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Commentary

THE 'Metric System' of weights and measures is to be adopted by Britain according to a written answer given in the House of Commons on 24 May by Mr. Douglas Jay, President of the Board of Trade. According to Mr. Jay countries using that system now take more than one half of our exports and the total proportion of world trade conducted in terms of metric units will no doubt continue to increase. His statement went on to say "Against that background the Government consider it desirable that British industries on a broadening front should adopt metric units, sector by sector, until that system can become in time the primary system of weights and measures for the country as a whole."

It was also stated that a necessary condition for advances in this field will be the provision of metric standards, whenever possible internationally recognized, which will enable particular sectors of industry to work in metric units. To this end the Government has requested the British Standards Institution to pay special attention to this work and to press on with it as speedily as possible. The Government recognizes that this will place a considerable burden on the B.S.I. and will necessitate larger grants-in-aid to the Institution.

According to Mr. Jay's statement, the Government hope that within ten years the greater part of the country's industry will have effected the change and to this end a small standing committee of representatives of Government Departments and industry will be established to facilitate the removal of obstacles and to keep under constant review the progress which is being achieved. It is also proposed to keep in touch with Commonwealth Governments on this matter.

Whether the advantages, mainly in terms of increased exports, of changing to a metric system are real or illusory is a matter for conjecture, but before commenting on some of the issues involved it may perhaps be pointed out that there is in fact, no such thing as 'the metric system'. There are metric units for length and weight but the needs of engineering, and even more certainly of science, go somewhat beyond the realms of linear measurement, and here there is as much chaos and confusion among the 'metric' countries as there is among the 'non-metric' countries. There are a number of systems based on metric units, such as the c.g.s. system, the m.k.s. system and the Giorgi system. All of these systems have enjoyed considerable popularity at one time or another, but none has found anything like universal acceptance; perhaps the latest contender, the S.I. system, will be more successful. It may also be pointed out that even in the realm of linear measurement there is no great measure of conformity in engineering specifications between the European countries.

The cost to this country of going completely 'metric' will be considerable. There are the tangible costs such as modifications to drawings, machine tools, taps, dies, spanners, scales, micrometers, etc.; there are also intangible costs such as an inevitable slowing of production during the transitional period; a fitter who has worked all his life in inches and 'thous' cannot overnight re-orientate his mind to millimetres and fractions thereof. It is necessary therefore that the gain from such a change should be correspondingly great and the evidence to this effect seems rather inconclusive. Do the industrial countries of the world really find these different units of measurement a handicap to trade? The U.S.A., the world's greatest exporter, does not seem to find the barrier all that great and Italy does not seem to experience undue difficulty in selling motor scooters to this country, complete with metric threads! Conversely, will our own motor car industry find it any more difficult to export its products to the U.S.A. after we have changed to metric threads and dimensions? It is certainly to be hoped that it will not. So far as the electronic industry is concerned the U.S.A. is the world's largest manufacturer and user of such equipment and will, it seems likely, continue to set the standards, whatever units of measurement the rest of the world may use. An example of this is the 0·1in grid for printed circuits.

Against all this, a strong case can be made for the desirability of standardization in industry and a system based on metric units is by far the most rational and easy to handle. The argument is even more forceful in the case of a science based industry such as electronic engineering. For example, it is obviously incongruous to dimension an aerial in feet and inches when its size has been determined by wavelengths measured in metres; or to dimension a field strength diagram in miles and then mark in the field strengths in millivolts per metre.

If world standardization of a measuring system can be achieved then one based on metric units undoubtedly offers the best long term hope of success but it is not a thing to be undertaken lightly in the present economic position of this country. It must be remembered too that excellent though standardization is in many ways it can, if carried too far, lead to wretchedly dull uniformity and conformity.

The Sing-Around Velocimeter and its Use in Measuring the Size of Turbulent Eddies in the Sea

By D. J. Dunn*, Ph.D.

Several methods of measuring variations in the velocity of flow of turbulent water are discussed. A brief description of the sing-around velocimeter is given and it is shown that it, too, can be used to measure flow velocity fluctuations.

(Voir page 492 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 499)

MEASUREMENTS of the velocity of flow at a fixed point in the sea show that fluctuations can occur having a very wide range of period, varying from 0·01sec as reported by Patterson[1] to a number of days as reported by Swallow and Hamon[2]. It therefore appears that the measurable periods of fluctuations are only limited by the frequency response of the measuring instrument and the duration of the record.

For studying high frequency fluctuations, the hot wire technique, used in atmospheric work, can be adapted to work underwater. This system has been described by Patterson[1]. Low frequency fluctuations can be measured by mechanical systems such as sets of revolving cups or propellors as used on ships' logs. These latter systems can only measure fluctuations of the order of several seconds and are rather bulky to handle.

In 1956 Bowden and Fairbairn[3] introduced an electromagnetic flowmeter which enabled measurements of both the vertical and longitudinal components of the eddy velocity to be made. In their system the measuring head of the meter produces a local magnetic field and the water flowing through this field has an e.m.f. induced in it. The potential gradient set up in the water is measured by two pairs of electrodes, so that the two components of the velocity fluctuations can be recorded simultaneously. The frequency response of this equipment falls to one half at ½c/s.

An alternative approach is to use a sing-around velocimeter to measure the velocity fluctuations of the turbulent eddies.

The Sing-Around Velocimeter

Sing-around velocimeters have been used for measuring the speed of sound in liquids for several years[4,5,6]. The principle of operation is well known but is included here for completeness. A short pulse from a blocking oscillator is fed directly to the transmitting transducer, as shown in Fig. 1. From this an acoustic pressure pulse travels through the medium whose properties are being investigated, to a receiving transducer in a time l_1/C_0 where l_1 is the path length between the two transducers and C_0 is the velocity of sound in the medium. The received pulse is amplified, reshaped and used to synchronize the original pulse-forming circuit. This process is continuous.

If τ_e represents the delay in the electrical circuits, the total time between two pulses is:

$$\tau_1 = \tau_e + l_1/C_0 \quad \text{(1)}$$

normally:

$$\tau_e \ll l_1/C_0$$

and so:

$$\tau_1 \simeq l_1/C_0 \quad \text{(2)}$$

If the system is allowed to transmit signals in a medium which is moving with a velocity V along the direction of measurement, then the time between pulses becomes:

$$\tau_1 = \frac{l_1}{C_0 + V} \quad \text{(3)}$$

The corresponding pulse repetition frequency is given by:

$$f_1 = 1/\tau_1 = \frac{C_0 + V}{l_1} \quad \text{(4)}$$

Thus the sing-around frequency f_1 is directly proportional to the velocity of flow of the medium. A single

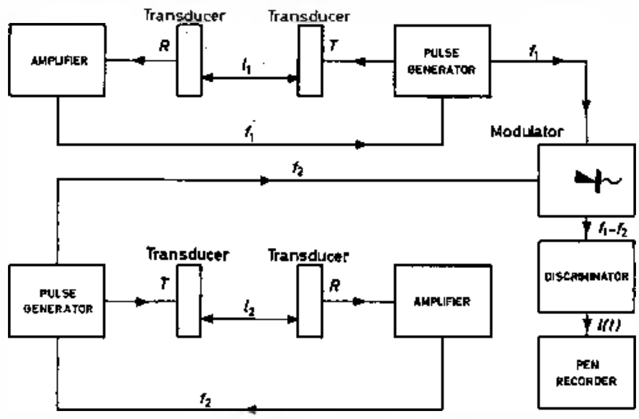

Fig. 1. The sing-around principle

velocimeter can therefore be used to measure the velocity of flow at any instant in time. If the sing-around frequency is modulated with an oscillator frequency of C_0/l_1, then the difference frequency produced is V/l_1. If this is fed into a discriminator circuit an output amplitude proportional to the velocity of flow of the medium is obtained. Unfortunately the velocity of propagation of sound in the medium is not constant but varies with temperature according to the formula[7]:

$$C_0 = 4626 + 13{\cdot}8\,T - 0{\cdot}12\,T^2 \text{ ft/sec}$$

where T is the temperature in deg. C.

The output from the discriminator would therefore fluctuate for variations of either temperature or velocity of flow.

This disadvantage can be overcome by using two velocimeters, placed close together, side by side, arranged to measure velocity variations in opposite directions.

Their sing-around velocities will then be:

$$f_1 = 1/\tau_1 = \frac{C_0 + V}{l_1} \quad \text{(5)}$$

and

$$f_2 = \frac{C_0 - V}{l_2} \quad \text{(6)}$$

where l_2 is the length of the second path.

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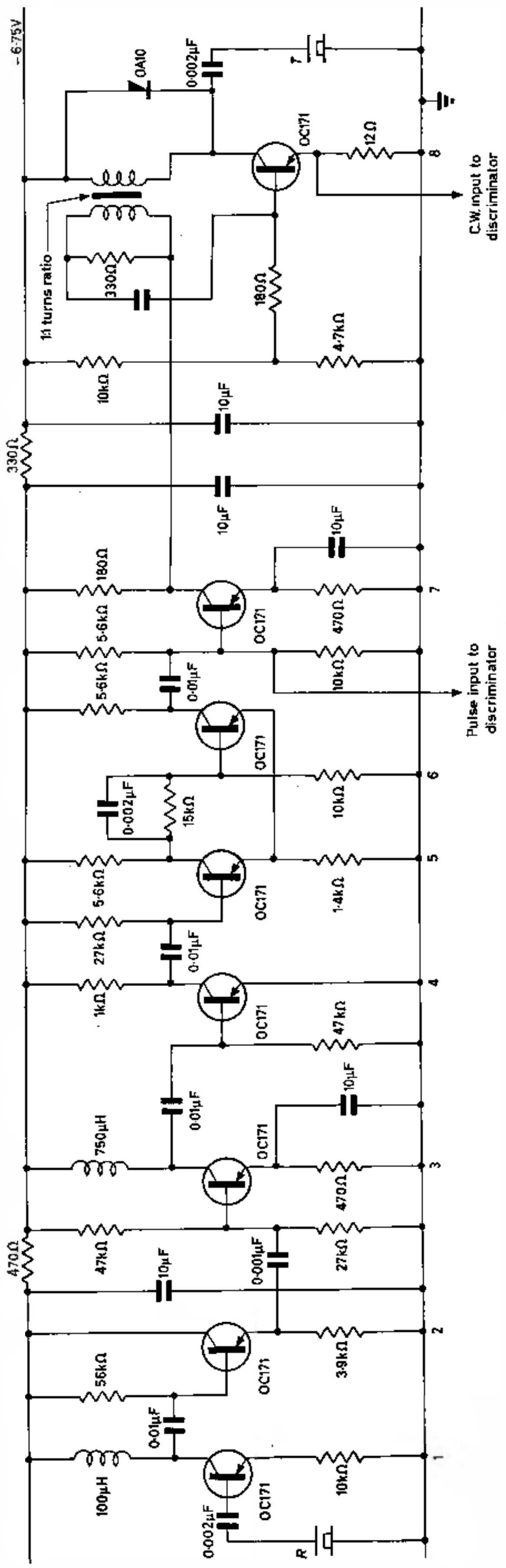


Fig. 2. The sing-around velocimeter circuit

If the sing-around frequencies of the two velocimeters are multiplied together in a modulator the output frequency, after passing through a low-pass filter, is:

$$\Delta f = f_1 - f_2 = \frac{C_0\,[l_2 - l_1]}{l_1 l_2} + V\,\frac{[l_2 + l_1]}{l_1 l_2} \quad \ldots\ldots\ldots\ldots\ldots\ldots \ (7)$$

Since $l_1 \simeq l_2 \simeq l$ this can be written as:

$$\Delta f = (\Delta l / l^2)\, C_0 + (2V/l) \quad \ldots\ldots\ldots\ldots \ (8)$$

where:

$$\Delta l = l_2 - l_1 \quad \ldots\ldots\ldots\ldots\ldots\ldots \ (9)$$

If $l_1 = l_2$

$$\Delta f = 2V/l \quad \ldots\ldots\ldots\ldots\ldots\ldots \ (10)$$

which is entirely independent of the velocity of propagation of sound in the medium.

In a turbulent medium V will be varying in a random manner because of the random motion of the turbulent eddies. If the velocimeter is used in such circumstances the resultant low frequency output Δf will vary in direct proportion to the velocity of any particular eddy, provided that the diameter of the eddy is considerably larger than the spacing of the velocimeter transducers l_1 and l_2.

In practical measurements the expression:

$$2V/l$$

can have either a positive or negative value depending on the motion of the eddy at any particular instant in time. If the term

$$\Delta l / l^2 \,.\, C_0$$

is made equal to zero then ambiguities will arise at the output of the modulator circuit, since it becomes impossible to determine which way the current is flowing merely by an inspection of the output difference frequency, which will always be positive. This ambiguity is easily overcome by making

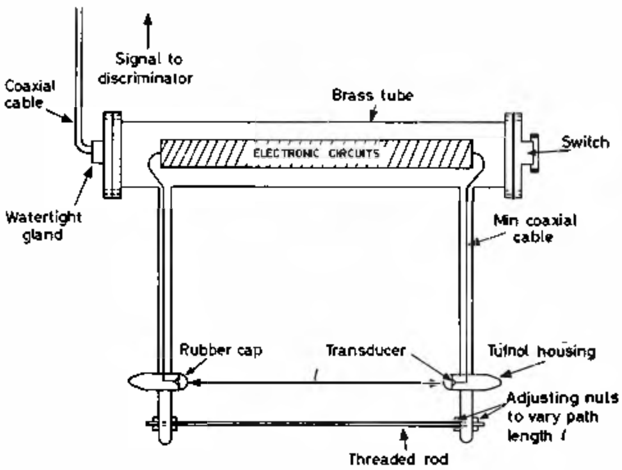
$$\Delta l / l^2 \,.\, C_0 > |2V/l|$$

The Practical Velocimeter

A circuit diagram of the sing-around velocimeter is given in Fig. 2. Each velocimeter uses eight OC171 transistors, the power required being 8mA at 6·5V. This can be obtained from a Mallory mercury cell, type TM 163.

