveau circuit pour la transformation directe des signaux SECAM, sans démodulation des signaux de chrominance, en signaux CCS. Ce circuit apporte une solution simple et au moins aussi efficace que celle utilisée avec le signal NTSC.

En outre le même dispositif permet de construire des transcodeurs SECAM de haute qualité.

Amélioration de la tenue des tubes de comptage à cathode froide   par F. Szlavik et L. Garamszegi

*Résumé de l'article
aux pages 828 à 831*

Les auteurs de cet article examinent quelques méthodes réalisées pour améliorer le comportement des tubes de comptage à gaz à vitesse de comptage élevée. Ils décrivent des circuits permettant d'augmenter la vitesse de comptage sûre d'impulsions bien au delà de la limite de fréquence indiquée dans la spécification du fabricant ou pouvant être obtenue par les montages décrits dans les notices. On a pu ainsi atteindre des fréquences de comptage de 100kHz et 2MHz pour des tubes à 40 et à 20 électrodes respectivement.

Tensions harmoniques dans les sorties de circuits redresseurs commandés   par R. Smith

*Résumé de l'article
aux pages 832 à 837*

Dans cet article, les tensions harmoniques se trouvant à la sortie de circuits redresseurs commandés sont obtenues mathématiquement. Les courbes de l'ondulation équivalente se rapportant à 800Hz (ondulation pondérée), telle qu'utilisée pour l'usage téléphonique, sont également indiquées. L'article donne en outre un tableau des rapports de pondération d'ondulations.

Un circuit à temps de propagation de phase compensé en fréquence   par A. De Sa et L. Molyneux

*Résumé de l'article
aux pages 838 à 839*

Ce circuit produit un temps de propagation de phase pouvant être fixé à n'importe quelle valeur entre 40 et 280°, mais qui varie très peu en fonction de la fréquence. Il s'agit donc en somme d'un circuit à temps de propagation de phase compensé en fréquence. Ce circuit peut être utilisé également pour la multiplication de fréquence.

Ondes stationnaires sur lignes de transmission   par W. S. Metcalf

*Résumé de l'article
aux pages 840 à 842*

Après avoir analysé les lignes de transmission, l'auteur en tire des formules sur la base desquelles il présente un tableau indiquant les trois rapports d'amplitude de tension d'ondes stationnaires, à savoir, le coefficient de réflexion, la puissance réfléchie, le rapport de puissance incidente à la puissance réfléchie et la perte de transmission.

Un voltmètre à tension continue de 0,2% d'un modèle simple et pratique   par M. R. Osborne

*Résumé de l'article
aux pages 843 à 845*

L'instrument dont il est question dans cet article peut mesurer des tensions allant jusqu'à 20V c.c. avec une précision de ±0,2% au-dessus de 0,5V et au-dessous de ±1mV. C'est un instrument de comparaison par lequel l'inconnue est mise en équation avec une tension connue obtenue d'un potentiomètre hélicoïdal de précision et d'une diode de référence au silicium. Le comparateur a une caractéristique de commande de gain automatique obviant à l'emploi d'une commande de sensibilité. Par sa précision et son prix, cet instrument comble le vide entre l'instrument de mesure classique à cadre mobile et le voltmètre numérique.

Un commutateur à action lente   par K. F. Knott

*Résumé de l'article
aux pages 846 à 848*

Cet article décrit la commutation d'un transistor par une méthode qui permet de rendre le temps de commutation aisément variable pendant plusieurs secondes. L'article décrit également un dispositif de déclenchement Schmitt à hystérésis à variation rapide.
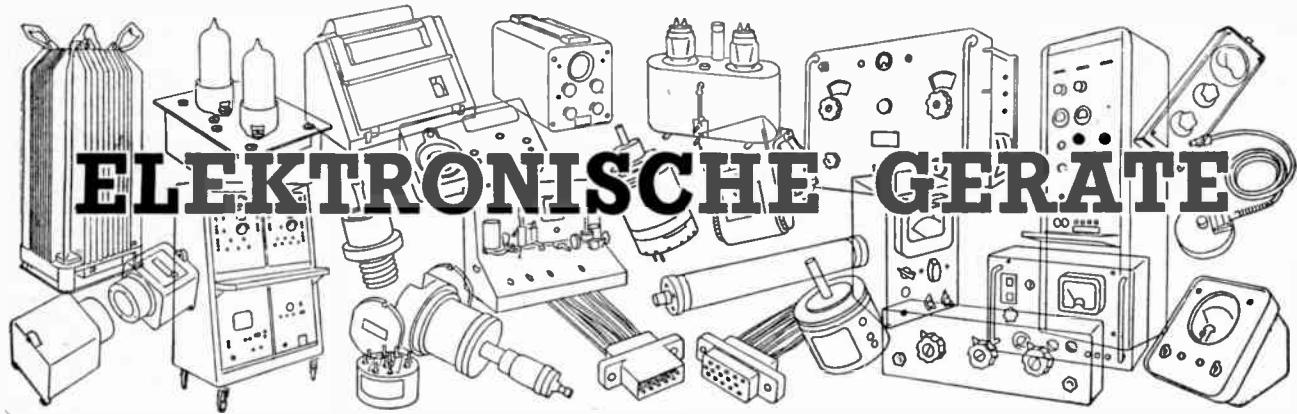
Beschreibung neuer Bauelemente, Zubehörteile und Prüfgeräte auf Grund der von Herstellern gemachten Angaben.

Übersetzung der Seiten 854 bis 859

Messwerterfassung

**Cambridge Instrument Co. Ltd,
13 Grosvenor Place, London, S.W.1**

(Abbildung Seite 854)

Die Cambridge-Allscott-Messwerterfassungsanlage wurde zusammen mit James Scott (Electronic Engineering) Ltd entwickelt. Die Messwerte einer beliebigen Anzahl von Kanälen werden zusammen mit dem Zeitpunkt der Messung und einer Kennziffer nacheinander in Digitalform angezeigt und registriert. Ein eingebautes einstellbares Alarmsystem gibt Sichtwarnung, wenn der gemessene Wert eines beliebigen Kanals oder einer Kanalgruppe vorgewählte Grenzwerte über- oder unterschreitet.

Die Anlage misst und registriert jede beliebige Variable, die sich in eine zweckmässige elektrische Spannung oder Widerstandsänderung umwandeln lässt; durch Wahl geeigneter Modul-Baugruppen und Primär-Messwertwandler kann man gemischte Eingänge nach beiden Methoden verarbeiten. Die Kanaleingänge werden nacheinander mittels einer Allscott-Abtasteinrichtung mit Allscott-Automatik für periodische Wiederholung an ein Cambridge-Servonachabgleich-Messgerät geschaltet. Ein mit einem Messgerät ausgerüsteter Binär-Drehgeber gibt einen Digitalausgang ab, der in Dezimalform umgewandelt und zusammen mit Kanalnummer und Messzeitpunkt gespeichert wird. Somit ist nicht nur eine weitere Messung möglich, während das Ergebnis ausgedruckt wird, sondern auch eine kontinuierliche Anzeige der Kanalnummer und gemessenen Werte. Kanäle werden nach dem gewählten Programm automatisch mit einer Messung alle drei Sekunden abgetastet; obendrein kann man jedoch jeden gewünschten Kanal durch Drücken der entsprechenden Taste eines Signalfeldes manuell messen.