The blocking oscillator is adjusted to run free at a frequency which is slightly lower than the sing-around frequency which will occur in practice. The oscillator produces a positive pulse of about 5V amplitude and 1μsec duration. This causes the transmitting transducer to execute

Fig. 3. Velocimeter construction



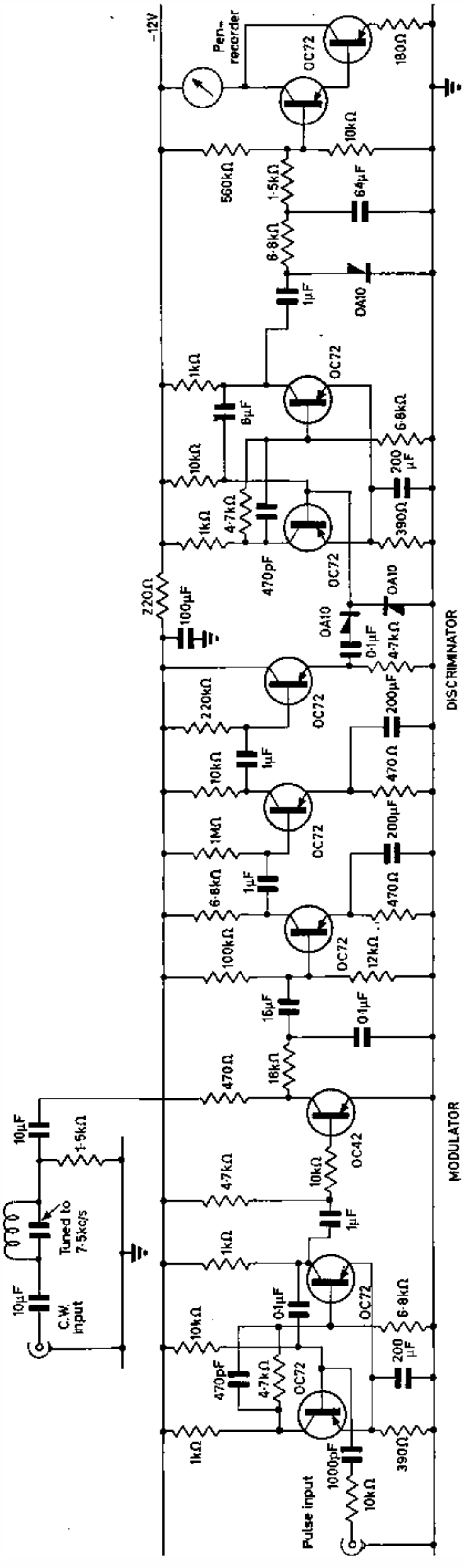


Fig. 4. Modulator and discriminator circuits

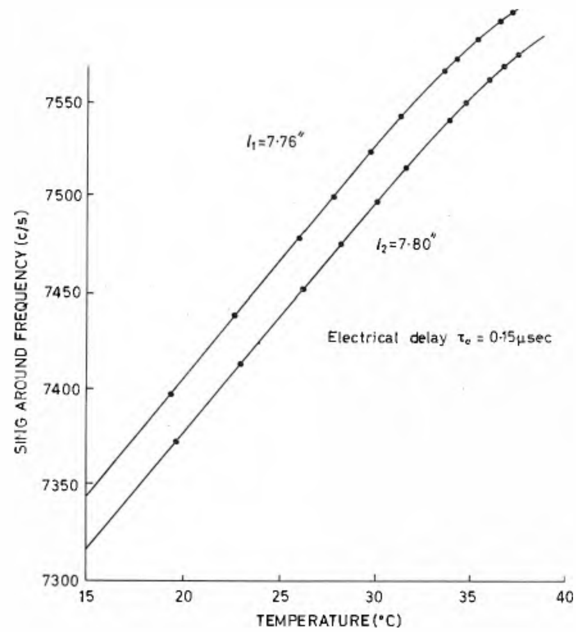


Fig. 5. Sing-around frequency against temperature

an exponentially damped vibration at its natural frequency of 3Mc/s and the corresponding pressure variation is transmitted through the water to the receiver. The receiving transducer is arranged to be polarized in the opposite direction to the transmitter so that the leading edge of the received signal, which is fed to an amplifier, is negative going.

The amplifier section comprises a common emitter stage followed by a common collector amplifier followed by another common emitter stage. This is connected to a detector which rectifies and amplifies the received signal, the output being about 3·5V positive going. The first four stages are used to sharpen the rise time of the received signal, stages 3 and 4 being heavily saturated. The output from the detector is fed to a Schmitt trigger circuit, whose corresponding output is a positive pulse of about 3V amplitude and 10μsec width. This is followed by an inverting, low output impedance power amplifier which produces the negative pulse used to synchronize the blocking oscillator. The output to the frequency measuring equipment is taken from the emitter of the blocking oscillator.

The transducers, lead zirconate titanate* PZT5 A disks, ¼in in diameter with a resonant frequency of 3Mc/s are positioned in two streamlined housings, approximately 8in apart, fixed to the watertight container holding the electronic equipment (Fig. 3). The output signals to be measured are taken through a coaxial cable to processing equipment in a boat overhead.

The output signal from the velocimeter operating at the higher frequency is turned into an approximately square pulse, by feeding it into a monostable multivibrator, which is then used to switch a simple modulator circuit. The output from the other velocimeter is passed through a tuned circuit, resonant at the sing-around frequency, and also fed into the modulator, the output of which is taken through a low-pass filter having a cut-off frequency of 50c/s and fed into a calibrated discriminator circuit (Fig. 4). The output from the discriminator is displayed on an Evershed and Vignoles recording milliameter of internal impedance 1·5kΩ.

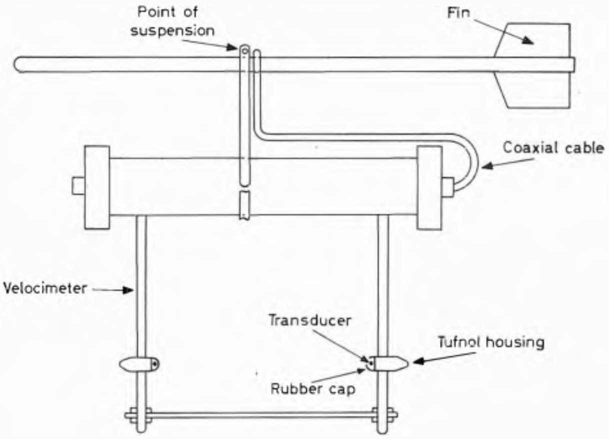


Fig. 6. Sing-around velocimeter showing stabilizing fin

Calibration and Stability of the Velocimeters

The calibration of the two velocimeters was carried out in a small tank of water. The two instruments were adjusted, by varying the acoustic path length l, so that their sing-around frequencies in water at room temperature (12°C) were 7 392 and 7 363c/s, thus having a difference of approximately 30c/s. They were allowed to sing-around freely for about an hour, while their p.r.f.s were measured with a pulse counter. It was found that both maintained their frequency within 0·2c/s over this period of time.

The water was then heated and a graph of sing-around frequency against temperature was plotted for each velocimeter (Fig. 5). By comparing these results with a theoretical curve of the velocity of sound in water against temperature, it was possible to determine the values of the acoustic path length and the time delay due to the electronic circuits. The lengths were $l_1 = 7{\cdot}76$in and $l_2 = 7{\cdot}80$in at 0°C and the electronic delays were both of the order of 0·15μsec.

The Use of Velocimeters for Measuring the Size of Turbulent Eddies

Experiments to determine the effectiveness of the velocimeter principle for measuring the sizes of turbulent eddies were conducted in the Menai Strait, near the Menai Bridge. The initial trials were made from a small dinghy, anchored in the Strait, but these were found to be unsatisfactory since any rocking of the boat tended to cause additional fluctuations in the recorded signals because of spurious vibrations induced in the pen recorder. These spurious fluctuations were eliminated in later trials by working from the 30ft vessel 'Lambda' belonging to the Marine Biology Station, Anglesey.

The velocimeters were arranged to point in the mean direction of current flow by connecting a large fin to the housing on the opposite side to the transducers as shown in Fig. 6. They were then lowered to a depth of about 5ft on the end of a thin cord. The outputs from the two instruments were taken through coaxial cables to the discriminator and pen recorder which were both in the boat. The system was allowed to run for several minutes and the resulting fluctuations were recorded. The velocimeters were then lowered further and the procedure repeated at 5ft intervals until a depth of 35ft was reached. Some of the recordings are shown in Fig. 7. The results were later analysed and the autocorrelation functions of the velocity fluctuations determined with the aid of a digital computer. These are shown plotted as a function of distance in Fig. 8. The point where the autocorrelation function reaches its minimum value corresponds to the radius of the mean eddy at any particular depth. From Fig. 8 it can be seen that the mean eddy size appears to vary with depth, reaching a maximum at the mid-depth of the medium at the point of measurement. (The Menai Strait is about 40ft deep at the point where these measurements were made).

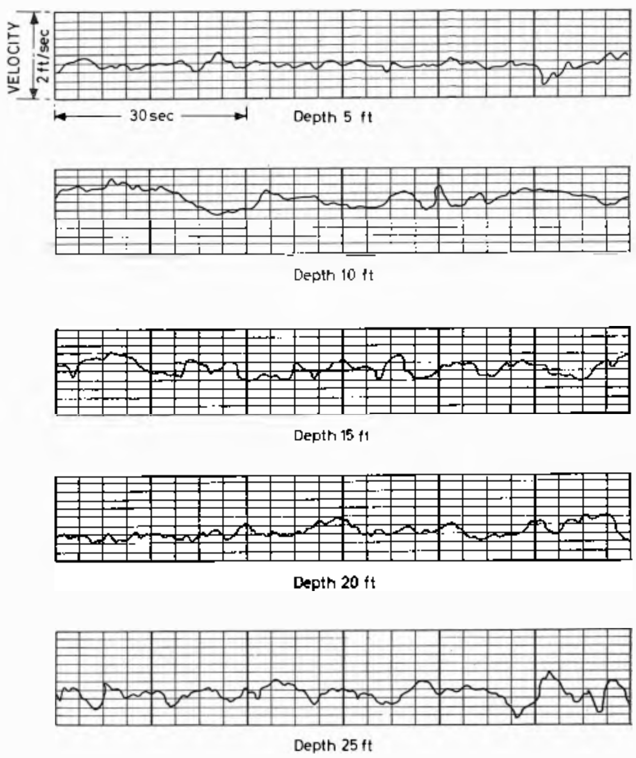


Fig. 7. Velocity fluctuations measured off the 'Lambda' at various depths in direction of current flow

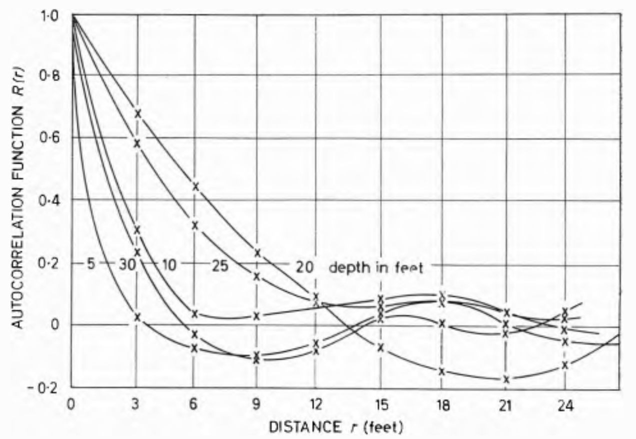
The velocimeters were also arranged to point in various other directions, relative to the mean direction of current flow, and more measurements taken.

A statistical analysis of the results showed that the turbulence at any particular depth appeared to be isotropic.

Conclusions

It has been shown that the sing-around velocimeter principle can be used to measure the variations in velocity of flow in a turbulent medium. When only one velocimeter is used, errors due to the variation of the velocity of

Fig. 8. Autocorrelation functions of turbulent eddies at various depths along line of current flow



sound with temperature affect the results. This failing can be minimized by using two separate velocimeters placed close together but measuring the flow velocity variations in opposite directions. Trials made with the instrument in the Menai Strait indicate that turbulent eddies vary in size with depth, but appear to be isotropic in character.

Acknowledgments

The author is indebted to Professor D. G. Tucker and Dr. B. K. Gazey for their continued assistance and interest throughout the work. He would also like to acknowledge the help given by Professor D. J. Crisp and Mr. E. I. Rees in placing the facilities of the Marine Science Laboratories, Anglesey, at his disposal on several occasions.

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Emley Moor Mast Nears Completion

Erection of the Independent Television Authority's new 1265ft mast—tallest in Europe—at Emley Moor, near Huddersfield is now nearing completion. It will be shared by the ITA and BBC and has been designed to carry the aerials of both organizations.

Main contractor for the supply of the new mast, aerials and feeder systems is EMI Electronics Ltd. Sub-contractor for the design, manufacture and erection of the mast is British Insulated Callenders Construction Co Ltd.

At Emley Moor a Band III transmitting aerial to be mounted on the mast will replace the ITA aerial on the existing 450ft tower.

The new aerial, of 80ft aperture, consists of eight tiers of full-wave dipole panels, enclosed in a 12ft diameter cylinder of glassfibre-reinforced plastic for protection against the weather and to facilitate all-the-year-round maintenance.

A directional horizontal radiation pattern was specified for this aerial to give a power of 200kW over an arc of 150° with a reduced power over the remaining sector. This radiation pattern has been achieved and demonstrated at the recent tests at the EMI aerial site at Hayes.

Two independent halves of the aerial—each of which continues to operate if a fault develops in the other—will each be fed by a 4½in diameter semi-flexible transmission line from the transmitters. Vertically polarized, the aerial will transmit on Channel 10.

The mast is one of three being built to a new design for the Independent Television Authority. The other two are at Winter Hill (Lancs.) and Belmont (Lincs.). The mast at Belmont will also be 1265ft tall but the one at Winter Hill has already reached its full height of 1015ft.

These masts set the pattern for a new generation of aerial supporting structures by departing from the open lattice type of stayed mast generally used in this country. The new design employs curved high tensile steel segments to form a 9ft diameter cylinder for the mast column which rises at Emley Moor to 900ft. A 350ft lattice section on top together with the capping cylinder brings the total height to 1265ft. The lattice portion, together with the upper part of the cylindrical steel section, is designed to carry five aerials capable of transmitting nine programmes.

This form of mast offers many advantages in addition to being able to support extra equipment without adding to the wind loading. Entry is at ground level through the 10ft high reinforced concrete superstructure on which the cylinder stands. Access to all parts of the structure from within is possible in the worst weather. All joints in the cylinder are weather proofed with sealing compound when they are bolted together and this gives complete protection to the aerial feeders and ancillary equipment within. The u.h.f. and v.h.f. aerial sections at the top will be enclosed in a glass-fibre shroud to maintain the cylindrical form and permit maintenance to be carried out in all weathers. The obstruction lights are hinged so that they can be swung inside for servicing. Maintenance men have the benefit of a lift in the cylindrical portion, travelling at about 100ft/min. Beyond that there is a hoist for carrying materials to the summit.

A feasibility study of cylindrical television masts was first started late in 1962 and, working in close collaboration with the I.T.A., BIC Construction Co. examined the special design and associated construction problems. A period of approximately one year was required to develop the final design for an aerodynamically stable structure of the height and strength required. The mast is designed to withstand wind loads equivalent to 120mile/h at the top. With this wind loading the mast head will deflect 6ft 3in from the vertical.

Hand in hand with the design work went investigations into the techniques of fabricating the curved steel sections which form the stressed shell of the mast. This was undertaken at the Hereford works of Painter Brothers Ltd. another member of the BICC Group, who designed special jigs and produced and assembled a trial section to prove design details and accuracy of fabrication before proceeding with the remainder of the mast.

New techniques had to be developed for building the mast. Erection of the first steel segments was carried out by an external mobile crane. When a height of 30ft was reached, a self-climbing cathead derrick was mounted centrally in the structure and attached to the interior of the cylinder by its two working platforms. When the mast had grown another 10ft, a small electric power winch on the upper working platform was used to raise the platform to its new position. With the upper platform secured, the lower working platform was detached from the cylinder and both it and the derrick pole, to the foot of which it is permanently attached, were lifted 10ft by the winch on the upper platform and secured in the new position. The cycle of operations completed, a further 10ft ring of segments could be erected.

The safe working load of the self-climbing derrick is 2 tons and the head may be slewed so that segments can be landed directly below it. The lifting bond runs from the ground-mounted diesel powered winch through a snatch block at the base of the mast, up the inside of the column and over the derrick head. Steelwork is controlled during lifting by guide wires or tail lines. When the cylindrical portion of the mast was completed the special derrick was removed and a conventional floating derrick used for the erection of the lattice part of the structure.

Safety and communications are vital matters on a constructional project of this size. A separate power winch operated a normal erector's chair for carrying men to the top of the structure during erection and they maintained constant contact with those on the ground by two way radio.