Üblicherweise hat die Ausrüstung sieben "hohe" und sieben "niedrige" einstellbare Alarmgrenzwerte, was 49 Kombinationen von Alarmtoleranzen ergibt. Auf Wunsch lassen sich zusätzliche Alarmpegel vorsehen. Jeder Alarmpunkt kann auf einem Schaltbuchsenfeld mit jedem Kanal oder jeder Kanalgruppe verbunden werden. Das Messgerät hat einen Schleifdrahtgeber, der an das Alarmpegelgerät ein dem gemessenen Wert des gewählten Kanals proportionales Bezugssignal abgibt. Wenn der gemessene Wert eines Kanals ausserhalb der Alarmgrenzen liegt, werden Kontakte erregt, und die entsprechende Lampe im Signalfeld leuchtet rot auf, bis sie von Hand gelöscht wird. Wenn der eingestellte Kanal im Alarmzustand ist, erfolgen Digitalanzeige und Registrierung in rot.

Kanalnummer, Messwert und Zeit werden normalerweise digital registriert, jedoch kann auch ein Streifenschreiber eingebaut werden, der die gemessenen Werte von bis zu 100 Kanälen in Analogform registriert.

EE 76 751 *für weitere Einzelheiten*

HF-Mikrovoltmeter

**Vertrieb: Claude Lyons Ltd,
76 Old Hall Street, Liverpool 3**

(Abbildung Seite 854)

Das Millivac - RF - Mikrovoltmeter MV-28B hat acht Spannungsbereiche von 1 mV bis zu 3 V Vollausschlag im Frequenzbereich 10 kHz...1,2 GHz bei einer niedrigsten ablesbaren Spannung von 300 μV. Die Messunsicherheit ist 3 Prozent zwischen 100 kHz und 60 MHz, 5 Prozent zwischen 50 kHz und 200 MHz und 10 Prozent über den Rest des Messbereiches. Mit dem hochohmigen Messkopffühler ist die Eingangsimpedanz ca. 2 ... 3 pF, 75 kΩ; ein 50-Ω-BNC-Abschlussadapter wird mitgeliefert, und als Sonderzubehör können u.a. kapazitive Spannungsteiler 100:1, N- und G. R.-Abschlüsse und ein T-Adapter Typ N bestellt werden.

Die Spiegelskala des Instrumentes ist sowohl in Volt wie auch dBm (auf 1 mW an 50 Ω bezogen) geeicht. Bis zu 30 mV ist die Anzeige fast in Effektivwerten, über 30 mV in Scheitelwerten; die Eichung gibt die entsprechenden Effektivwerte einer Sinuswelle. Es ist bemerkenswert, dass der Nutzbereich als effektivwertempfindliches Instrument durch Einsatz des kapazitiven Spannungsteilers 100 : 1 auf 3 V erweitert werden kann.

EE 76 752 *für weitere Einzelheiten*

Feuchtigkeitsmesser

Kappa Electronics Ltd, 159 Hammersmith Road, London, W.6

(Abbildung Seite 854)

Der Transistor-Feuchtigkeitsmesser AB66 ist ein tragbares batteriegespeistes Instrument zum Messen des Feuchtigkeitsgehaltes von Platten und Stoffen oder Substanzen von einer Seite, soweit sie eine annehmbar flache Oberfläche haben, die mit der "Fühlerplatte" im Boden des Instrumentgehäuses Kontakt machen kann.

Nach diesem Verfahren misst das Gerät den Einfluss des Stoffes auf das "hochfrequente" elektrostatische Feld, das von der Oberfläche der "Fühlerplatte" ausgestrahlt wird. Die Wirkung beruht auf den elektrochemischen Eigenschaften des Materials, unter denen der Feuchtigkeitsgehalt für viele Substanzen ein Hauptfaktor ist.

Das Instrument gibt je nach Zusammensetzung und Feuchtigkeitsgehalt des Materials eine Anzeige, die sich entweder zur Dielektrizitätskonstante oder—bei hochleitenden Substanzen—zu deren hochfrequenter Impedanz in Beziehung bringen lässt. Die Überlagerungs-Detektorschaltung gibt ein gutes Unterscheidungsvermögen für Dielektrizitätskonstante und Impedanz. Das Messergebnis wird auf der von Hand verstellten Skala in willkürlichen Einheiten angezeigt.

In Betrieb wird das Instrument auf eine Oberfläche des Materials gestellt und der geeichte Abstimmknopf verstellt, bis das eingebaute Messwerk einen maximalen Ausschlag gibt. Die Einstellung des Abstimmknopfes wird dann abgelesen und mittels einer mit Hilfe von Mustern bekannter Feuchtigkeit vorbereiteten Kurve oder Tabelle in den Feuchtigkeitsgehalt umgewandelt.

Beim Messen relativ dünner Platten werden dieselben entweder auf die flache Oberfläche des mitgelieferten Untersatzes oder—wo das bequemer ist—auf eine grosse, mit Metall hinterlegte Polystyrolschaumplatte gelegt.

Verhältnismässig dicke Materialien kann man an Ort und Stelle messen, jedoch hängt das von der gewählten "Fühlerplatte" und der Tiefe der Felddurchdringung ab. Man muss auch darauf achten, dass die Bedingungen an Ort und Stelle reproduzierbar sind.

EE 76 753 *für weitere Einzelheiten*

Röhrenvoltmeter
K.L.B. Electric Ltd, 335 Whitehorse Road, Croydon, Surrey
(Abbildung Seite 855)

Das PACO-Röhrenvoltmeter V.70 ist ein preisgünstiges und vielseitiges Messgerät mit sieben Gleichstrombereichen von 0...1500 V, sieben Wechselstrombereichen von 0...1500 V_{eff}, 0...4000 V_{ss}, einem Widerstandsbereich von 0...1 GΩ und einem Dezibelbereich von -6 dB...+66 dB. Das Gerät kommt komplett mit dreipoligem Messfühler. Als Zubehör sind ein HF-Kristallmesskopf und ein Hochspannungsmesskopf für unmittelbares Messen von Spannungen bis zu 60 kV lieferbar.

EE 76 754 *für weitere Einzelheiten*

Präzisionsvorgelege
Ancillary Developments Ltd, Blackwater Station Estate, Blackwater, Camberley, Surrey
(Abbildung Seite 855)

Bei Ancillary Developments Ltd wird in Kürze die Fertigung eines umfangreichen Sortimentes von Präzisionsvorgelegen und Zahnraduntersetzern für elektromechanische Servoeinrichtungen hoher Qualität anlaufen. Ausser vier üblichen soll es sich um die ersten britischen Baugruppen in der Internationalen Gehäusegrösse 08 handeln. Obwohl das ganze Programm den Forderungen der britischen RAE-Pflichtenblätter entspricht, sind die Anwendungsmöglichkeiten nicht auf das Flugwesen beschränkt.

Die verwendeten Werkstoffe sind durchweg korrosionsfest; für die Wellen sind wahlweise Gleit- oder Kugellager lieferbar. Zur Anpassung der Vorgelege und Untersetzer an die Servomotoren verschiedener Grösse sind Adaptorplatten lieferbar. Alle Einheiten können auf Wunsch entweder mit Zentrierung oder Klemmplatte geliefert werden.