The steel stay ropes used to support the mast are all of locked coil construction giving greater strength than other types of rope. The largest stays are 2¼in diameter and weigh 15·29lb/ft, a 1140ft length thus weighing 7·8 tons. It is made up from 205 high tensile steel wires and is designed for a safe working load of 110 tons. The total weight of all the stay ropes is 85·4 tons and the total length of the wires that went into their make up is 520 miles.

The mast is stayed at six levels with three equally spaced stays at each level. The lower ends of the eighteen stays are secured by nine anchor blocks, the largest of which required 250 tons of concrete. Over 300 tons of concrete went into the mast base raft and superstructure which support a maximum thrust of just under 1000 tons from the mast column.

Work is already in hand for the next stage of the project, the fitting of the aerials and glass-fibre shrouds.

A Transistor Ratemeter with 2μsec Resolution

By B. Shepherd*

This article describes a random pulse ratemeter having seven rate ranges from 10p/s to 10kp/s and six averaging time-constants from 0·25sec to 50sec. Each input pulse generates a stable width square pulse which, in turn, forces a fixed quantity of charge into a storage capacitor. The resultant voltage is amplified in a chopper amplifier enclosed in a Miller loop to give the required integrating time. Meter output indication and 1mA and 100mV recorder outputs with an accuracy of 2 per cent are available.

(Voir page 492 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 499)

THE ratemeter described in this article was designed to be part of a transistorized scaling system for use in the nucleonic field. It was necessary that the unit be an independent plug-in module with controls on the front and power supplies, input and outputs taken from the rear.

When determining the rate of random pulses the accuracy of the final reading is determined by the resolution of the ratemeter. If an input pulse effectively paralyses the ratemeter for a certain time t and another input pulse occurs during time t the second pulse will not be included. The observed count rate N_o will thus be in error from the true count rate N_c. The error can be computed from the formula:

$$N_c = \frac{N_o}{1 - N_o t} \text{ counts/sec}$$

If necessary this modification factor can be included, and, knowing the paralysis time t the result can be corrected. If, however, it is desired that the accuracy only can be within 2 per cent then:

$$\frac{N_c - N_o}{N_o} \leqslant 0{\cdot}02 \text{ giving } t \leqslant \frac{0{\cdot}02}{N_o}$$

A maximum rate was initially chosen of 10kp/s giving a required maximum resolution time of $0{\cdot}02/10^4$ or 2μsec.

Each input pulse is amplified and the leading edge triggers a monostable pulse generator, Fig. 1. It is necessary that the generated pulse width remains constant for every input pulse regardless of the spacing of individual pulses. The pulse rise and fall times must be short. The width (w) is altered for each range of the ratemeter but must not exceed $0{\cdot}02/N_f$ where N_f is the full scale average rate. A value of

$$w = 0{\cdot}01/N_f \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (1)$$

i.e. 1μsec for the highest range, was selected.

The pulse output of the monostable is used to operate the constant charge pump. This diverts a particular value of current, decided by a stable reference, into a known capacitor, for a time equal to the monostable pulse width. The total charge input to the capacitor over 1sec will thus be directly proportional to the input pulse rate. Charge is removed from the capacitor by a parallel load resistor at a rate dependent upon the voltage across it. Equilibrium is therefore reached when the charge lost into the load resistor is equal to the total charge given by the pump.

$It = qNt$ $\quad$ I = load current
N = rate (p/s)
t = time
q = charge for one input pulse

$V/R = qN$ $\quad$ V = load voltage
R = load resistor

$$V = qNR \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots \quad (2)$$

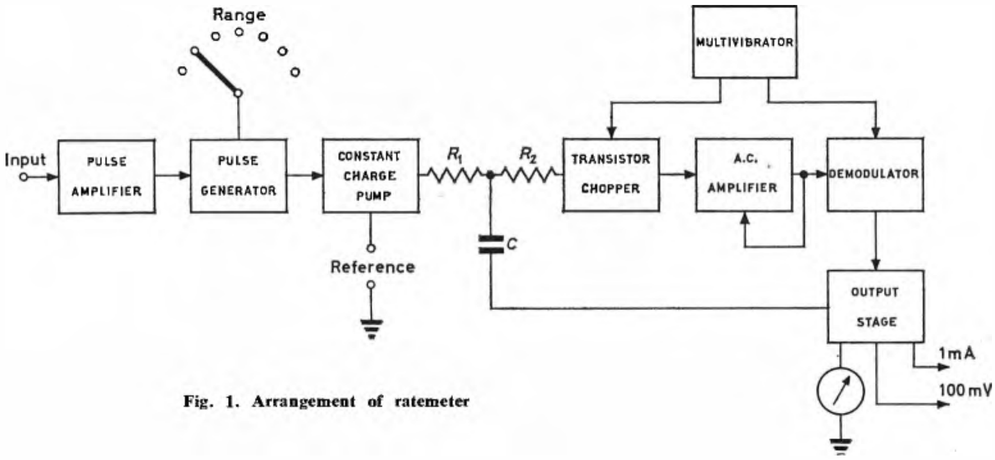


Fig. 1. Arrangement of ratemeter

Thus for a fixed quantity of charge and a known load resistor the output voltage is proportional to the input pulse rate.

The next problem, having derived a voltage proportional to rate, is to average this voltage out. The actual value of V can fluctuate wildly over small intervals of time particularly at low values of N, due to the random arrival of input pulses.

It is necessary to integrate the voltage preferably over a known time interval. The highest count rate may only need 0·25sec but the lowest, 10p/s, can require 50sec. This large time-constant could be obtained with a series resistor and a large electrolytic capacitor, say 100kΩ and 500μF. The relatively low leakage resistance of an electrolytic capacitor, particularly at high temperatures, precludes the use of this arrangement. The highest value non-electrolytic capacitor of a size suitable for transistor equipment is about 10μF. This involves a resistor of value 5MΩ. The only suitable transistor d.c. amplifier capable of this sort of input impedance, apart from costly field effect devices was a chopper amplifier.

A simple transistor chopper circuit is operated at 160c/s by a square wave derived from a multivibrator. It converts the d.c. obtained through the large resistors R_1 and R_2 in Fig. 1 to a small voltage square wave. This is amplified by an a.c. amplifier of adjustable gain and

* *Isotope Developments Ltd.*

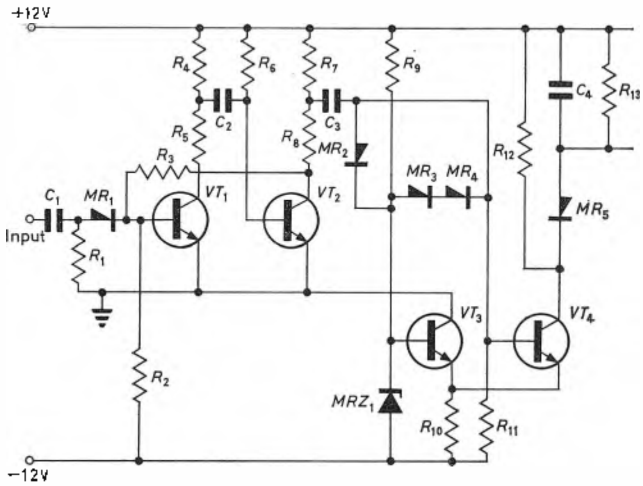


Fig. 2. Monostable and constant charge pump

then synchronously demodulated at the output of the amplifier. The resultant d.c. is filtered and buffered to feed several outputs. An increase in overall time-constant is obtained by connecting the integrating capacitor C to the output stage and not to earth.

Basic Ratemeter

The input pulse required to drive the ratemeter is negative with a sharp rise time of 100nsec or less. It is differentiated and inverted and the positive edge applied to the monostable of Fig. 2. As the output square pulse only switches the constant charge pump it is not necessary that its amplitude be very stable. The pulse width however is very important as errors will be directly proportional to any change.

The pulse width of a monostable in the configuration shown is dependent upon four things.

(a) Charging capacitor C_2

(b) Timing resistor R_6

(c) Supply voltage

(d) Cut-in point of transistor VT_2

The timing resistor and capacitor need to be stable components. R_6 was a 1 per cent metal film resistor and for C_2 1 per cent silver mica capacitors were used for the low values and 2 per cent paper capacitors for the high values.

The supply voltage affects the width in two ways. Any fall in the positive supply will decrease the edge fed through C_2 to VT_2 base and tend to reduce the pulse width. The same fall will reduce the voltage to which C_2 is charging via R_6 on the return and tend to increase the pulse width. These two effects tend to cancel.

The width will also be dependent to some extent upon V_{be} and β of transistor VT_2 as they affect the cut in point of VT_2 on the charging waveform. With the transistor used these effects were considered small.

The constant charge pump consists of two transistors VT_3 and VT_4 with common emitters. The base of VT_3 is held to a constant potential defined by the stable reference diode MRZ_1. The base of VT_4 is normally held about 1·2V below VT_3 base by the diodes MR_3 and MR_4 forward biased by resistor R_{11}. The tail current defined by R_{10} thus flows through VT_3 to earth.

The pulse derived from the monostable switches on VT_4. The base of VT_4 is held about 0·7V positive with respect to the base of VT_3 for the duration of the pulse by MR_2. The common emitters follow VT_4 base, cutting off VT_3. The tail current is therefore diverted to the collector of VT_4. A small portion flows into R_{12} forward biasing MR_5. The remainder of the current flows into C_4, the storage capacitor.

Temperature could affect this circuit in several ways and steps have been taken to overcome these effects. Firstly, the reference chosen is one of the composite devices having a stability better than the normal Zener. The nominal reference voltage is 8·2V and the maximum deviation at a fixed reference current is 15mV, over a wide temperature range. Changes in the V_{be} of VT_4 could affect the reference current but a degree of compensation is obtained by the diode forward voltage change of MR_2 as both VT_4 and MR_2 are silicon.

The leakage current of VT_4 in the off condition could be very important. The tail current chosen was 50mA. As the pulse width is $0{\cdot}01/N_t$ this represents an average maximum current of 500μA. A leakage of 5μA would thus cause an error of 1 per cent. Diode MR_5 reduces this effect. In the interval between pulses MR_5 is reverse biased with a maximum of 2·5V across it. It is a low leakage silicon diode keeping the leakage below 1μA even at 60°C. A leakage of 5μA through VT_4 will develop only 100mV across R_{12}, insufficient to affect MR_5.

C_4 is therefore charged with a pulse of current for every input pulse and discharges continuously into R_{13}.

From equation (1):

$$q = 0{\cdot}01/N_t \times i$$

where i = current pulse amplitude.

Substituting for q in equation (2):

$$V = \frac{0{\cdot}01iNR}{N_t}$$

At full scale on any range $N = N_t$

$$V = 0{\cdot}01iR$$

When i = 50mA and $R = R_{13}$ = 5kΩ

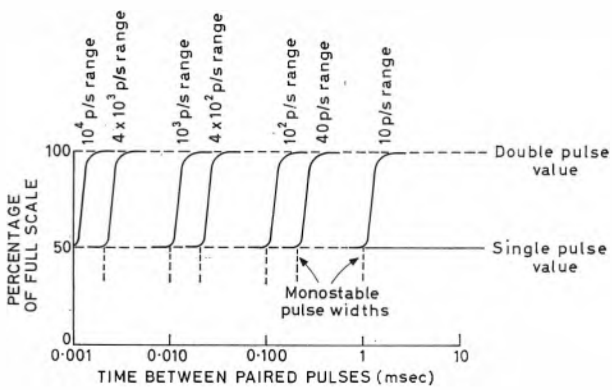
$$V = 2{\cdot}5\text{V}$$

It is, of course, possible to obtain a higher output voltage up to about 15V by increasing R_{13}. With random pulses, however, the voltage across R_{13} varies in a random manner, although the average is still 2·5V. Every input pulse will develop a voltage change across the storage capacitor C_4, of $v = wi/C$.

To reduce this change C_4 should be as large as possible. An inherent integrating time is developed by C_4 and R_{13} and it is preferable that this be kept as low as possible.

With the lowest rate range 10p/s the width of the pulse is 1msec. With a capacitor of 10μF the voltage step would

Fig. 3. Range resolution



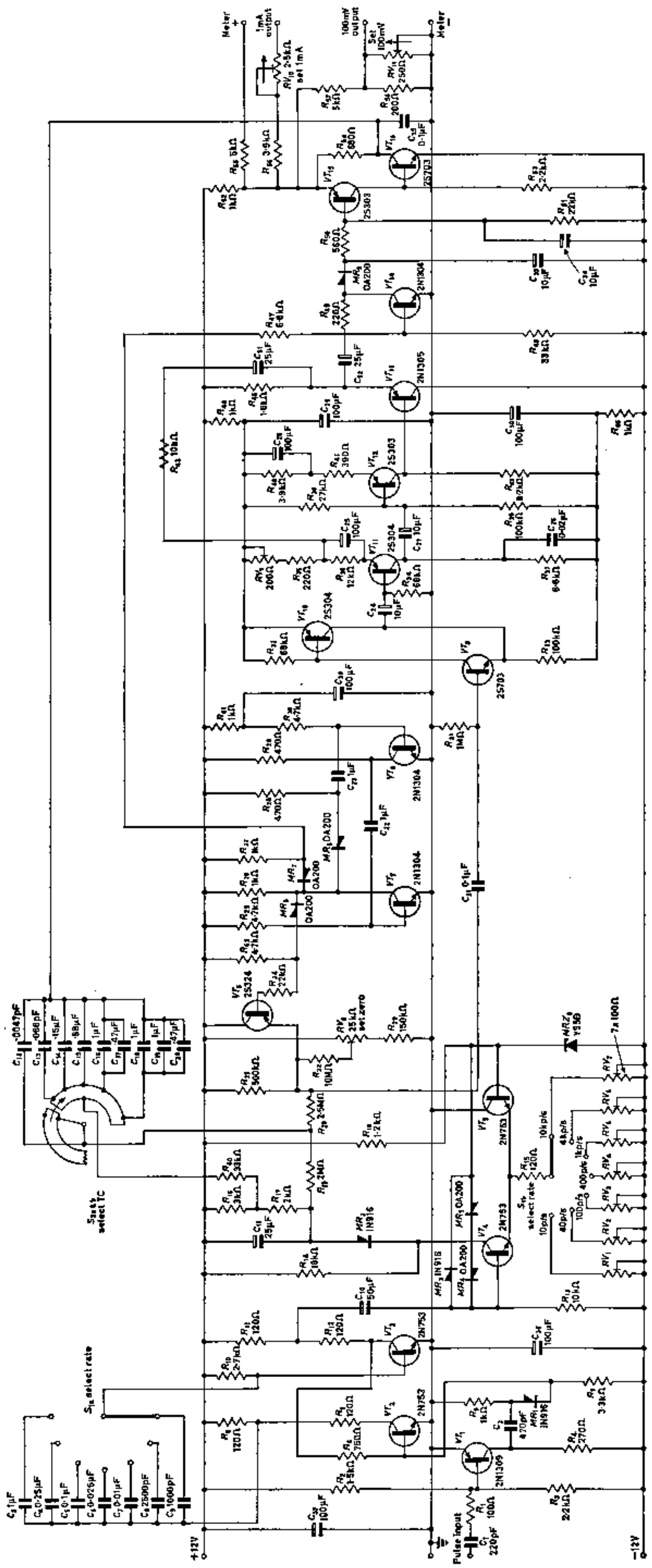


Fig. 4. Complete circuit of ratemeter

be 5V. Three or even more pulses could follow in quick succession, building up to bottom VT_4 in the on condition and making the circuit non-linear. A capacitor of 25μF, with a voltage step of 2V is statistically unlikely to cause bottoming.

With a storage capacitor of 25μF and a load of 5kΩ an inherent time-constant of 125msec is obtained. As the lowest time-constant required is 250msec this is allowable.