Auch für die Ausgangswelle sind alternative Ausführungen lieferbar, z.B. glatt, mit Stirnzahnrad, oder als Keilwelle mit Gewinde.

Untersetzungen bis zu 1000 : 1 werden in Gehäusegrössen 08, 10 und 11, bis zu 5000 : 1 in Grössen 15 und 18 geliefert.

EE 76 755 *für weitere Einzelheiten*

Festkörper-Regler
Kent Precision Electronics Ltd, Vale Road, Tonbridge, Kent
(Abbildung Seite 855)

Kent Precision Electronics Ltd hat einen geeichten Festkörper-Regler für empfindliche Regel-, Schutz- oder Alarmsysteme entwickelt, der mit Gleich- oder Wechselstromeingängen durch Festkörperschalten Lasten bis in den Kilowattbereich steuert oder ein elektromagnetisches Relais betätigt, dessen isolierte Umschaltkontakte für vielseitige Aufgaben zur Verfügung stehen. Diese Instrumente können in allen Anwendungsgebieten das herkömmliche Kontaktmessgerät ersetzen, wobei durch moderne Festkörperschaltungen Zuverlässigkeit und lange, wartungsfreie Lebensdauer erreicht wird.

Die lineare Skala des Messwerkes ist in Einheiten des zu regelnden Parameters geeicht, und Einzel- oder Doppelmarkierungszeiger können auf die Werte eingestellt werden, bei denen Regelung erfolgen soll. Der Leistungsschalter oder die Relaiskontakte im Ausgang ändern ihren Zustand, wenn der Parametereingang über die eingestellten Messpunkte geht.

Das Instrument besteht grundsätzlich aus einem Transistor-Gleichstromverstärker mit niedriger Drift, der als Differenzdetektor zwischen einer geregelten Bezugsspannung und dem Eingangsparameter wirkt. Der Ausgang dieses Verstärkers speist eine Triggerschaltung, die ihrerseits eine Thyristor-Ausgangsschaltung für das Schalten der Last steuert, und zwar proportional oder ein-aus. Wenn die zu regelnde Leistung über der Nennleistung lieferbarer Thyristoren liegt, wird eine Thyristorschaltung niedrigerer Leistung zur Betätigung eines elektromagnetischen Relais oder Kontaktgebers mit geeigneter Nennleistung eingesetzt.

Normalerweise werden die Instrumente mit Standardrelais mit Kontakten für 250 V, 5 A~ ausgerüstet, sie lassen sich aber auch für andere Nennleistungen bemessen. Kundenwünschen entsprechend können Standard-Thyristorausgänge bis zu 600 V und 100 A geliefert werden. Speisung erfolgt aus einem Hilfsnetzgerät über einen Trenntransformator mit Regelung für Schwankungen von $\pm 20\%$. Der Temperaturkoeffizient ist für alle Bereiche innerhalb 0,05%/°C, und das Instrument ist für Betrieb in Umgebungstemperaturen bis zu 85°C geeignet.

EE 76 756 *für weitere Einzelheiten*

Schulungs-Spannungsteiler
Croydon Precision Co., Hampton Road, Croydon, Surrey
(Abbildung Seite 855)

Der Schulungs-Spannungsteiler' P.1 wurde für technische Lehranstalten und Gymnasien entwickelt, wo er die einfachen Schleifdraht-Spannungsteiler des Banjo-Types ersetzt oder ihnen folgt. Das robust gebaute Instrument hat eine Genauigkeit, die Einsatz für ernsthafte Aufgaben mit garantiert wiederholbaren Ergebnissen ermöglicht.

Der Spannungsmessumfang ist 0...1,1 V bei einer kleinsten Unterteilung von 0,001 V. Es sind zwei Messskalen vorgesehen: die erste hat 10 Schritte von je 0,1 V, und die zweite ist in Form eines Schleifdrahtes mit 0,1 V Skalenendwert und Unterteilung in 0,001 V. Die Messunsicherheit ist 0,1 Prozent oder 0,002 V, wobei immer der grössere Wert gilt. Eine eingebaute Eichschaltung ermöglicht Nacheichen unabhängig von der Skaleneinstellung; der Potentiometerstrom wird mittels eines mehrgängigen "Cropico"-Drehwiderstandes mit konstantem Kontaktwiderstand eingeregelt. Alle erforderlichen Schalter und Anschlüsse für Normalelement, Galvanometer, 1,5-V-Speisebatterie und die zu messende Spannung sind eingebaut.

Eins der Hauptmerkmale dieses Spannungsteilers ist die grafische Deckplatte, die in deutlicher Weise Verdrahtung und Widerstandsnetzwerke zeigt. Das erleichtert dem Auszubildenden, die Wirkungsweise der verschiedenen Bedienelemente usw. zu verstehen, und der Ausbilder braucht weniger Zeit zur Erläuterung der Arbeitsweise und Aufgaben des Instrumentes.

Als Zubehör wird ein Weston-Normalelement geliefert, das bei 20° C eine Absolutspannung von 1,01859 V hat. Das Element wird in einem robusten Kunststoffgehäuse mit isolierten Klemmen geliefert.

Als Detektor ist ein tragbares Drehspulgalvanometer mit geneigter Frontplatte lieferbar. Die Empfindlichkeit von 2 μA je Skalenteilung reicht aus, unabhängig davon, ob das Potentiometer nachgeeicht wird oder kurzgeschlossen ist.

EE 76 757 *für weitere Einzelheiten*

Beharrungsgedämpfte Motoren
Bowmar Instrument Ltd, Sutherland Road, London, E.17
(Abbildung Seite 856)

Bowmar kann nunmehr neue Typen der beharrungsgedämpften Motoren für 400 Hz in Grösse 08 liefern. Sie sind für Bezugsspannungen von 115 oder 26 V und Kundenwünschen entsprechenden Steuerspannungen zu haben. Typ IM83-002 hat eine Steuerwicklung für 33 V Spaltphase, ein Kurzschlussdrehmoment von 23,5 gcm und eine Lastgeschwindigkeit von 6500 UPM. Das Trägheitsmoment des Schwungrades ist 4,6 gcm² mit 245 Dyn/cm/Sek.

IM82-001 hat eine 33-V-Steuerwicklung mit Mittelabgriff, 21 gcm Kurzschlussdrehmoment und eine Leerlaufgeschwindigkeit von 5500 UPM. Das Trägheitsmoment des Schwungrades ist 1,6 gcm² mit 30 Dyn/cm/Sek Dämpfung.

Beide Motoren haben 30 cm lange Zuleitungen, Ausgangskeilwellen 120DP mit 13 Zähnen und einen Betriebstemperaturbereich von —65° C...+ 125° C. Diese Motoren können mit Getriebeuntersetzungen bis zu 2000 : 1 geliefert werden. Einheiten in Grösse 11 sind mit ähn-

lichen Wicklungen und Dämpfungsfaktoren zu haben.

EE 76 758 *für weitere Einzelheiten*

Leiterplatten-Bohrmaschine
Drill Service (Horley) Ltd, 89 Albert Road, Horley, Surrey

Auf Grund vieler Anfragen nach Bohrkopfeinheiten für Sonderzweckmaschinen für Leiterplatten bietet Drill Service nunmehr eine komplette Sonderzweckmaschine an.

Nach der bereits erprobten und bekannten Methode werden von oben in einer Schablone positionierte Plattenpakete von unten gebohrt.

Die zu bohrenden Platten werden in einer Vorrichtung gestapelt, deren Oberseite eine Schablone ist. Vorrichtung und Platten werden dann unter einen Schablonenstift gebracht, der in die Schablonenlöcher einrastet. Der Schablonenstift federt und übt zur Unterstützung des Rastens einen konstanten, leichten Druck aus. Sowie der Stift eingerastet ist, betätigt der Arbeiter einen Fussschalter und leitet dadurch den automatischen Bohrvorgang ein.

Durch Betätigung des Fussschalters wird der Vorschubmechanik des Bohrers Druckluft zugeführt, und gleichzeitig werden die Platten durch eine Klampe in der Bohrposition festgeklemmt.

Sobald der Bohrvorgang beendet ist, wird der Bohrer zurückgefahren und der Klemmdruck aufgehoben, um weiteres Positionieren zu ermöglichen.

Der Arbeiter hat beide Hände zum Positionieren der Schablone frei. Eine Bahn auf der Schablone, der der federnde Schablonenstift von Loch zu Loch folgt, kann den Arbeitsablauf beschleunigen.

Der Bohrkopf Modell 111 hat folgende Eigenschaften:

Stufenlos regelbare Geschwindigkeit bis zu 15 000 UPM;

Stufenlos regelbarer Vorschub.

Um saubere Löcher zu gewährleisten, werden Vorschub und Geschwindigkeit geregelt.

Durch einen Zeitschalter wird der Bohrer zwecks Räumen von Spänen in regelmässigen Zeitabständen ausgefahren, um Verstopfen bei schnellem Vorschub durch hohe Plattenpakete zu vermeiden.

Der Bohrkopf ist genau mit der Buchse und dem Schablonenstift ausgerichtet, so dass selbst ungelernte Arbeiter Pakete von 6 bis 8 Leiterplatten sehr schnell mit Hartmetallbohrern bearbeiten können, ohne zufälligen Bruch befürchten zu müssen. Diese Bohrvorrichtung ist besonders für Betrieb mit extralangen Hartmetallbohrern für Leiterplatten geeignet. Die Bohrer behalten ihre Schneidkante und Durchmesser für längere Arbeitsspannen, so dass Rüstzeit und unnötiger Ausschuss von Platten und Bauelementen beim Zusammenbau vermieden werden.

EE 76 759 *für weitere Einzelheiten*

Feuchtigkeitsmesser
The Wayne Kerr Laboratories Ltd, Sycamore Grove, New Malden, Surrey
(Abbildung Seite 856)

Als erstes in ihrem neuen sogenannten "Midge"-Programm kompakter Messgeräte kündigt Wayne Kerr den Feuchtigkeitsmesser MB100 an. Dieses in sich geschlossene Gerät braucht weder eingestellt, noch nachgeeicht zu werden und ist jederzeit betriebsfertig. In Verbindung mit einem preisgünstigen Messwertwandler gibt das Gerät direkte Anzeigen der relativen Feuchtigkeit von 20...100 Prozent in zwei Bereichen. Dichte Verpackungen können so ausgelegt werden, dass jede einen Messwandler mit Zuleitung zu einer Steckverbindung an der Aussenseite der Kiste enthält und der Zustand der für Lagerung oder Verschiffung in feuchter Umgebung verpackten Ausrüstung jederzeit ohne Auspacken überprüft werden kann.

Die Summenmessgenauigkeit ist einschliesslich Messwertwandler 5 Prozent. Die Messwertwandler können Temperaturschwankungen zwischen −40° C und +100° C aushalten, und ihre Leistungsfähigkeit wird weder durch lange Lagerung, noch Einsatz unter extremen Feuchtigkeitsbedingungen beeinflusst. Im wesentlichen ist die Messschaltung eine Übertragerbrücke, die bei 2 kHz arbeitet und elektronisch mittels eines gegengekoppelten Hochleistungsverstärkers abgeglichen wird. Quelle, Detektor und Batterien sind zusammen in einem Gehäuse (120 × 70 × 159 mm) untergebracht. Für Registrierzwecke usw. kann man einen Ausgangsstrom von 0...100 μA entnehmen.

EE 76 760 *für weitere Einzelheiten*

Metall-Oxyd-Silizium-Transistor
Mullard Ltd, Mullard House, Torrington Place, London, W.C.1

Bei der Ankündigung eines neuen Metall-Oxyd-Silizium-Transistors (MOST) gab Mullard Ltd auch bekannt, dass er bei Massenfertigung wesentlich billiger sein würde, als jedes zur Zeit in England lieferbare Element gleicher Leistung. Grossfertigung ist geplant, und in der Zwischenzeit können Mustermengen geliefert werden. Durch Anwendung der erprobten Planartechnik wird hohe Zuverlässigkeit gewährleistet.

Der MOST-Entwicklungstyp 95BFY hat einen Eingangswiderstand von einer Million Megohm (10^{12} Ω)—viel höher als der eines herkömmlichen Transistors oder selbst eines Feldeffekt-Transistors—und kann z.B. als hoch-niederohmiger Impedanzwandler oder als NF-Verstärker in Instrumentierungs- und Regelschaltungen, Elektronenrechnern und Fernmeldeausrüstungen Verwendung finden. Er ist auch für logische Schaltungen geeignet.

Da der MOST nach dem Majoritätsträgerprinzip arbeitet, wird er durch Strahlungen nicht beeinflusst und eignet sich daher für Satelliten und andere Raumfahrzeuge.

Die ersten Anwendungsstudien haben gezeigt, dass der Typ als Oszillator für Frequenzen bis zu 150 MHz, als Mischstufe mit 20 dB Mischverstärkung (bei Benutzung der Substrate als Anschluss für das zweite Signal), oder als "anodenmodulierte" Endstufe für Leistungen bis zu mehreren hundert Milliwatt bei Frequenzen bis zu 100 MHz eingesetzt werden kann. Seine Leistungsfähigkeit als Chopper ist besonders eindrucksvoll, da er bei Zerhackerfrequenzen bis zu 100 kHz keine "Verschiebungs"-Spannung und einen "Ein"-Widerstand von 400 Ω aufweist. Ausserdem hat er keine Speichereffekte, und da die Schaltzeit nur von der RC-Zeitkonstante der über den Strompfadwiderstand aufladenden Gate-Kapazität abhängt, kann er für Schaltzwecke eingesetzt werden.

Der 95BFY wird durch Diffundieren von zwei eng nebeneinander liegenden n+-Zonen (Source und Drain) in einen p-Siliziumträger gefertigt. Auf der Substrate wächst zwischen Source und Drain eine dünne Siliziumoxydschicht, auf die eine Gate-Elektrode aus Aluminium aufgedampft wird.

Durch Anlegen einer positiven Spannung an die Gate-Elektrode wird auf der Oberfläche der Substrate eine negative Ladung hervorgerufen, wodurch eine n-Zone entsteht, die einen Strompfad zwischen Source und Drain bildet. Wenn jetzt zwischen Source und Drain eine Spannung angelegt wird, fliesst ein Strom, der durch die Spannung am Gate gesteuert wird.