The resolution of the arrangement of Fig. 2 can be checked, using a double pulse generator where the overall number of pulses per second corresponds to the full scale rate but the pulses are in pairs and the time between paired pulses can be varied. As this time is reduced the error will increase and at some point the instrument will respond to only one pulse giving a reading of half full scale, Fig. 3. If the error at twice the monostable pulse width is less than 2 per cent the resolution is adequate.

Chopper Amplifier

It has been shown that the maximum average direct voltage out of the constant charge pump is 2·5V from a source resistor of 5kΩ. To obtain adequate averaging time-constants which would not introduce temperature effects paper capacitors were used. Values of above a few microfarads are physically too large, and the lower values required resistors of several megohms.

Although balanced transistor differential amplifiers with low drift levels are available they do not lend themselves to high source resistors as the leakage currents to the bases cannot be predicted or compensated for with any degree of reliability. A chopper circuit, however, with a low switching frequency can be used.

The problem involved in the choice of chopper circuit was not the usual one. Chopper circuits normally introduce drift due to the change in offset voltage with temperature. This drift, however, amounts to a few microvolts per degree Celsius, insufficient to cause any trouble in this application. The offset current, or leakage current, however, will have an effect. This is the current flowing through the chopper transistor in its 'off' condition. As it is difficult to do any compensation for leakage current the simplest possible chopper can be used. A special low leakage silicon chopper transistor, VT_6 in Fig. 4 was used, in the inverted condition. The emitter voltage in the 'off' condition is only 0·25V helping to keep the leakage very low.

Too high a chopping frequency will create substantial transients on the output waveform particularly with high impedance levels. These are due mainly to stored base charges. The only lower limit to the frequency is the overall time response of the amplifier and the difficulty of passing a low frequency square wave satisfactorily through an a.c. amplifier. A frequency of 160c/s was chosen.

The chopping waveform is obtained from the multivibrator of VT_7 and VT_8. The isolating diode MR_8 removes the inherent poor rise time of VT_7 collector giving a clean square wave for the chopper drive. As the base of VT_6 must be at exactly its collector potential during the 'off' condition, MR_6 prevents the leakage of VT_7 at high temperatures from biasing VT_6 base slightly 'on'. R_{61} and C_{38} initiate consistent starting of the multivibrator.

The square wave from VT_6 emitter representing the d.c. input is buffered by the composite emitter-follower of $VT_{9/10}$ and fed to the a.c. amplifier of VT_{11}, VT_{12}, VT_{13}.

The a.c. gain of this circuit is adjustable by means of RV_9 and stabilized against variations by the feedback via R_{43}.

The amplified square wave is synchronously demodulated by VT_{14} operated from the same square wave as that which operated VT_6. In the 'on' condition it clamps to earth d.c. restoring the square wave output. The diode MR_9 charges capacitor C_{33} to the square wave peak/peak value, with respect to earth. This type of system makes the gain relatively independent of the mark-to-space ratio of the square wave. Output filtering and a slight fixed d.c. shift are accomplished by R_{50}, R_{51} and C_{34}.

The set zero control RV_8 puts a small direct current into the input of VT_6 overcoming the fixed d.c. shift. This enables the output to be adjusted to zero for no input.

The Miller Increase

Three outputs were required.

(1) 0 to 1mA for a meter

(2) 0 to 1mA to operate a recorder

(3) 0 to 100mV to operate a recorder

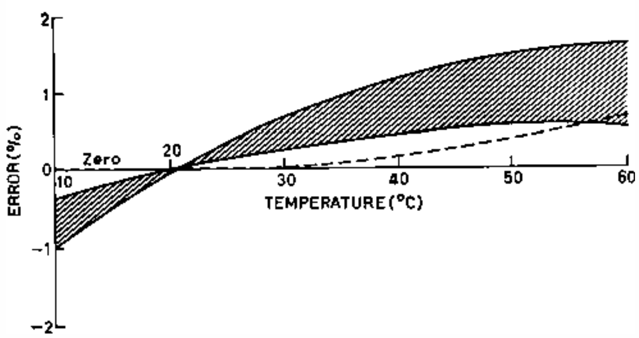
An output d.c. change of 5V was chosen, feeding the appropriate resistor networks. The compound emitter-follower of VT_{15} and VT_{16} buffers the filtered voltage to the output loads. The addition of R_{54} gives a point at VT_{16} collector, where an output swing of 11V can be obtained.

It can be seen in Fig. 1 that the timing capacitor C is returned to the amplifier output stage and not to earth. By using this technique the capacitance is effectively increased by a factor of $A + 1$ where A is the voltage gain between the changes at each end of the capacitor. This is similar to the increase of input capacitance of a valve due to the Miller effect.

The voltage change at the junction of R_{19}, R_{20} is 1·5V for a change at VT_{16} collector of 11V. The time-constant increase is therefore approximately 8·3.

The integrating time of the system depends upon the time response of the amplifier together with the time-constant decided by the resistors R_{19}, R_{20} and the chosen feedback capacitor. The time response of the amplifier is approximately 100msec, due largely to the output filtering.

Fig. 5. Temperature response of full ratemeter



Although a longer time-constant can be obtained by changing the ratio of R_{19} and R_{20}, keeping their sum constant, the resultant loading of the chopper circuit decreases the effective gain. No advantage is therefore gained.

With the values used an integrating time of 50sec was obtained using a 5μF capacitor. The leakage of this capacitor needs to be exceptionally low. An error of 1 per cent will be introduced for a resistance of 1 000MΩ.

In circuits of this type having long time-constants it is advisable, if possible, to arrange for the time-constant capacitors which are not being used to be pre-charged. If this is not done it is necessary to wait approximately five time-constants for each occasion that the time-constant switch is operated before a reading can be taken. In this case the load resistor of the constant charge pump is tapped to give the same ratio as R_{19} to R_{20} and R_{21}. Timing capacitors which are not in circuit are held to this potential via R_{60}.

Operation and Limits

The overall linearity of the system is good. On all ranges the linearity is within 0·2 per cent of full scale and will thus largely depend upon the quality of the meter used.

As with nearly all analogue devices the unit is dependent upon the stability of the supply lines. The error introduced by the positive supply is about ½ per cent for a 1 per cent change and that on the negative line ¼ per cent for a 1 per cent change. These effects tend to cancel when both supplies increase or decrease together giving a total change of ¼ per cent for 1 per cent change on both lines. With stabilized supplies this effect can be considered small.

The effect of temperature upon the monostable and constant charge pump has been mentioned. A further cause of drift due to temperature however is the high value resistors representing the integrating time, R_{19} and R_{20}. To obtain satisfactory results metal film resistors were used. The printed circuit boards on which the high value resistors, chopper circuit, and timing capacitors were mounted were found to be important. The leakage across an s.r.b.p. board at high temperatures was found to be excessive. An improvement of zero drift from 4 to 0·2 per cent of full scale over a 30°C temperature change was made by using fibre glass boards.

The temperature stability of the circuit is shown in Fig. 5. One curve indicates the range with the least drift and the other that with the greatest drift. The dotted line shows the zero drift which is common for all ranges. The change of drift with range indicates that the timing capacitors are responsible. The higher rate ranges with the silver mica capacitors gave the best results. The three lowest rate ranges with the large value paper capacitors had the greatest drift.

Temperature tests on the integrating times showed that they remained within 5 per cent of their 25°C value at 50°C.

Potentiometers RV_{10} and RV_{11} were connected to the recorder outputs to enable the recorders to be adjusted to their full scale.

Initial setting up is facilitated if the voltage across C_{31} is adjusted to 2·5V for a calibrated full scale input. The amplifier gain potentiometer RV_9 can then be adjusted to give full scale deflexion on the meter. After each range has been calibrated RV_9 can be adjusted if necessary to set full scale for all ranges and RV_8 to set zero for all ranges.

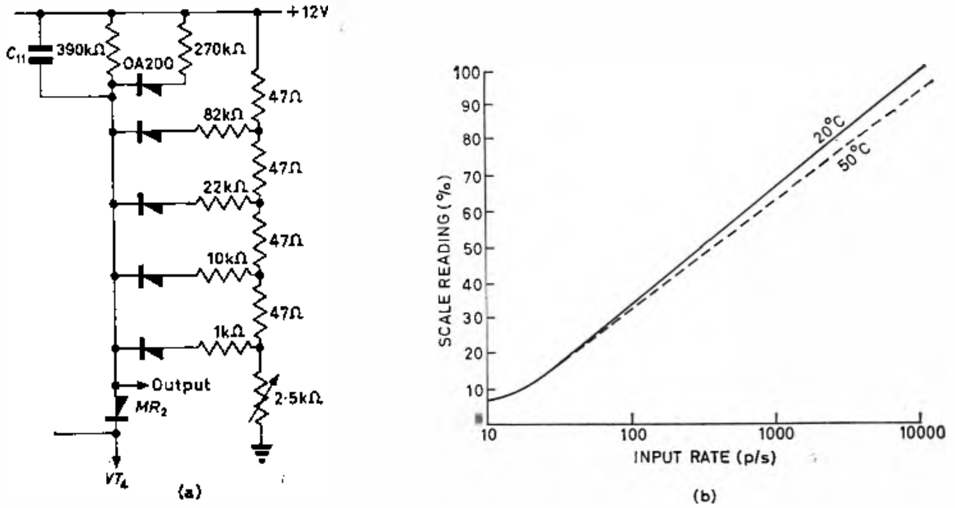


Fig. 6. (a) Circuit for logarithmic law (b) Law with above circuit

Modifications

Although this unit has been constructed to a specification there are several interesting extensions.

(a) *Time-constant increase*: The polyester capacitors chosen for the integrating times are by no means large, being a total of 5μF and this total could be increased if necessary to give a time of 100sec or more.

(b) *Time-constant reduction*: Although the full unit cannot be used for a time-constant lower than 0·25sec without major redesign it is possible for oscillographs to monitor across C_{11}. This necessitates reducing its value and disconnecting R_{60}. It can be useful for sweeping techniques with a high count rate, i.e. a time-constant of 0·02sec (with C_{11} about 4μF) showing the pulse shape derived from, for example, a radioisotope-labelled blood sample as it passes a detector.

(c) *Range extension*: To extend the ranges downwards to say 1p/s is feasible by increasing the monostable capacitor. It is necessary, however, to increase C_{11} and the integrating time also.

It is also possible to use the ratemeter with a maximum count rate of 100kp/s. The monostable width has to be reduced to 100nsec. This is possible with C_2 in Fig. 2 about 100pF. The input pulse edge needs to be sharp and C_1 and C_2, Fig. 4, need to be reduced.

The switching time of the constant charge pump is also adequate but the temperature dependence is slightly worse. A known paralysis time and a correction factor may be considered more reliable.

(d) *Logarithmic scale*: It is possible, by replacing the load resistor R_{18} in Fig. 2 by some other non-linear device to obtain an output voltage which is not proportional to the input rate but which follows some other known law. Where the rate is completely unknown or may vary over a large range while the ratemeter is left unattended feeding a recorder a logarithmic scale may be desirable. This may cover three decades: for example, a full scale deflexion of 10kp/s, but still giving a large deflexion, say 0·3 × full scale for 100p/s. A simple germanium diode chain could give this approximate logarithmic effect but it could not be stretched to three decades and it would be very temperature dependent.

The system indicated in Fig. 6(a) can be used to follow approximately the type of law desired. Adjustment of the cut-in points of the various diodes together with the parallel loads added can give an outline of the required curve. The greater the number of diode networks used the more faithfully the curve will be reproduced. A second advantage is that the integrating time will automatically be increased at the lower rates as the time-constant of C_{11} and the effective load resistor is increased.

Naturally a great consistent accuracy could not be expected of this arrangement but experimental results are shown in Fig. 6(b).

(e) *D.C. Amplifier:* It is possible that in the near future this amplifier could be partially replaced by a circuit comprising field effect transistors. At the present time the drift would still be much worse than that experienced with the chopper amplifier unless very expensive units are used. New developments and price reductions, however, may soon reverse this position.

Specification

Several modules have been built. The following is a typical specification.

Dimensions 9in × 3½in × 3¾in excluding meter

Rate Ranges 10; 40; 100; 400; 1 000; 4 000; 10 000p/s

Time-Constant
Ranges 0·25, 1, 2·5, 10, 25, 50sec

Paralysis time error $<$ 1·5 per cent

Linearity: Better than 0·2 per cent full scale

Zero drift $<$ 0·01 per cent/°C }
Signal drift $<$ 0·05 per cent/°C } 10 to 60°C

Supply stability 1 per cent for 4 per cent supply variations—Stabilized supply used

Chopper ripple at output $<$ 1 per cent of full scale peak-to-peak.

Conclusions

A random pulse ratemeter has been described combining two new features. The more usual diode pump circuit is replaced by a transistor constant charge pump having a far superior linearity at the small voltage levels involved in transistor circuits. Also the relatively large time-constants required for useful integration present a problem which has been overcome in a relatively drift-free manner by the use of a chopper amplifier having a high input impedance. The increase of the effective time-constant by the addition of a Miller capacitor illustrates how the chopper amplifier can be considered as any other d.c. amplifier providing the ripple is adequately removed at the output.

It has been shown that various improvements can be made to extend the ranges, in particular to 100kp/s and that logarithmic or other laws are obtainable.

Acknowledgments

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A Self-checking Cartesian A.C. Potentiometer for Use in the 100c/s to 10kc/s Range

By E. R. Wigan, B.Sc.(Eng), A.M.I.E.E.

A new form of Cartesian a.c. potentiometer, based upon a set of CR networks is described, and the theory, design and applications are discussed in terms of a model instrument.

By a self-checking routine the performance parameters of the model are calibrated in absolute terms. Subsequently the precision of the calibration is demonstrated by using the model to determine vector voltage ratios in some circuits of critical performance.

The frequency range runs in 100c/s steps from 100c/s to 10kc/s, and by an artifice, interpolations within this step are possible. A frequency-bridge forms a part of the Cartesian network and serves to calibrate the associated driving oscillator.

Although not precision built, the model is capable of 0·5 per cent measurement accuracy, which can be improved to 0·1 per cent by adopting a replication routine.

(Voir page 492 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 499)

THIS a.c. potentiometer measures, in Cartesian co-ordinates, the vector ratio between pairs of alternating p.d.'s without taking any current from the tested circuit; this has the advantage that, once the argument of the ratio is known as well as the modulus, methods of circuit analysis become possible which may be more rapid, and are always more revealing than conventional techniques based on voltage magnitude alone.

Measurements are made in terms of the readings of X and Y slide-wires that carry currents of substantially equal magnitude and quadrature phase. In the potentiometer circuit—of which Fig. 1 is a simplification—capacitance and resistance are used as the controlling elements; as a result the potentiometer is largely insensitive to stray magnetic fields.

The circuit can be treated as a passive 'bridge' and is operated in much the same way by bringing the detector D to a series of 'null balances'. Because the 'null' can be established so precisely ratio measurements can be made with a precision (circa 0·5 per cent) impossible with a pointer instrument. With the experimental model arranged as in Fig. 2 tests can be made at frequencies between 100c/s and 10kc/s, this range being covered either in 100c/s preset steps, or by a slight modification of test-procedure, in much smaller increments.

The following description is based on the network of Fig. 1 which is electrically equivalent to Fig. 2.

The two slide-wires SW_1 and SW_2 have linearly divided scales with centre zeros and equal positive and negative markings (150 divisions). As they are effectively connected in series by the transformer T_3, the slide-wire p.d.'s are additive. Voltage vectors in the tested network which lie anywhere in the four quadrants of the complex plane can therefore be dealt with.

In making a measurement the 'potential leads' P_1 and P_2 are bridged across the unknown p.d. (e_1) appearing in the test-object Z; the X and Y dials are operated to bring D to zero (thus ensuring that no current flows out of Z) and, assuming that the a.c. potentiometer needs no calibration corrections, the unknown p.d. is registered as $(X_1 + jY_1)$.

When a second p.d. (e_2) has been registered similarly as $(X_2 + jY_2)$ the vector ratio e_2/e_1 is known to be

$$e_2/e_1 = (X_2 + jY_2)/(X_1 + jY_1)$$

Although tiresome to handle in this form, the ratio can be reduced to a single term in X and Y by arranging that

$$X_1 = 100 \text{ and } Y_1 = 0$$

For this the phase-shifter in Fig. 1 is adjusted so that the p.d. (e_1) in the denominator of the ratio is adjusted to align precisely with the potentiometer dial setting (100 + j0·0). This being done the ratio e_2/e_1 is read direct from the dials as

$$e_2/e_1 = (1/100).(X_2 + jY_2) \quad \ldots\ldots\ldots \quad (2)$$

Equation (1) is true only if the slide-wire currents I_2 and I_1 have equal magnitude and quadrature phase. The actual ratio will be expressed by $I_2/I_1 = (\alpha + j\beta)$ in which the factor α should ideally be zero, and β unity. In practice α will be finite and β will not be unity, and to be realistic equation (1) must be re-written

$$e_2/e_1 = (1/100).(X_2 + \alpha Y_2 + j\beta Y_2) \quad \ldots\ldots\ldots \quad (2)$$

The exactitude of the ratio measurement therefore depends not only on the relatively simple matter of scaling the X and Y dials (purely linearly), but also on measure-

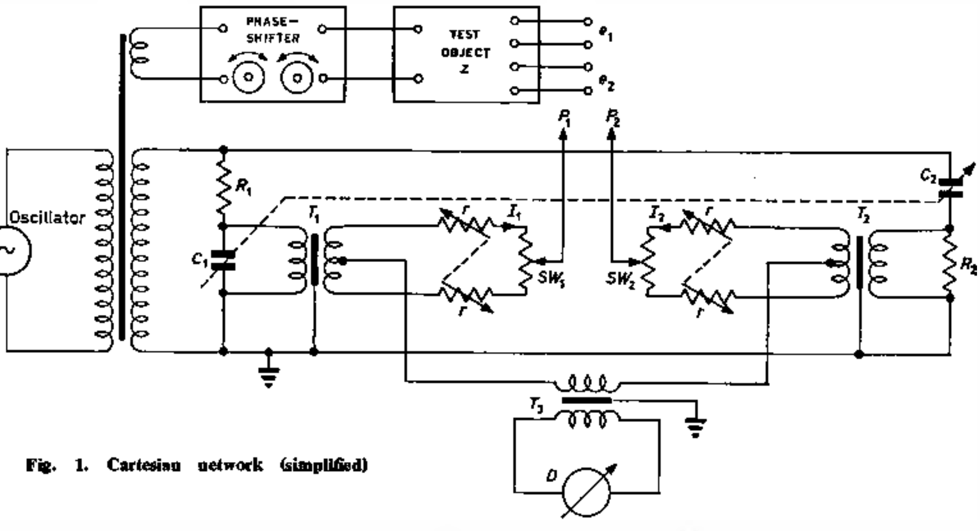


Fig. 1. Cartesian network (simplified)

ment of α and β. These parameters are measured by the self-checking procedure dealt with later when the model instrument is calibrated. It is shown that in the model α has a maximum value of $0{\cdot}27 \times 10^{-2}$ and β varies irregularly from 0·955 to 0·995 over the frequency range 100c/s to 10kc/s.

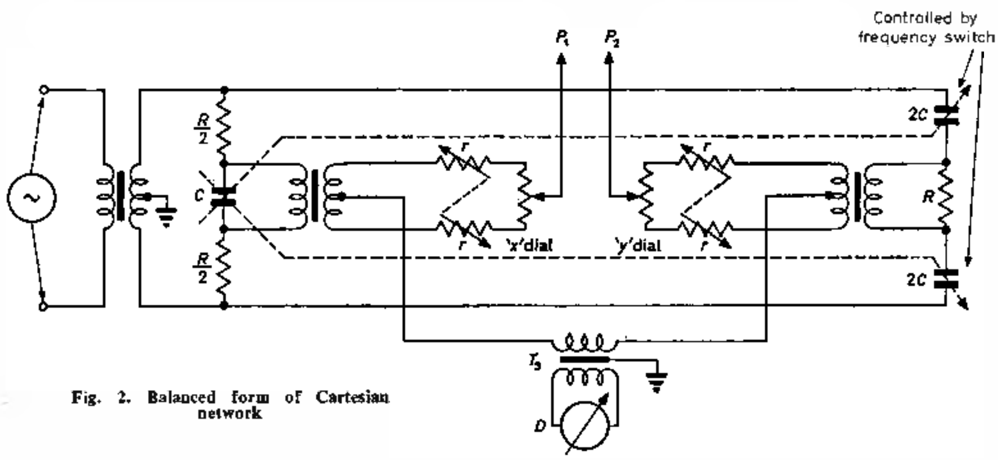
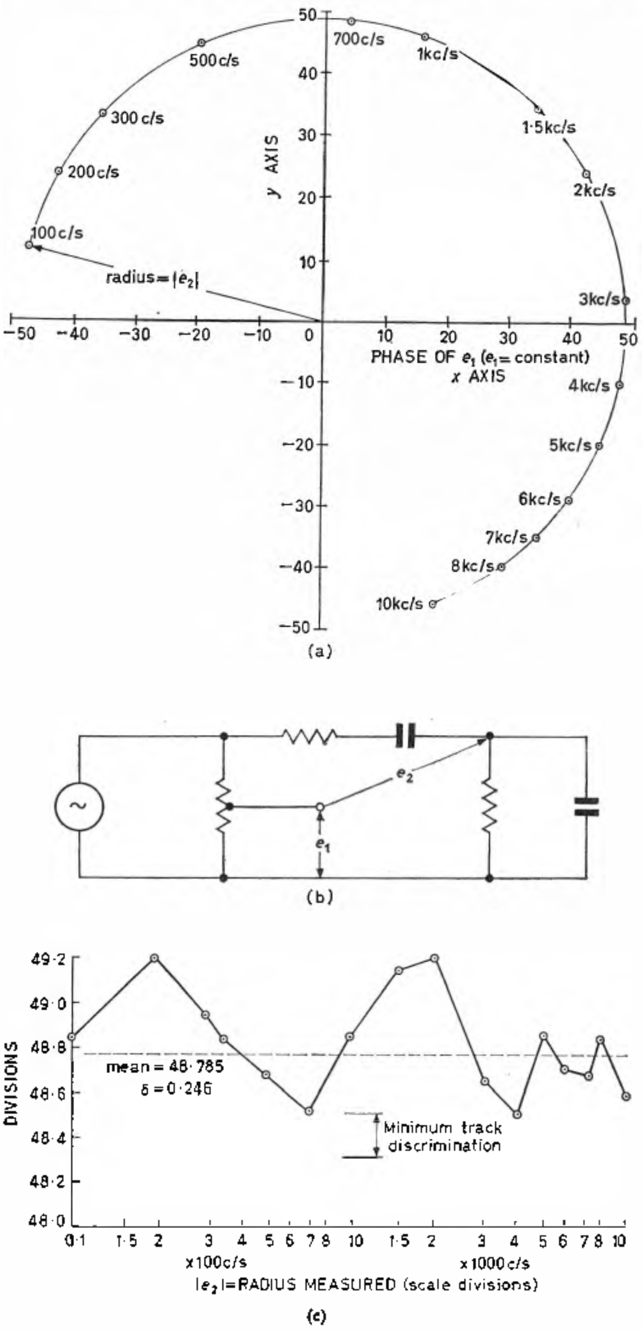


Fig. 2. Balanced form of Cartesian network

The model was not built to precise limits because, no matter how big the calibration corrections of the instrument might be, the self-checking procedure would always lead to precise measurement. A simple precision test is illustrated by Fig. 3(a) where the plotted points fall very closely indeed on the theoretical circle. To obtain this precision, however, both α and β had to be allowed for, and it will be appreciated that tedious computation could be avoided if α were made very small and β very nearly unity; this could have been approached with precision-grade components.

In the next section the various factors controlling α and β are dealt with, α getting smaller as the Q-factor of the controlling capacitors improves, and β depending on the capacitance values selected; in the model cheap industrial grade capacitors were used and adjusted to 1 per cent. The controlling feature of the design, however, is the circuit arrangement now to be dealt with.

Fig. 3. (below) (a) Simple precision test (b) RC circuit used (c) Plot of deviation



Theory of the Cartesian Network

Even with ideal circuit components the desired equality and quadrature between slide-wire currents depends on the following circuit theorem[1].

When two electrical elements, of impedance A and B, are connected in series across a generator of e.m.f. E, and internal impedance zero, a third element Q, connected in turn across A and B, will be found to be carrying currents I_A and I_B, respectively, which bear the ratio

$$I_A/I_B = A/B$$

this ratio being completely unaffected by the nature of Q.

The skeleton diagram, Fig. 4, applies this theorem to the a.c. potentiometer. Notice that the A, B and Q elements of the theorem have been duplicated so that the 'switching operation' suggested in the theorem has been avoided. Then, provided the duplication of the elements is perfect,

$$A_1 = A_2;\ B_1 = B_2;\ Q_1 = Q_2$$

leading to

$$B/A = I_{p1}/I_{p2}, \quad \text{(the ratio of transformer input currents).}$$

The first two equalities are met by choosing $R_1 = R_2$ and $C_1 = C_2$; it is also relatively easy to make $Q_1 = Q_2$ at all frequencies provided the slide-wire resistances, SW_1 and SW_2, are equal and the transformers are of identical construction. Notice that there is no need for transformers of perfect performance, but that it is necessary that at any working frequency their imperfections shall be alike; it will then follow that their current-transfer ratios will be approximately the same also. Finally if the transformers are similar the slide-wire current will have very nearly the same ratio as the currents I_{p1} and I_{p2} so that

$$I_1/I_2 = B/A \quad \ldots\ldots\ldots\ldots\ldots\ldots \text{(3)}$$

In fact the transformers will never be exactly alike; for example, in the case of the unity-ratio Post Office line transformers used in the model, each loaded by 3 200Ω slide-wires, Q_1 and Q_2 were (2660 + j45) and (2600 + j44·0) at 100c/s and (3110 + j0) and (3060 + j4·0) at 10kc/s; there were also minor differences in current transfer ratio.

To this extent therefore the conditions stipulated by the circuit theorem are not met. But even if the transformers were perfect and the slide-wire currents strictly in the ratio B/A, imperfections in the capacitors would introduce the quadrature correction α; moreover the factor β would be unity only if the capacitors C_1 and C_2 had precisely the capacitance suited to the frequency of measurement, i.e. $C = 1/2\pi fR$. The decade switches of Fig. 5 control the capacitance, and the frequency is established by balancing the Wien bridge described in the next section.

The Frequency Bridge

By turning a 'function' switch from MEAS to FREQ the R and C elements of Fig. 1 are converted into the

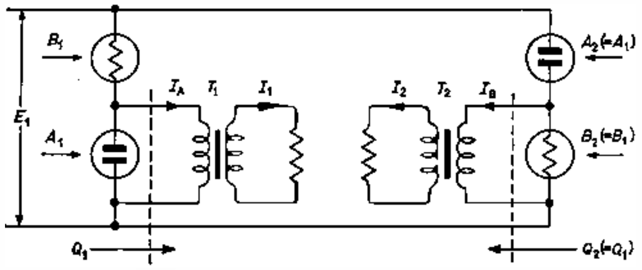


Fig. 4. Skeleton net (illustrating circuit theories)

Wien bridge of Fig. 6. In the model instrument $R_1 = R_2 =$ 1592Ω. The bridge is balanced by varying the frequency and/or the capacitance. It must then follow that the reactance of each capacitor is equal to each resistance; that is to say

$$1/C_1 = 2\pi f \,.\, R_1 \text{ and } 1/C_2 = 2\pi f \,.\, R_2 \quad \ldots\ldots \quad (4)$$

The function switch is returned to MEAS, once the bridge has been balanced, which restores the R and C elements to the Cartesian network; there is now some assurance that $|I_1|$ will very nearly equal $|I_2|$, that is $\beta \simeq 1{\cdot}0$.

The only unusual feature of the bridge is the switching scheme of Fig. 5 which operates on the 'decade-additive' principle which is explained as follows by taking advantage of equation (4):

If for frequency f' some capacitance C' is needed; where $f' = K/C'$; and if for frequency f'' some capacitance C'' is needed; where $f'' = K/C''$; then it must follow that for a frequency:

$$(f' + f'') = K(1/C' + 1/C''),$$

the capacitance needed must be C' and C'' in series.

For example, in the model potentiometer 0·1μF is needed for 1kc/s, and 0·05μF for 2kc/s. For 3kc/s, therefore, 0·1μF and 0·05μF are connected in series; similarly a frequency-increment of 100c/s is obtained by adding, in series, 1·0μF. By adopting this principle the selector switches of Fig. 5 have been arranged to indicate the balancing frequency by the sum of their markings. Here it must be pointed out that the model circuit of Fig. 2, although the electrical equivalent of Fig. 1, is in fact balanced to earth about a central line to reduce stray capacitance currents. In consequence Fig. 2 calls for two extra switches wired as in Fig. 5 but with double-value capacitors; the fixed resistors are also redisposed.

Because the capacitors in the model have been selected to no finer limits than ±1 per cent, the balancing frequency does not precisely agree with the switch-markings; that this is of little importance will be shown in the next section.

Self-Calibration Procedure[3]

The procedure is analogous to the method of getting a true weighing from an inaccurate chemical balance by exchanging the known and unknown weights on the pans. To calibrate the potentiometer at a chosen frequency a fixed voltage ratio, called here the 'reference ratio', is measured in eight different ways in terms of certain X

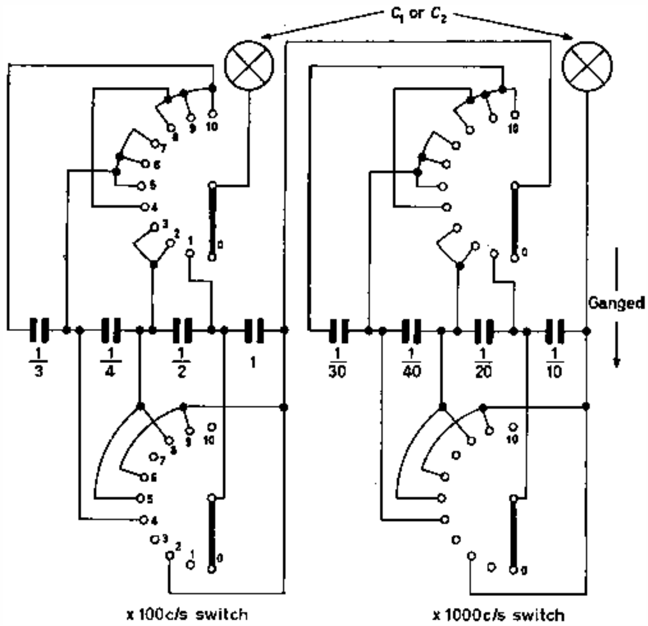


Fig. 5. Decade capacitor switching
(Used with $R_1 = R_2 = 1592\Omega$ in Fig. 1)

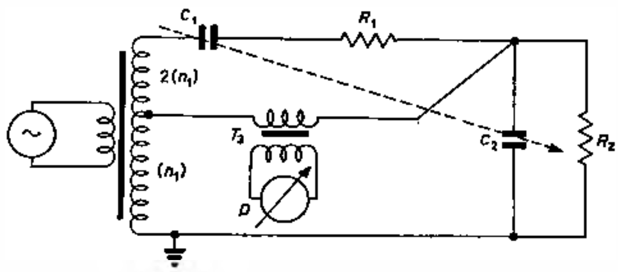


Fig. 6. Frequency bridge

and Y dial readings. Readings have to be assumed subject to unknown corrections, but when they are all 'averaged' the true measure of the reference ratio is derived and, by inference, all the potentiometer corrections are derived as well. In this way the quantities α and β are established together with the scale zeros on each dial without having to make reference to any standard quantity. This test-routine will be called the 'eight-point test'. (Attention is drawn to Appendix B which deals with a simpler three-point test applicable when the scale zeros are known to be correct).