EE 76 761 *für weitere Einzelheiten*

Schwingungsanlage
Derritron Electronic Vibrations Ltd, 24 Upper Brook Street, London, W.1
(Abbildung Seite 857)

Das eine Schubkraft von 2721 kg entwickelnde Schwingungssystem besteht aus einem 30-kVA-Kraftverstärker, der für Testen mit regellosen und sinusförmigen Schwingungen ausgelegt ist, und einem Schwingungsgeber mit Steuerausrüstung. Der Verstärker hat eine maximale stetige Anodenverlustleistung von 50 kW, bei voller Leistung einen Frequenzbereich von 15 Hz...5 kHz, und vollen Strom gibt es bis zu 5 Hz herunter. Die neusten English-Electric-Röhren mit Verdampfungskühlung werden benutzt, Brummen und Rauschen sind—auf die Höchstleistung bezogen—mindestens −70 dB.

Die Treibspule des Schwingungsgebers VP.6000 für 2721 kg Schub ist mit einem Kühlungssystem ausgerüstet, das einen äusserst steifen Tisch- und Spulenzusammenbau ergibt. Üblicherweise wird für die Kühlung destilliertes Wasser verwendet, für Klimakammern unter besonders rauhen Bedingungen kann man jedoch Ölkühlung benutzen. Technische Kurzdaten für diesen Schwingungsgeber sind 2721 kg Vektorschub, Frequenzbereich etwa 5 Hz...3 kHz nutzbar bis zu 5 kHz mit einer ersten grösseren

störenden Resonanz bei 2,5 kHz. Das niedrige Tischgewicht von 32 kg erlaubt unbelastete Tischbeschleunigungen bis zu 85 g.

EE 76 762 für weitere Einzelheiten

Spannungsabhängige Relais
Newton Brothers (Derby) Ltd, Alfreton Road, Derby
(Abbildung Seite 857)

Die spannungsabhängigen Relais TVS der Firma wurden als Schaltmittel für vorgegebene Spannungspegel entwickelt und können in Schutzschaltungen gegen Unter- oder Überspannung, in denen hohe Genauigkeit erforderlich ist, eingesetzt werden.

Die Serie überdeckt Schaltspannungen zwischen 10 und 220 V— bei ±1% Empfindlichkeit und 1% Differential. Die Ansprechzeit ist 10...15 ms; ist jedoch Auslösung durch Überlastung bei Einschwingvorgängen unerwünscht, kann eine Verzögerung von ein paar Millisekunden bis zu zwei Minuten vorgesehen werden.

Die Bauelemente sind mit 10poligen Sockeln ausgerüstet und lassen sich ohne Schwierigkeiten in Steuergeräte einbauen; sie sind mit zwei normalerweise geschlossenen und einem normalerweise offenen Hochleistungskontakt bestückt.

EE 76 763 für weitere Einzelheiten

Bürstenlose Gleichstromservomotoren
Vactric Control Equipment Ltd, Garth Road, Morden, Surrey
(Abbildung Seite 857)

Diese neuentwickelten bürstenlosen Gleichstromservomotoren in Grösse 11 wurden herausgebracht, um den Bedarf an Motoren mit viel längerer Lebensdauer als herkömmliche Gleichstrommotoren zu decken, und sind besonders für Anwendungszwecke geeignet, in denen Funkenstörung eine Rolle spielt. Sie haben weder Bürsten noch Kollektor, und die Abwesenheit von Bogenbildung und Bürstenabnutzung führt zu längerer Lebensdauer. Sie haben eine Spezialwicklung für Aufnahme der von einem getrennten elektronischen Baustein abgegebenen Wellenformen.

Der Motor ist durchgehend aus korrosionsfesten Werkstoffen konstruiert und arbeitet im Umgebungstemperaturbereich —65° C...+85° C und Höhen bis zu 18 280 m.

Der elektronische Baustein enthält einen Multivibrator, der über geeignete logische Schaltungen eine Reihe von Flip-Flops steuert, deren Leistungstransistoren die Motorwicklungen speisen. Er hat ein Volumen von ungefähr 33 m³ und kann dieselben Umgebungsbedingungen aushalten wie der Motor. Der Baustein kann in ein Zylindergehäuse Grösse 11 vergossen mit Lötstiften oder —für Einsatz unter weniger strengen Bedingungen—in Steckkartenform geliefert werden.

EE 76 764 für weitere Einzelheiten

Messbrücke
K.L.B. Electric Ltd, 335 Whitehorse Road, Croydon, Surrey
(Abbildung Seite 857)

Die Messbrücke M.300 für Bauelemente ist ein vielseitiges und preiswertes Instrument mit vier Kapazitätsbereichen von 10 pF...1000 μF und vier Widerstandsbereichen von 0,05 Ω...100 MΩ, sowie einer Einrichtung zum Testen auf Kondensator-Reststrom mit regelbarer Testspannung bis zu 500 V—. Verhältnismessungen in den Bereichen 0,05 : 1 und 20 : 1 können zwischen jeden beliebigen zwei Kondensatoren, Spulen oder Widerständen durchgeführt werden. Dieser Test ist für das Abgleichen von Bauelementen gegen ein Normal bestimmt.

EE 76 765 für weitere Einzelheiten

Umschaltbare Zirkulatoren
The Marconi Co. Ltd, Chelmsford, Essex
(Abbildung Seite 858)

Die Marconi Company hat einen neuen umschaltbaren Zirkulator mit drei Toren für Hohlleiter Nr.15 entwickelt, der den Frequenzbereich 7,75...8,5 GHz überstreicht. Schnelles Umschalten der Leitungsrichtung wird auf elektromagnetischem Wege erreicht.

Der neue Zirkulator wurde für Verwendung in Mikrowellen-Nachrichtenausrüstungen entwickelt und bietet in bezug auf Zuverlässigkeit, Schaltgeschwindigkeit und Raumeinsparung wesentliche Vorteile gegenüber den bisherigen Schaltmethoden mit mechanischen Einrichtungen. Er ist für alle Zweiweg-Hohlleiterschaltzwecke geeignet; ein typisches Beispiel sind Weichen, die Mikrowellensender und -empfänger an ihre Antennen anschliessen.

Dieser umschaltbare Zirkulator ist eine Ausführung des dreiarmigen Standard-Verbindungszirkulators, in der jedoch das magnetische Vorspannungsfeld elektromagnetisch angelegt und für Richtungsänderung umgekehrt wird. Der Elektromagnet und Zirkulatorkörper sind so konstruiert, dass das Magnetfeld schnell mittels eines normalen Transistor-Schalters umgekehrt werden kann. Die Stromversorgung des Elektromagneten regelt die Schaltzeit des Zirkulators. Bei 25 W zugeführter Leistung lassen sich Schaltzeiten von unter 0,5 ms erzielen.

EE 76 766 für weitere Einzelheiten

FS-Nutzfeldmesser
Lab-Craft Ltd, Gainsborough Road, Woodford Bridge, Essex
(Abbildung Seite 858)

Der Nutzfeldmesser 215 ist unter Berücksichtigung neuster technischer Erkenntnisse und hohen Anforderungen entsprechend konstruiert. Alle Verstärkerstufen sind mit Spanngitterröhren bestückt, und die Konstantstromversorgung macht Drift und Verstärkungsschwankungen über den Gesamtbereich

der Netzschwankungen vernachlässigbar klein.