The reference ratio is provided by a series combination of R and C connected at the output of the phase-shifter; R and C are chosen so that the ratio e_C/e_R approximates 1·0, 90° at the frequency at which the test is made. The approximation can be very crude and a modulus between 0·8 and 1·0 and an argument of 80° to 90° are acceptable.

Fig. 7(a) illustrates the first two (x_1y_1 and x_2y_2) of the eight successive measurements of the ratio e_C/e_R; all eight are set out in Fig. 7(b) and in Table 1. The phase-shifter is used to rotate the reference voltages so that they lie in each of the quadrants in turn. At the outset of each ratio measurement one or other of the voltages e_C or e_R is aligned with ± 100 divisions-mark on the X or Y axis of the potentiometer as shown in column 2 of Table 1; the X and Y co-ordinates of the non-aligned vector constitute the test-data, x_1, y_1; x_2, y_2; etc. The measurements in column 4 will be used to demonstrate the analysis.

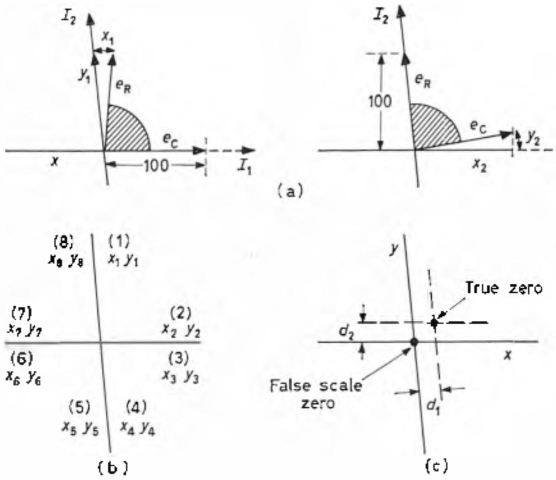


Fig. 7. (a) Calibration example (b) Test location (c) Location of electrical centres

TABLE 1

(1) TEST SERIAL NUMBER	(2) ALIGNMENT OF VOLTAGE e_C		(3) CO-ORDINATES OF VOLTAGE e_R		(4) DEMONSTRATION DATA AS COL. 3 (SIGNS AS MEASURED)	
	X	Y	X	Y	X	Y
1	+100	0	X_1	Y_1	+2·6	+107·6
3*	0	−100	X_3	Y_3	+97·0	+1·6
5*	−100	0	X_5	Y_5	−3·15	−106·0
7	0	+100	X_7	Y_7	−96·6	−1·75
	ALIGNMENT OF VOLTAGE e_R		CO-ORDINATES OF VOLTAGE e_C			
4	+100	0	X_4	Y_4	−2·0	−103·3
6*	0	−100	X_6	Y_6	−94·6	−2·35
8*	−100	0	X_8	Y_8	+1·3	−103·1
2	0	+100	X_2	Y_2	+93·05	+2·75

In Col. 4 the smaller of each pair of figures are designated as P and the larger Q. Thus $P_1 = +2{\cdot}6$; $Q_2 = +93{\cdot}05$.

First the P-data appearing in tests 3, 5, 6 and 8, and marked by an asterisk, have to be reversed in sign; thus P_3 becomes $-1{\cdot}6$ and P_5 becomes $+3{\cdot}15$ and so on. Then it can be shown that

From data of Col. 4

$$\alpha = (1/800)\,.\,[(P_3 + P_4 + P_7 + P_8) - (P_1 + P_2 + P_5 + P_6)]$$

$$\alpha = -2{\cdot}19 \times 10^{-2}$$

and the cotangent (γ) of the argument of the reference ratio is

$$\gamma = (1/800)\,.\,(P_1 + P_2 + \ldots + P_7 + P_8) \qquad \gamma = 0{\cdot}525 \times 10^{-2}$$

The location of the electrical centre of each slide-wire, defined by the (positive) quantities d_1 and d_2 in Fig. 7(c), is deduced:

$$d_1 = (1/8)\,.\,[(P_1 + P_2 + P_3 + P_4) - (P_5 + P_6 + P_7 + P_8)]$$

$$d_1 = +0{\cdot}04 \text{ divs}$$

$$d_2 = (1/8)\,.\,[(P_1 + P_2 + P_7 + P_8) - (P_3 + P_4 + P_5 + P_6)]$$

$$d_2 = +0{\cdot}05 \text{ divs}$$

(To make use of the corrections d_1 and d_2 in practice, they have to be reversed in sign and added to the test-data under observation.)

The values of d_1 and d_2 just deduced are very small because the zero-adjusters r in Fig. 1 were set in advance*. Because of the finite number of turns of resistance wire in the slide-wires, the slide-wire voltage changes in steps of 0·2 divisions, but as the average of eight individual measurements is being dealt with, differences as small as 0·05 divisions are significant; on the other hand the 0·05 intervals shown in the data recorded in column 4 have, of course, been derived by visual estimation between these steps, and are subject to different forms of error, some mechanical and others visual.

Some useful checks on the P-data can be made before dealing with the Q-data. If all measurements were made with absolute precision, P_1 would equal P_2 (see the two vector diagrams in Fig. 7(a)); moreover the same sort of equality must exist in each of the other quadrants of Fig. 7(b) provided the signs of P are adjusted as shown in column 1 of the Table. It follows that as a first check the adjusted P-data should fall into sequential pairs of the same sign; as a second check, if the measurements have been made carefully the magnitude discrepancy should not exceed $2 \times 0{\cdot}2$ divs. The quantity 0·2 appears twice because the measurement is concerned with a ratio, and the alignment setting (column 2 of Table 1) is just as likely to be in error as the P-data.

Although P_5 and P_6 in the Table differ by 0·80 divs, all other pairs are in order. A re-test always corrects the faulty data but in this case this single error will have negligible influence on the average and has been ignored.

Analysis of the Q-data is simplified if the quantity $|Q| - 100$ is dealt with: then only small quantities are involved. Writing this difference as Q' there are eight values, Q_1', Q_2', etc. up to Q_8', and these again fall into sequential pairs of equal magnitude (but here they have opposite sign). Tolerance $2 \times 0{\cdot}2$ divs again applies and it is seen that this is exceeded in the case of $Q_1' = +7{\cdot}6$ and $Q_2' = -6{\cdot}95$. The difference between other pairs being less than 0·4, one may ignore the single discrepancy and deduce

$$F = \frac{\text{(Voltage drop per scale-division on the } Y\text{-dial)}}{\text{(Voltage drop per scale-division on the } X\text{-dial)}}$$

$$= \text{'Scale Factor'} = \sqrt{[\alpha^2 + \beta^2]}$$

Deduced from Col. 4 Table 1

$$= 1 + (1/800)\,[(Q_2' + Q_3' + Q_6' + Q_7') - (Q_1' + Q_4' + Q_5' + Q_8')]$$

$$F = 0{\cdot}9526$$

(note that since α is very much smaller than β, $F \simeq \beta$ very closely)

and $e_C/e_R =$

$$1 + (1/800)\,.\,[(Q_1' + Q_3' + Q_5' + Q_7') - (Q_2' + Q_4' + Q_6' + Q_8')]$$

$$e_C/e_R = 1{\cdot}017$$

The 'reference' voltage ratio e_C/e_R used in this demonstration was obtained from a good class resistor and a mica capacitor (which accounts for the very small value of γ) and the frequency used was 800c/s. (The reader may notice a discrepancy between the α and β given above and those plotted in Fig. 8. The explanation is that, owing to

* *Both dials set to '0'; P_1 and P_2 clipped together; r-r adjusted for null balance at D.*

an error in the capacitance used in the frequency-bridge when set to 800c/s, the frequency used for the tests of Fig. 8 was in fact 770c/s which resolves the discrepancy: the reactance of the capacitor C_1 would be higher in that case, and the slide-wire p.d. also higher, in the ratio 800/770; hence β_{800} will exceed β_{770} in the reverse proportion.)

Usage

General

Some applications of the instrument will now be dealt with.

It is true that the 'phase-sensitive voltmeter' covers some of the same ground, but the null-balance technique of the a.c. potentiometer gives an inherent reading accuracy much higher than the pointer instrument. Again, as a ratio-measuring device, the potentiometer is self-calibrating and (see below) it can be voltage-calibrated if required.

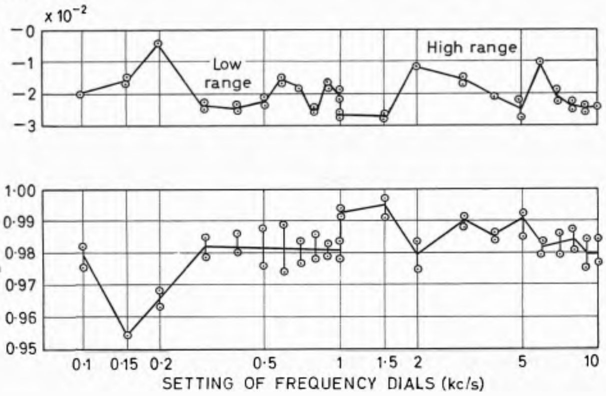


Fig. 8. α **and** β **by three-point test**
(average of $\alpha_1\,\alpha_2$ *and* $\beta_1\,\beta_2$*)*

(a) *Voltage Calibration*

It is a simple matter to include, somewhere within the tested network, a calibrated voltmeter; then in the course of the potentiometer ratio measurements the potential leads are connected to the meter terminals, thus yielding a calibration of the dials in voltage. This confers on the instrument the properties of a phase-sensitive voltmeter of effectively infinite input impedance, high reading accuracy and effectively isolated from ground.

(b) *Side-Wire Voltage Available* (E_s)

(Note: the total voltage (E_s) on the slide-wires consists of $+E_s/2$ and $-E_s/2$ with a central zero.)

Let E_1 be the p.d. delivered from the oscillator to the input terminals of the potentiometer network. Then, assuming unity-ratio slide-wire transformers (as in the model net) and no transformer losses:

$$|E_1/E_s|^2 = 1 + (1 + 1/q)^2;$$

($q = Q/R$, the ratio of the slide-wire transformer input impedance, when loaded, to the reference resistance R.)

In the model $q \simeq 2{\cdot}0$ whence, ideally,

$$|E_s| = (0{\cdot}555)\,E_1$$

Of this half, $(0{\cdot}277)\,E_1$, is available as the maximum positive or negative slide-wire voltage. In practice the transformer power-losses reduce this figure to a minimum of about $(0{\cdot}200)\,E_1$, near 200c/s, or $(0{\cdot}22)\,E_1$ at higher frequencies.

Unless there is a specific demand for higher voltages, the one-to-one slide-wire transformer ratio has advantages: higher ratios, though giving higher voltages, increase the capacitance to ground (or shield) and may affect measurement accuracy (see Conclusion).

When, in spite of this, the transformer ratio is increased the ratio should 'match' the slide-wire resistance to $R/\sqrt{2}$, not to the reference resistance R; this is because the reference capacitor C influences the impedance of the effective 'source' from which the slide-wire transformer is supplied.

(c) *Artifice for Making Tests at Other than 100c/s Intervals*

(This technique was used to obtain the data in Fig. 8 at 150c/s.)

Tests at frequencies not marked on the frequency-selector switches can be made, provided a calibrated oscillator is available. The selector switches are set to the nearest (lower or higher) frequency-marking (f_m) and the oscillator to the desired frequency (f_o) at which tests are to be made. Suppose f_o is 150c/s and the switches have been set to '100c/s' (f_m).

As the applied frequency (f_o) is higher than the selected frequency (f_m) the reactance of the selected capacitors will be too low in the ratio 2/3 and the voltage available to generate the current in the Y slide-wire will be similarly reduced relative to the X. That is

$$I_Y/I_X = f_m/f_o = 2/3, \text{ in this case.}$$

Provided this discrepancy is allowed for, by multiplying all readings taken from the Y-dial by f_m/f_o (as well as by the value of β, shown against f_m in Fig. 8) the performance of the instrument is unaffected.

An uncalibrated oscillator may in fact be used, provided it is first 'calibrated' at adjacent frequencies, using the potentiometer frequency-bridge at the 100c/s intervals, graphical interpolation being made within the interval. (It is assumed here that the frequency bridge is accurate—unlike that used in the model.)

(d) *Measuring Amplifier Gain and Linearity*

By comparing some fraction of the amplifier output voltage with the input, the instrument can measure, to well within 1 per cent, changes of gain due to changes of amplifier input level. To change the amplifier input level the voltage fed in common to the amplifier and potentiometer is altered, the potentiometer having been adjusted to balance the amplifier gain prior to the change of input; then the subsequent alteration of dial setting is observed, and thence the gain-change deduced.

It is to be noted that, arising from the basic circuit theorem, any change in non-linearity in the two slide-wire transformers resulting from the altered signal level will have no effect so long as the transformers have similar parameters. Moreover, as distinct from conventional methods of gain-measurement, the null-balance technique is quite uninfluenced by any distortion terms which may be generated in the amplifier (see Conclusion).

(e) *Measurement in the Presence of D.C.*

As no part of the slide-wire circuit is grounded, blocking capacitors in the potential leads will allow alternating voltages to be measured in valve or transistor amplifiers without upsetting the d.c. conditions.

(f) *Impedance Measurement In Situ*

Provided there is, in series with some branch of the tested network, a circuit element of known impedance, the input current to the branch can be measured and compared (in modulus and argument) with the input voltage to derive the input impedance of the branch.

By this means the conditions operating in, for instance, an active network can be explored without disturbing the power-supplies; a specific application is to the output stage of a feedback amplifier.

In passive networks also (see (a) below) the various local impedances can be observed in this way without disturbing the tested system.

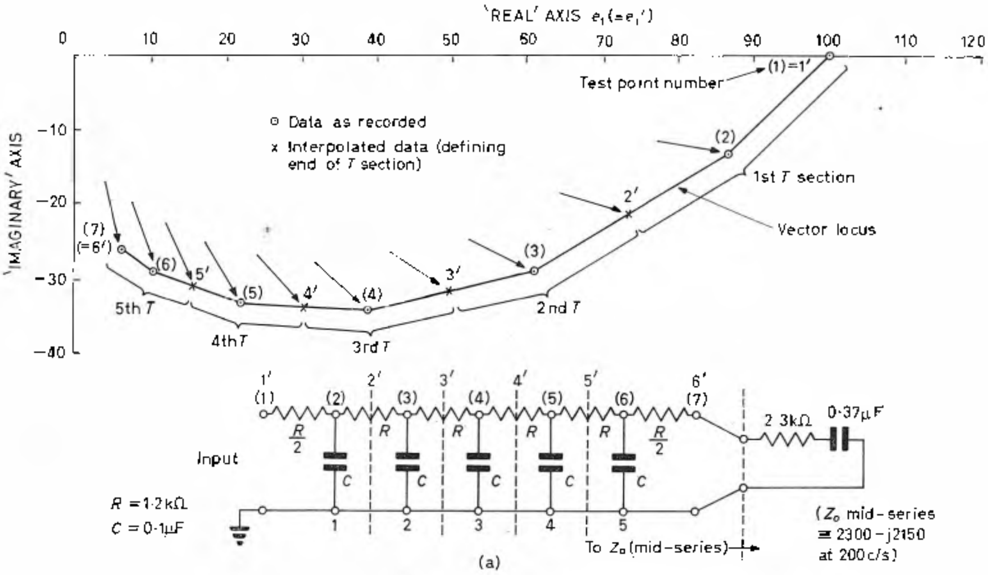
The technique can be extended to 'pseudo-impedances'[4]. By this means an effectively zero impedance can be simulated and the true short-circuit current measured in phase and magnitude—an application of special interest to transformer designers.

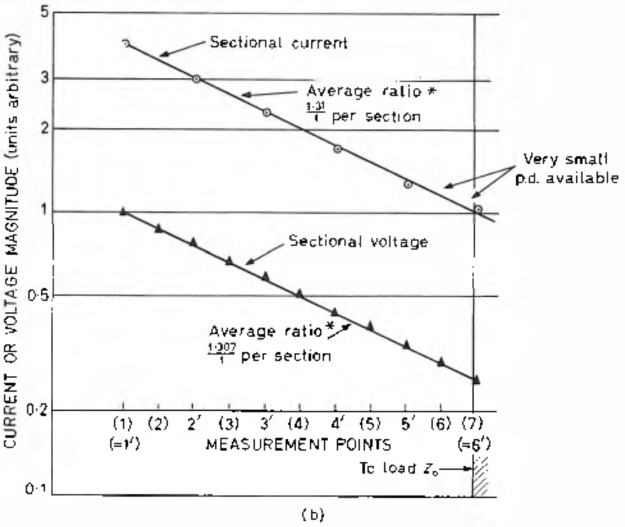
DEMONSTRATIONS OF PRECISION

(a) *RC Transmission-Line* (*'Kelvin Cable'*)

The voltage-vectors measured (at 200c/s) in all parts of a five-section artificial line are illustrated in Fig. 9(a). The subsidiary figures demonstrate the precision achieved after the α and β corrections of Fig. 8 have been made.

As an example of (f) above, the 'mid-series' and 'mid-shunt' impedances of each section of the line were deduced using the p.d. measured across the series element of each T-section to derive the current. The current vectors so derived yielded the data plotted in Fig. 9(b) which confirms the agreement required by theory between voltage and current attenuation in a symmetrical iterative network such as this.





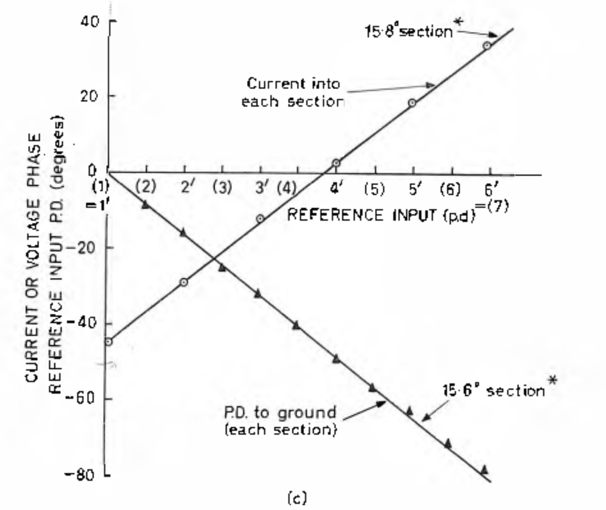


Fig. 9 (a). Network and test points. (b) Current and voltage attenuation (* *Theory predicts 1·320/1*) **(c) Current and voltage phase-shift** (* *Theory predicts 15·8°/section*)

(b) *Precision Over the Full Frequency-Range; a Circle Diagram*

In the simple *RC* network of Fig. 3(b) it can be arranged that the voltage vector e_2 lies on a circle when the input frequency is varied from 100c/s to 10kc/s, provided that at each frequency the voltage e_1 is 'normalized' in modulus and argument.

The data plotted in Fig. 3(a) were measured by the potentiometer with e_1 normalized to $X_1 = 50$ and $Y_1 =$ zero; (later the data were adjusted very slightly to bring the centre of the circle to the centre of the diagram).

The scale zero errors were removed by adjusting the controls *r* of Fig. 2 and the potentiometer data corrected for the values of α and β shown in Fig. 8 before plotting. To derive the radius precisely, the corrected co-ordinates (X_2', Y_2') of each point were squared, added and square-rooted.

The mean radius is 48·78 scale-divisions with roughly equal maximum and minimum deviations of 0·4 divisions, and standard deviation 0·25 divisions (see Fig. 3(c).

The irregularity ($\pm 2 \times 0{\cdot}2$ divisions in Fig. 3(c)) in the measured radius is in line with the $\pm 0{\cdot}2$ divisions uncertainty in slide-wide p.d. for this quantity appears twice during the measurement—once when e_1 is normalized, and once when X_2 and Y_2 are measured. Defects in individual measurements of the radius, however, can be seen to approach $\pm 2 \times 0{\cdot}2$ divisions only twice, and are in general very much smaller. The maximum possible defect would have been $\sqrt{2}$ times greater if it had happened that both X' and Y' suffered discrepancies of the same sign simultaneously; the lower value found suggests that some degree of random addition of such errors must have taken place.

Indirectly this demonstration proves the reliability of the α and β data of Fig. 8, for there is no repetition in Fig. 3(b) of the irregularities which are so obvious in Fig. 8.

Conclusion

It has been shown that provided the instrument is self-checked and the α and β corrections derived and made use of, the reliability of the data is set largely by the discrete steps in which the p.d. can be picked off the slide-wires.

From time to time other factors, not dealt with above, have to be taken into account; for example, the unwanted currents flowing from the tested network to ground through the transformer winding-to-earth capacitance. There are so many ways in which these can cause minor errors that it is simplest to give general instructions only for avoiding them.

The most obvious is that the secondary windings of the slide-wire transformers should have as small a capacitance as possible to ground (or to the shield, if provided), consistent with adequate secondary voltage. Here it has to be remembered that a relatively large secondary leakage inductance—due to physical separation between windings, or winding and shield, for instance—is likely to be a good rather than a bad design feature, the power-efficiency of these transformers being of secondary importance provided they can be made in similar pairs. In this respect the choice of a Post Office line transformer for the model instrument, with excellent conventional performance, tended to degrade rather than improve the performance of the potentiometer, at least at the higher working frequencies.

Even so, the winding to earth capacitance of the model transformers causes trouble only when the test object has a considerable p.d. to earth, as for instance when the potential difference under examination forms some small part of a chain of impedances earthed at one end; the stray capacitive current flowing to earth through the slide-wires and transformers then modifies the X and Y p.d.s. and causes errors. Naturally, the percentage error in readings will be greatest when the p.d. measured is much smaller than the interfering signal, but it will also increase with frequency. Tests show that the 'balanced' circuit of Fig. 2 gives good immunity because the capacitance of the slide-wire transformer secondaries are 'balanced' to their screen. (It is found no advantage to earth the screen physically, however.) However, when an interfering p.d. to earth equal to the full slide-wire p.d. exists between the measuring point and earth, and the frequency is below 1kc/s, readings will not be significantly affected.

If separate terminals from the transformer screens were provided which could be connected to a Wagner earth adjusted to suit conditions it might be possible to remove entirely the stray current to earth.

It is found that the potential leads can be extended through quite large resistances (e.g. 10kΩ) without upsetting the accuracy of p.d. measurement, even at 10kc/s, but this is true only when stray capacitive currents are negligible; that is to say, when the tested potential has roughly the same p.d. to earth as the potential leads.

When the instrument is in use for measuring the p.d.'s at the terminals of a transducer such as a probe-microphone—or of the associated amplifier—stray currents of this kind can be made negligible at all frequencies.

As this article is concerned with the a.c. potentiometer measurement principle, auxiliary gear has received little attention, but it should be obvious that the precision of measurement can never exceed the precision of 'null' balance. To achieve the accuracy demonstrated in the diagrams a c.r.o. was always used and on occasion a selective amplifier as well. The c.r.o. beam was deflected horizontally by the driving oscillator, and at the same time deflected vertically by an amplifier driven from the 'detector' terminals of the potentiometer.

This arrangement had two great advantages; first, any noise or hum in the tested system appeared as a non-coherent interference superimposed on the steady pattern on the tube-face which merely masked but could not alter the null-balance point; secondly, any non-linearity distortion in the driving oscillator did not degrade the null balance, a characteristically looped figure appearing which closed to a curved line when balance was reached; at the same time the extremities of the loop fell on a horizontal line. In general it happened that the closing of the loop depended on one slide wire setting and the 'tilt' of the diagram on the other; thus 'convergent' adjustments were achieved rapidly even when the signal-to-noise ratio was substantially less than unity.

Used in this way the a.c. potentiometer is the only practical means of measuring single-frequency probe-microphone voltages in the presence of ambient room noise; it was for this purpose that the instrument was originally developed.

Because it is so important, a statement made at the outset will be repeated: accurate Cartesian measurements can be made even with an imperfectly constructed instrument because the necessary corrections can always be derived by self-calibration, the basic requirement being that the X and Y dials shall bear a calibration which has a one-to-one relation to the p.d.'s picked off the slide-wires. As a rider to this, the slide-wires should be evenly wound and have as many turns of wire as possible: for this reason a high resistance, many-turn, high-resistance, rheostat can be used with advantage if shunted down to the required overall resistance.

A potentiometer that reads in polar co-ordinates has been designed on the principles set out in this article and will be described in the Complete Specification subsequent to Ref. 2.

Acknowledgment

Most of the work reported here was done at the BBC Laboratories at Kingswood Warren and acknowledgment is made for permission to publish it; the basic theoretical work of Ref. 3 was done at Signals Research and Development Establishment, Christchurch, while under the Ministry of Supply.

APPENDIX

(A) PHASE-SHIFTING ARRANGEMENTS

Without a subsidiary phase-shifter potentiometer measurement of voltage ratios is so cumbersome and open to so many errors as to be impracticable. On the other hand, with the assistance of a phase-shifter one of the voltages can be represented by a simple number $(X_1 + j0)$ and the ratio can then be read off directly from the dials in Cartesian co-ordinates as $(X_2 + jY_2)/X_1$. (See Introduction.)

A very simple one will serve, for the requirements are elementary. No phase or magnitude calibration is needed; the performance need not be consistent over the working frequency range; finally, the input and output impedances may be allowed to vary while phase adjustments are being made. The design merely has to provide for smooth and continuous adjustment of phase and magnitude over the required range. The ideal arrangement has 360° coverage, or alternatively slightly more than 180°, with a reversing switch at the output to complete the circle. A simple RC assembly can meet such conditions.

In Fig. 1 the phase-shifter is required to shift the phase and magnitude of one of the voltages (e_1, e_2, etc) in the test-object Z, relative to the slide-wire p.d.'s. Now as this is a purely relative quantity the performance of the associated potentiometer cannot be affected by variations of the input impedance of the phase-shifter. (Imagine, for example, that the input terminals of the phase-shifter are shunted by some low impedance; the potentiometer, being in parallel, will be affected similarly and tested and testing p.d.'s will shrink equally; consequently the potentiometer null-balance will not be disturbed.) It follows that changes of the input impedance of the phase-shifter, arising during adjustments will be permissible. Put in other words, the phase-shifter can be designed as though it were supplied from a source of zero internal impedance—this markedly simplifies the circuit.

Variations of the output impedance during adjustment merely make its use slightly more tedious.

Finally, the phase-shifter network can have an overall voltage loss of the order of 1/5. This arises because the highest possible voltage ever required from the phase-shifter will be that due to both X and Y dials set at maximum, and in the main text (USAGE: GENERAL (b)) it is shown that E_s is of the order 1/5 of the input voltage (E_1) supplied to the potentiometer. As E_1 is also the input of the phase-shifter this implies that it need never supply more than $E_1/5$.

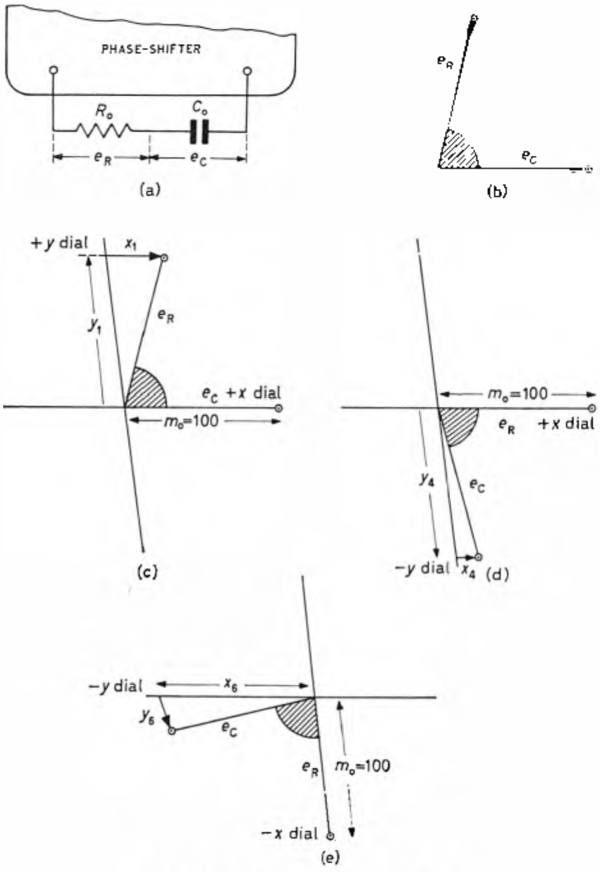


Fig. 10. Three point calibration

It may be objected that, when the performance of the test-object (e.g. an amplifier operating very near to overload) is conditioned by the absolute magnitude of the voltage delivered to it by the phase-shifter, the adjustments of voltage-loss through the phase-shifter in the course of setting up the null-balance of the potentiometer will upset the testing condition. This is true. But, by altering the voltage supplied to the assembly from the testing oscillator, the voltages in all parts of the assembly can be adjusted simultaneously without upsetting any balance conditions which may exist. Thus after a first null balance has been established, the input signal is set to produce in the test-object the required operating conditions; only residual adjustment of the X and Y dials will then be needed, and very often the phase-shifter does not have to be altered at all.

(B) RAPID METHOD OF CALIBRATION

Instead of carrying through the eight-point test of the main text, which may take 10 to 15min, the following 'three-point' test can be substituted provided the scale-zeros are corrected.

As the test will give two separate estimates of α and β, which act as checks, and takes only 2 to 3min it can be used to prove that the instrument is in working order. The two values of α and β will never precisely agree because of the limited voltage-discrimination of the slide-wires but the average may be taken as representative.

Unlike the eight-point test which, in order to validate the 'averaging' procedure, requires that e_C/e_R (see Fig. 10(b)) should be approximately equal to 1·0, 90°, this three-point test is valid for any value whatever of e_C/e_R. For purely practical reasons a ratio near to the ideal is preferred as it helps to reveal, at a glance, any gross errors of measurement procedure.

As in Fig. 7, the X and Y axes of the potentiometer have been shown set obliquely, but exaggerated, to show the effect of a negative value of α (see example below).