Der Kern des Instrumentes ist ein Standard-UKW-Tuner, der einen ZF-Verstärker mit Verstärkungsregelung speist. Der geregelte Strom einer der ZF-Verstärkerröhren wird auf einem Messgerät angezeigt. Die direkt in Mikrovolt und Millivolt geeichte Skala gibt eindeutig ablesbare Eichungen vom Rand- zum Überdeckpegel, ohne dass getrennte Einsteckabschwächer erforderlich werden. Ansteigende Signalpegel führen zu einer Reduktion des durch das Messwerk fliessenden Stroms. Diese Anordnung schützt das Messwerk gegen zufällige Überlastung.

Die Feinabstimmung wird für jeden Kanal einmal eingestellt und stellt sich danach jedes Mal, wenn der Tuner geschaltet wird, automatisch ein. Diese Einrichtung sowie die nicht reihenmässige Anordnung der Spulen machen Umschalten von Band I auf Band III für die meisten Gegenden momentan.

Wegen seiner Hochempfindlichkeit mit Anzeige bis zu 10 μV herunter kann das Modell in extremen Randgebieten benutzt werden, wo solch ein Instrument besonders agebracht ist. Die Antennenbuchse ist hinten am Gerät angeordnet, und bei Benutzung einer kurzen Teleskopantenne lässt sich der erforderliche Antennengewinn überall relativ einschätzen.

EE 76 767 für weitere Einzelheiten

Selbstklemmerige Steckverbindungen
Vertrieb: Nutec Electronics Ltd, Mercantor House, East Street, Shoreham-by-Sea, Sussex
(Abbildung Seite 858)

Die von dem Schweizer Ingenieur Leon Mouttet entwickelten Lemo-Steckverbindungen haben eine patentierte Rasteinrichtung, die dem unbeabsichtigten Auseinanderfallen durch Schwingungen oder Zug am Kabel des Steckers oder der Kupplung vorbeugt.

Die Kontaktoberflächen sind stark vergoldet, um sehr niedrigen und konstanten Kontaktwiderstand zu gewährleisten, und es finden durchweg nur Isolierstoffe höchster Qualität Verwendung.

Leo-Steckverbindungen kommen in drei Grundtypen: koaxial, zwei- und dreiachsig für Hochspannung oder mehrpolig. Jeden Typ gibt es in sieben Grössen, die alle für Mehrfach-Variationen zusammengebaut werden können.

Ausser den Standardmodellen und -kombinationen gibt es Hochspannungstypen, wasser- und vakuumdichte Steckverbindungen, Sonderausführungen für hohe und niedrige Spannungen, koaxiale Stecker und Systeme für das Fernsehen.

Die verschiedenen Zangen-Klemmeinrichtungen können eine breite Auswahl von Kabeln aufnehmen.

EE 76 768 für weitere Einzelheiten

Subminiatur-Messwertgeber

Associated Engineering Ltd, Cawston, Rugby
(Abbildung Seite 858)

Für Weg- und Schwingungsmessungen unter ungünstigen Umgebungsbedingungen wurde ein neuer induktiver Subminiatur-Weggeber entwickelt.

Der in den zwei abgebildeten Grössen hergestellte Messwertgeber kann mit Trägerfrequenzen bis zu 300 kHz eingesetzt werden, so dass man Schwingungen oder Bewegung irgendwelcher metallischer Stoffe bei Frequenzen von 0...60 kHz nachweisen kann. Die beiden Messwertgeber haben einen Messumfang von 0,25 mm, können Umgebungstemperaturen bis zu 150° C ohne merkbare Änderung der Empfindlichkeit aushalten und bei geeigneter Montage ohne Ausfall bis zu 2000 g beschleunigt werden.

Abmessungen und Gewicht der Messwertgeber einschliesslich der Anschlüsse für gedruckte Schaltungen sind:

Grosses Modell: 10,9 mm lang, 6,35 mm Durchmesser und 7 g Gewicht;

Kleines Modell: 10,2 mm lang, 4 mm Durchmesser und 5 g Gewicht.

·Die Messwertgeber werden für Montage auf Wunsch mit dem links im Foto sichtbaren Gewindering hergestellt. Sonst werden sie wie in der Mitte und rechts abgebildet geliefert und entweder festgeklemmt oder mit geeignetem Kitt oder Epoxydharz befestigt.

EE 76 769 *für weitere Einzelheiten*

Lötkolben

Vertrieb: Lugton & Co. Ltd,
209-212 Tottenham Court Road, London, W.1
Die in Kanada hergestellten Imperial-Ungar-Lötkolben können nunmehr von Lugton & Co bezogen werden.

Die Haupteigenschaft dieses Lötkolbens ist seine Anpassungsfähigkeit, da alle Teile schnell ausgetauscht werden können. So kann man z.B. einen 25-W-Kolben mit kleiner keilförmiger Lötspitze innerhalb von Minuten in einen 40-W-Kolben mit langer, verjüngter Keilspitze umwandeln. Ein breites Lötspitzen-Sortiment ist lieferbar.

Die in rostfreien Stahl eingeschlossene Heizpatrone gewährleistet lange Lebensdauer, beseitigt Zundern und gibt maximale Wärmeübertragung. Farbenkennzeichnung gestattet sofortiges Unterscheiden von Kolben unterschiedlicher Leistung.

EE 76 770 *für weitere Einzelheiten*

Wellenformgenerator

Feedback Ltd, Crowborough, Sussex
(Abbildung Seite 859)

Der Wellenformgenerator TWG 300 ist ein vielseitiges Instrument, das für die Untersuchung der Leistung von Regelsystemen bestimmt ist.

Die Arbeitsweise des TWG 300 beruht auf Integrationsschaltungen und Schaltvorrichtungen, die Dreieck- und Rechteckwellen erzeugen. Die Dreieckwellenform wird in einem Diodennetzwerk in eine Sinuswelle konstanter Amplitude umgeformt, die über den gesamten Frequenzbereich einen niedrigen Oberwellengehalt hat. Ein weiteres Netzwerk erzeugt die abgekappte Dreieckwellenform, bei der durch Abschneiden der Spitzen von der Dreieckwelle abwechselnd Dachflächen oder -schrägen entstehen. Die Abkappschaltung ist regelbar, so dass die Neigung der Dachschräge variiert werden kann, die konstante Amplitude jedoch beibehalten wird, woraus sich ein Verhältnis regelbare Dachschräge : Rest ergibt.

Der Frequenzumfang ist 0,001...1000 Hz in sechs 10:1 Teilbereichen mit kontinuierlicher Einstellung in jedem Bereich. Die Skalen sind in Frequenz und Periode geeicht und überlappen an beiden Enden um 20 Prozent, wodurch kontinuierliches Überstreichen gewährleistet ist. Die Skalengenauigkeit ist besser als 0,1 Prozent. Für Frequenzeinstellung gegen Zähler oder Frequenznormale ist die Unterscheidung besser als 1 Prozent der Frequenz. Die gewählte Wellenform steht gleichzeitig an zwei Ausgängen zur Verfügung: entweder mit konstantem Pegel von 40 V_s bei 20 mA Höchststrom und 10 Ω Ausgangsimpedanz, oder über einen kontinuierlich regelbaren 2-kΩ-Abschwächer, der für 0...40 V geeicht ist, und eine Reihe von Schaltern, die ×1 — ×0,1—Aus geben. Die Ausgangsamplitude ist unabhängig von der Frequenzeinstellung konstant. Ausgänge werden zwischen "stromführenden" und "gemeinsamen" Klemmen entnommen. Die am Gestell liegende Erd- oder Masseklemme wird üblicherweise mit der "gemeinsamen" Klemme verbunden. Nach Entfernung dieser Verbindung kann das Instrument durch externe Stromversorgungen auf ein erdfreies Potential von 250 V gebracht werden. Vorsorge muss getroffen werden, um der externen Stromversorgung zur Vermeidung von Brummeinkopplung einen niederohmigen Pfad zu geben.

Monitorausgänge für Dreieck- und Rechteckwellen haben eine konstante Amplitude von 5 V_s und 20 kΩ Quellenimpedanz. Diese Ausgänge sind mit der Hauptwellenform synchronisiert.

EE 76 771 *für weitere Einzelheiten*

Zeitbasis und Zähler

Specto Avionics Ltd, Feltham, Middlesex
(Abbildung Seite 859)

Ein Universalzeitmessgerät zum Messen von Frequenz, Periode, Zeit und Impulsbreite wurde von Specto Avionics Ltd entwickelt. Es ist für Messungen ausgelegt, die besonders hohe Genauigkeit verlangen, und wird vor allem in Forschungsinstituten, Universitäten und Industrie von Nutzen sein.

Das Instrument ist eine Kombination von zwei Standardgeräten der Specto Avionics, und zwar dem Dezimikrosekunden-Chronometer 222 und der Zeitbasis 233. Die langfristige Genauigkeit der kombinierten Geräte ist ±2 × 10⁻⁷, die kurzfristige ±1 × 10⁻⁸. Wie alle anderen Geräte dieser Serie ist die Zeitbasis 233 mit Silizium-Transistoren bestückt, in modernster Druckschaltungstechnik und leichter, modularer Konstruktion ausgeführt und kann entweder in Gestelle eingebaut oder einzeln auf dem Labortisch benutzt werden. Die Abmessungen der Zeitbasis sind 450 mm × 89 mm × 368 mm, und wenn sie mit dem Zehntelmikrosekunden-Chronometer zusammen ins Gestell eingebaut wird, ist die kombinierte Höhe 222 mm.

Die technischen Kurzdaten des kombinierten Instrumentes sind:
Zeit- und Impulsdauermessungen: 0,1...10⁸ s
Direktes Eingangssignal: ±1 V_s... ±5 V_s Impulse bei 1 kΩ Last
Eingangssignal über Abschwächer: ±5 V_s ... ±25 V_s Impulse bei 1 kΩ Last
Frequenz: 30 Hz ...20 MHz
Periode: 1 ... 10⁸ Ereignisse; Mindestperiodendauer 0,05 μs
Verhältnis: f_1 = 20 MHz max; f_2 = 15 MHz max
Eingangsempfindlichkeit: 100 mV$_{eff}$ ... 5 V$_{eff}$
Eingangsimpedanz: 10 kΩ
Temperaturbereich: −20°C ... +40°C.

EE 76 772 *für weitere Einzelheiten*

Bürstenlose Drehmelder

Elliott-Automation Ltd, Century Works,
Lewisham, London, S.E.13
(Abbildung Seite 859)

Die Servo Components Division der Elliott-Automation kündigt einen neuen bürstenlosen Drehmelder für Verwendungszwecke an, die hohe Bauelementzuverlässigkeit erfordern.

Die neuen bürstenlosen Elliott-Ellsyn-Drehmelder sind als Steuergeber und -transformatoren Grösse 11 lieferbar. Sie haben viele Vorteile gegenüber herkömmlichen Bauelementen, haben aber ähnliche elektrische und mechanische Kenndaten beibehalten. Bürsten und Schleifringe wurden durch Miniatur-Hilfstransformatoren ersetzt, was Bürstenkontaktprobleme beseitigt und die Reibungsverluste auf ein Viertel der in herkömmlichen Drehmeldern üblichen reduziert. Diese neuen Bauelemente sind daher besonders für Anwendung in der Fluginstrumentierung geeignet.

Die Drehmelder sind für Betrieb in Temperaturen bis zu +125° C mit maximalen Fehlern von ±14 Bogenminuten geeignet. Weiterentwicklungen zur Verbesserung dieser Leistung und Erhöhung der Betriebstemperaturgrenze auf +150° C sind im Gange.

Sorgfältige Konstruktion ermöglichte die Längenmasse der Steuertransformatoren· denen der herkömmlichen Aus-

führungen gleich zu halten. Der Steuertransformator, der mit einer höheren Leistung arbeitet, ist 12,7 mm länger als der Standard-Drehmelder.

Sowohl die 115-V- als auch die 26-V-Typen des lieferbaren Ellsyn-Programmes bürstenloser Drehmelder sind so konstruiert, dass sie gegen herkömmliche Typen auswechselbar sind.

EE 76 773 *für weitere Einzelheiten*

Image-Orthikon
English Electric Valve Co. Ltd, Chelmsford, Essex
(Abbildung Seite 859)

English Electric Valve Co Ltd hat die Neuentwicklung eines Werkstoffes für die Speicherplatten ihrer Image-Orthikons bekanntgegeben.

Der mit ELCON bezeichnete Werkstoff bringt die garantierte Lebensdauer der Image-Orthikons mit mindestens 750 Stunden auf ein neues Hoch. Protokolle über Muster der Nullserie lassen Betriebslebensdauern von 3000 bis 5000 Stunden erwarten.

Die ELCON-Speicherplatte erreicht auch völlige Abwesenheit von Bildbeibehaltung, Konstanz der Kontrastwiedergabe, wesentlich reduziertes Abfallen der Empfindlichkeit und sofortiges Einschalten.

Die ELCON-Speicherplatte führt auch ein neues Arbeitsprinzip ein, nach dem anstelle der ionischen nunmehr eine elektronische Leitung die Übertragung des Ladungsmusters von der Bildseite auf die Strahlseite übernimmt. Ionenwanderung ist daher praktisch beseitigt, folglich gibt es keine bedeutsame Leistungsverschlechterung mit Zeit. Empfindlichkeit und Konstanz der Kontrastwiedergabe (Gamma) bleiben daher unveränderlich, und "Bildkonservierung" kommt während der Lebensdauer nicht vor.

EEV betont, dass der neue Werkstoff keine verborgenen Untugenden besitzt und dass er eine sehr zufriedenstellende Lösung der bei jedem normalen Image-Orthikontyp auftretenden Probleme darstellt.

Das volle EEV-Sortiment von 3″-Image-Orthikons ist mit der ELCON-Speicherplatte ausgestattet und in Serienfertigung.

In Kürze wird es auch möglich sein, das 4¼″ Orthikon mit der neuen Speicherplatte zu bemustern; Serienfertigung wird Anfang 1965 anlaufen.

Die Bezeichnung ELCON wurde von den Eigenschaften des neuen Speicherplattenwerkstoffes abgeleitet, d.h. ELectronic CONducting (elektronisch leitend), um einen Unterschied gegenüber den üblicherweise verwendeten ionischen Leitern zu machen.

EE 76 774 *für weitere Einzelheiten*

Zusammenfassung
der wichtigsten Beiträge

Ein linearer Transistor-Schwärzungsmesser mit vier Dekaden   von E. Gordy, P. Hasenpusch und G. F. Sieber

Zusammenfassung des Beitrages auf Seite 808-812
Eine lineare Festkörper-Schwärzungsmesserschaltung mit einem Arbeitsbereich von null bis vier für optische Dichte wurde mit Dynoden-Spannungsgegenkopplung in Kombination mit Linearitätskorrektur mittels vorgespannter Dioden entworfen. Die Linearität des Ausgangssignals in Bezug auf optische Dichte ist ±0,5 Prozent. Je eine Ausführung mit einem Germanium-Transistor und einem Silizium-Transistor wird beschrieben.

Ein versuchsmässiges elektronisches Zündungssystem   von W. Alexander und P. Sample

Zusammenfassung des Beitrages auf Seite 813-816
Eine versuchsmässige Untersuchung der Vor- und Nachteile eines elektronischen Zündungssystems für Kraftfahrzeugmotoren wird beschrieben. In der angewandten Methode wird ein Kondensator in die Primärwicklung eines Umspanners entladen, und das Schalten erfolgt mittels eines steuerbaren Siliziumgleichrichters. Es wird gezeigt, dass eine erhöhte Motorenleistung erreicht werden kann.

Die Demodulation eines mit zwei Tönen modulierten FM-Signales   von G. May

Zusammenfassung des Beitrages auf Seite 816-819
Das Modulationsverfahren und das zusammengedrängte Spektrum der Seitenfrequenzen, das als Resultat des Frequenzmodulierens eines Trägers mit zwei Tönen auftritt, wurde bereits in mehreren Artikeln beschrieben (1, 2, 3, 4). Überraschenderweise hat der umgekehrte Prozess der Demodulation,

in dem das FM-Signal in einem typischen Diskriminator in eine amplitudenmodulierte Welle umgewandelt wird, wenig Aufmerksamkeit erregt.

Die Aufgabe dieses Beitrages ist, zu zeigen, wie diese Seitenfrequenzen — vor allem die Kombinationsfrequenzen—während der Demodulation benutzt werden.

Ein elektronisches Spektrometer für Ton- und Tiefstfrequenzen   von D. Llanwyn Jones

Zusammenfassung des Beitrages auf Seite 820-823

Die Theorie eines einfachen Spektrometers für Ton- und Tiefstfrequenzen wird entwickelt und die Selektivität und Stabilität der Schaltung abgeleitet. Eine praktische Schaltung wird gegeben und in Bezug auf die Ergebnisse der theoretischen Analyse diskutiert.

Decodieren von SECAM-Signalen für Einstrahlsystem-Bildröhren   von P. Cassagne und M. Melchoir

Zusammenfassung des Beitrages auf Seite 824-827

Die Verarbeitung von SECAM-Signalen in Fernsehempfängern mit Einstrahlsystem-Bildröhren wird untersucht. Eine neue Schaltung für direkte Umwandlung in CCS-Signale ohne Demodulation der Buntsignale wird beschrieben, die eine sehr einfache und mindestens ebenso effektive Lösung darstellt wie die mit NTSC-Signalen benutzte.

Ausserdem gestattet diese Anordnung die Konstruktion eines hochwertigen SECAM-Transcoders.

Verbesserung der Leistung von Kaltkathoden-Zählrohren   von F. Szlavik und L. Garamszegi

Zusammenfassung des Beitrages auf Seite 828-831

Einige Methoden zur Verbesserung der Leistung von gasgefüllten Zählrohren bei hohen Zählraten werden besprochen und Schaltungen beschrieben, mit deren Hilfe die Rate der noch zuverlässig gezählten Impulse weit über die Grenzfrequenz im Pflichtenblatt des Herstellers oder die mit in der Literatur beschriebenen Anordnungen erreichbare hinaus erhöht wird.

Auf diese Weise konnten für Zählrohre mit 40 bzw. 20 Elektroden Zählfrequenzen von 100 kHz und 2 MHz erzielt werden.

Harmonische Spannungen im Ausgang von Schaltungen mit steuerbaren Siliziumgleichrichtern   von R. Smith

Zusammenfassung des Beitrages auf Seite 832-837

In diesem Beitrag werden die im Ausgang von Schaltungen mit steuerbaren Siliziumgleichrichtern vorhandenen harmonischen Spannungen mathematisch abgeleitet. Die im Fernmeldewesen benutzten Kurven für äquivalente, auf 800 Hz bezogene Restwelligkeit (bewertete Restwelligkeit) werden ebenfalls gegeben. Eine Tabelle der Restwelligkeit-Bewertungsfaktoren ist einbegriffen.

Eine frequenzkompensierte Phasenverzögerungsschaltung   von A. De Sa und L. Molyneux

Zusammenfassung des Beitrages auf Seite 838-839

Diese Schaltung erzeugt eine Phasenverschiebung, die für jeden Wert zwischen 40° und 280° eingestellt werden kann, sich jedoch wenig mit der Frequenz ändert. Im Grunde genommen ist es eine frequenzkompensierte Zeitverzögerungsschaltung. Die Schaltung kann auch für Frequenzmultiplikation Anwendung finden.

Stehwellen an Übertragungsleitungen   von W. S. Metcalf

Zusammenfassung des Beitrages auf Seite 840-842

Es wird eine Analyse der Übertragungsleitungen gegeben und mittels abgeleiteter Formeln eine Tabelle aufgestellt, die die drei Stehwellenverhältnisse, Spannung, Reflexionsfaktor, reflektierte Leistung, das Verhältnis der vorlaufenden zur reflektierten Leistung, sowie die Übertragungsverluste gibt.

Ein einfaches und zweckdienliches 0,2%-Gleichstromvoltmeter   von M. R. Osborne

Zusammenfassung des Beitrages auf Seite 843-845

Ein beschriebenes Gerät kann Spannungen bis zu 20 V– mit einer Unsicherheit von ±0,2 Prozent über 0,5 V und ±1 mV darunter messen. Es ist ein Vergleichsgerät, in dem die unbekannte Spannung einer von einem Präzisions-Wendelpotentiometer und einer Bezugsdiode abgeleiteten bekannten Spannung gleichgesetzt wird. Der Komparator hat eine AVR-Charakteristik, die eine Empfindlichkeitsregelung überflüssig macht. Das Instrument füllt in Genauigkeit und Preis eine Lücke zwischen dem herkömmlichen Drehspulinstrument und dem Digitalvoltmeter.

Ein langsamer Schalter   von K. F. Knott

Zusammenfassung des Beitrages auf Seite 846-848

Der Beitrag beschreibt das Schalten eines Transistors nach einer Methode, die ohne Schwierigkeiten Regelung der Schaltzeit über mehrere Sekunden ermöglicht. Ausserdem wird ein Schmitt-Trigger mit leicht regelbarer Hysteresis beschrieben.