The calibration is carried out exactly as in the eight-point test (see Table 1) but only three of the eight tests there listed are made: Nos. 1, 4 and 6, which can be seen to correspond to Figs. 10(c), 10(d) and 10(e). It should be pointed out that the 'alignment' figure $M_o = 100$ used in Fig. 10 is not essential; any round number will do, but 100 makes calculation of α and β very simple. The following equations are true for any value of M_o.

Theory

It can be shown that if the X and Y slide-wires carry currents I_1 and I_2, respectively related by

$$I_2/I_1 = (\alpha + j\beta)$$

then, using tests (1) and (4) of Table 1:

$$\alpha_1 = -(1/2).(X_1/Y_1 + X_4/Y_4) \quad \text{(B1)}$$

$$\text{and } (\alpha_1^2 + \beta_1^2) = -(M_o)^2/(Y_1 . Y_4) \simeq (\beta_1)^2 \quad \text{(B2)}$$

Or, using tests (4) and (6) of Table 1:

$$\alpha_2 = -(1/2).(X_4 + Y_6)/Y_4 \quad \text{(B3)}$$

$$\text{and } (\alpha_2^2 + \beta_2^2) = X_6/Y_4 \simeq (\beta_2)^2 \quad \text{(B4)}$$

It is important to notice that the X and Y data read off the dials have to be inserted into these equations without the changes of sign called for in Table 1. It follows that X_6, Y_4 and Y_6 are essentially negative, and as Y_1 is essentially positive, equation (B2) is viable.

Example Test made at 8kc/s for use in Fig. 8:

Potentiometer data

$M_o = 100$ $X_1 = +3{\cdot}9$ $Y_1 = +104{\cdot}6$

$X_4 = -1{\cdot}0$ $Y_4 = -98{\cdot}85$ $X_6 = -96{\cdot}75$

$Y_6 = -3{\cdot}5$

Whence: from equations (B1) and (B2)

$\alpha_1 = -2{\cdot}43 \times 10^{-2}$ $\beta_1 = 0{\cdot}981$

and from equations (B3) and (B4)

$\alpha_2 = -2{\cdot}26 \times 10^{-2}$ $\beta_2 = 0{\cdot}989$

which yields average values:

$\alpha_{av} = -2{\cdot}345 \times 10^{-2}$ $\beta_{av} = 0{\cdot}985$

Notes

The negative sign of α implies that, as in the figures, the top right quadrant of the Cartesian axes exceeds 90°. The negative sign of X_4 disagrees with the positive sign suggested in Fig. 10(d) because the reference voltages used in the test example above were much more closely in quadrature than in Fig. 10(b).

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A Large-Signal Transistor Analogue

By G. G. Bloodworth, M.A., D.U.S., A.M.I.E.R.E., and D. N. Axford*, M.A., M.Sc.

An analogue has been constructed which imitates the behaviour of the transistor model which is normally assumed in circuit design, except that the effective cut-off frequency is scaled down. The electrical behaviour of the base region is represented by a linear passive network, and the necessary non-linear voltage transformations at the junctions are generated from the characteristics of diodes. The analogue can be used to demonstrate the significance of the various parameters of a transistor and to predict its performance in circuit applications.

(Voir page 493 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 500)

THE behaviour of any real transistor is so complex that it cannot be represented exactly by an analogue, nor, indeed, described completely by a finite number of measured parameters. It is, however, possible to identify certain characteristics, for example the current gain β, which are essentially constant over certain ranges of operating conditions, and others with a variation which is almost exponential over wide ranges. Thus an acceptable model is obtained, supported by semiconductor theory and described by equations involving parameters which are constant, and therefore usefully measurable.

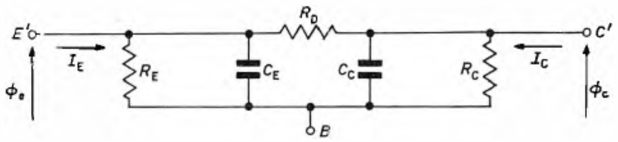

Fig. 1. The linear base region model

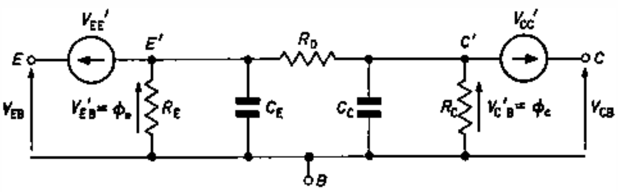

Fig. 2. The intrinsic transistor model

From such a model, transistor behaviour can be predicted either by solving the equations, or by building an analogue. Both methods of attack become progressively easier as the model is simplified, but each simplification involves restricting the operating range or accepting increased uncertainty in prediction. Eventually, theoretical analysis becomes so easy that no analogue need be considered. Thus, if the transistor is to amplify signals with very restricted amplitude and bandwidth, it can be represented simply by four complex numbers, for example the admittance parameters appropriate to the configuration.

Between these extremes there is a model which is accurate enough for many applications, but involves equations which are non-linear, and therefore not easy to deal with. For this model an analogue has been constructed which is relatively easy to use, and is adjustable so that it can represent a wide range of transistors. The analogue can be used to demonstrate the meaning and effect of each of the characteristics of a transistor, and also to predict its circuit performance.

The Transistor Model

The basic assumption made is that the processes of charge transfer and recombination within the base are linear, and that the emission efficiencies of the junctions are independent of current. This assumption implies operation at a low level or over a restricted range of current. The effects of base width modulation and junction breakdown are also disregarded in the basic model. With these assumptions it has been shown[1] that the transistor can be represented by the equivalent circuit shown in Fig. 1. This circuit is applicable to graded-base transistors, as no restrictions are placed on the variation of impurity concentration throughout the base or on the geometrical arrangement. A pnp structure will be assumed throughout for convenience.

The voltages applied to the π-section represent the excess charge products ϕ_e and ϕ_c in the base adjacent to the emitter and collector junctions respectively, that is the product of the densities of electrons and excess holes in these regions. The currents entering at the nodes E' and C' are then equal to the net currents flowing into the base through the two junctions, and the charge carried by the capacitances equals the charge stored in the base.

The excess charge products are determined by the bias voltages V_{EB} and V_{CB} (positive for forward bias) according to the equations

$$\left.\begin{aligned} V_{E'B} &= \phi_e = \phi_o(\exp qV_{EB}/kT - 1) \\ V_{C'B} &= \phi_c = \phi_o(\exp qV_{CB}/kT - 1) \end{aligned}\right\} \quad \ldots\ldots\ldots \quad (1)$$

Here q is the electronic charge, k is Boltzmann's constant, and T is the absolute temperature. The constant ϕ_o is given by:

$$\phi_o = qp_N n_N$$

where $p_N n_N$ is the product of the equilibrium densities of holes and electrons, which is constant throughout the base. This model gives the same direct currents as the well-known equations of Ebers and Moll, and it is assumed that during transient changes the base charge distribution does not lag appreciably behind the junction voltages. The grounded-base current gain in the normal direction is given by

$$\alpha_N = \alpha_{oN}/(1 + jf/f_{1N})$$

where $\alpha_{oN} = R_E/(R_E + R_D)$

and the cut-off frequency is given by

$$f_{1N} = (R_E + R_D)/2\pi R_E R_D C_E = 1/2\pi\alpha_{oN} R_D C_E \quad \ldots. \quad (2)$$

For inverse operation, R_C and C_C replace R_E and C_E and there are corresponding expressions for α_{oI} and f_{1I}.

For any particular transistor, the ratios between the values of the five components are determined by the four quantities α_{oN}, α_{oI}, f_{1N}, f_{1I}. The actual values are determined[1] by the voltage ϕ_o, which is chosen arbitrarily, together with the transistor current for any one set of bias conditions, for example the collector saturation current I_{CBO}.

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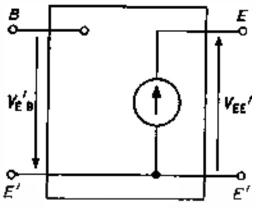


Fig. 3. The non-linear amplifier

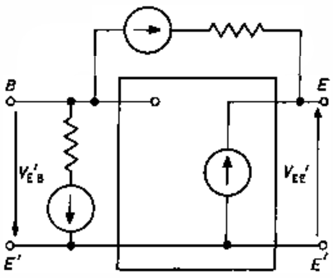


Fig. 4. Generators applied to the amplifier

The Transistor Analogue

In the analogue which has been constructed, the above equivalent circuit is represented by real components, and ϕ_e and ϕ_c are real voltages at the nodes E' and C' relative to the base terminal. To represent the emitter junction correctly the analogue must be arranged so that the currents flowing into E' and into the external emitter terminal E are always equal, and yet the voltages at these nodes relative to B must obey equation (1). This requirement has been met, as shown in Fig. 2, by introducing a voltage generator between E and E' with low impedance and voltage equal to the required difference $V_{EE'}$ given by

$$\left.\begin{aligned} V_{EE'} &= V_{EB} - V_{E'B} \\ &= (kT/q)\ln(1+V_{E'B}/\phi_o) - V_{E'B} \end{aligned}\right\} \quad \ldots\ldots\ (3)$$

This voltage generator is the output of a non-linear amplifier whose input voltage is $V_{E'B}$, as shown in Fig. 3. The output impedance must be very low, because the voltage produced must be due to the current flowing through R_E, R_D etc., but independent of the current in the branch EE'. This is evident, since the junction voltage at any instant is determined directly by the excess charge product in the base, and not by the emitter current. The input resistance of the amplifier should be very high as it is in parallel with R_E, though in practice R_E could be chosen to take account of it.

Suppose another voltage generator is connected between the external terminals E and B instead of, or in addition to, the generator applied between B and E', as shown in Fig. 4. If both generators had zero impedance the corresponding terminal voltages would be independent variables and so the equations (3) could not be obeyed in general. However, a finite impedance is provided between B and E' in the analogue by R_E, R_D etc., and consequently the equations are obeyed, within the dynamic range of the amplifier, even if an ideal voltage generator is connected between emitter and base.

The same equations must be satisfied at the collector junction by the voltages $V_{CC'}$, V_{CB}, and $V_{C'B}$, because ϕ_o has the same value on both sides. A second identical amplifier is used for this purpose.

The effects of stored charge in the depletion layers are simulated in the analogue by ordinary capacitors connected between the base and the other two terminals. According to theory[2] the stored charge is proportional to $U^{\frac{1}{2}}$ for an abrupt junction or $U^{2/3}$ for a graded junction, where U is the magnitude of the potential hill. A typical curve is shown in Fig. 5. The ordinary capacitor gives a straight line passing through zero bias voltage. As the total change in charge under transient conditions is the most important factor, this representation is reasonably accurate provided the straight line is parallel to the chord of the appropriate curve over the voltage range to be traversed. This simplification is made almost invariably in theoretical calculations of transient responses, and so the analogue accords with the usual model.

The analogue is completed, as shown in Fig. 6, by adding the depletion capacitances C_{dE} and C_{dC}, and the lumped resistances of the three regions of the transistor $r_{bb'}$, $r_{ee'}$, and $r_{cc'}$. The variations of these resistances with current due to conductivity modulation is disregarded; this again implies a restricted range of operation.

To avoid errors due to stray reactances and the finite bandwidth of the amplifiers, the cut-off frequency of the analogue is reduced by multiplying each of the four capacitors by a suitable constant. As the magnitudes of any reactances of the form ωL and $1/\omega C$ in the external circuit must not change, each inductance and capacitance must also be multiplied by the same scaling factor.

The Non-Linear Function Generator

The required non-linear voltage transformation of equation (3) is most conveniently generated from a pn junction diode whose characteristic is, ideally,

$$I_D = I_S\,(\exp qV_D/kT\ -\ 1)$$

or

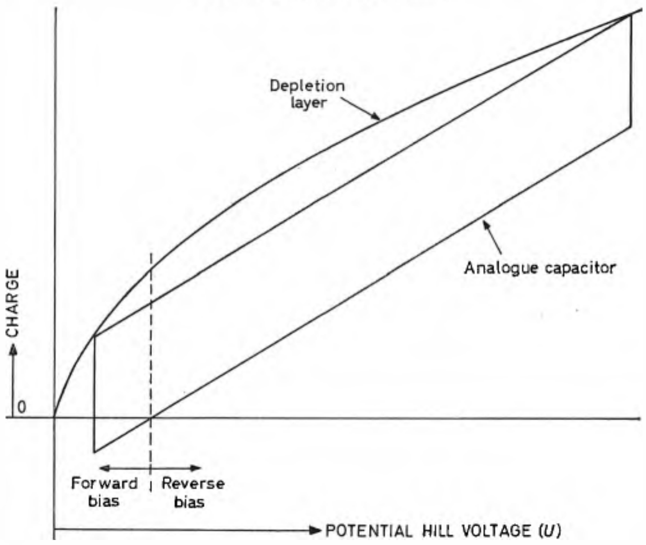
$$V_D = (kT/q)\ln(1 + I_D/I_S)$$

Thus if a current proportional to $V_{E'B}$ is passed through a suitable diode, the voltage developed provides the first term in equation (3) and $V_{EE'}$ can then be generated by subtracting $V_{E'B}$. This operation can be realized with the arrangement shown in Fig. 7.

The first amplifier has high open-loop gain, and uses series-current feedback through R_F to maintain the voltage across this resistor closely equal to $V_{E'B}$. The current in R_F also flows into the emitter of a transistor and the collector current from this transistor flows into the diode. Thus the diode is driven from a source of high impedance, represented by R_L, with a current equal to $V_{E'B}/R_F$ if the chord resistance of the diode is much smaller than R_L. The diode is also driven by the output voltage $V_{E'B}$ of the buffer amplifier, which, considered alone, produces a diode current $V_{E'B}/R_L$. This current is negligible provided R_L is much greater than R_F, and this condition is easily met.

The buffer amplifiers have high input impedance and low output impedance, and are arranged so that the output voltages follow the respective input voltages as closely

Fig. 5. Depletion layer charge



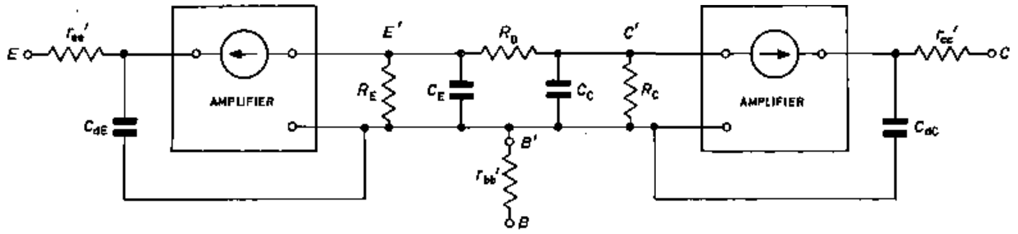


Fig. 6. The complete analogue

as possible. The connexion shown in Fig. 7 enables $V_{EE'}$ to be produced, as required, by subtracting $V_{E'B}$ from V_D.

When the diode is forward biased, the resistance R_L can be ignored and the current in the diode is very nearly

$$I_D = V_{E'B}/R_F = I_S\,(\exp qV_D/kT - 1)$$

Assuming for simplicity that $V_{C'B} = 0$, then from the equivalent circuit

$$V_{E'B} = I_E R_D R_E/(R_D + R_E) = I_E R_D\,\alpha_{oN}$$

and also, assuming perfect subtraction in the amplifier, V_{EB} equals V_D.

Therefore:

$$V_{E'B} = I_E R_D \alpha_{oN} = I_D R_F = R_F I_S\,(\exp qV_{EB}/kT - 1) \quad \ldots\ldots\ldots\ldots (4)$$

A comparison with equation (1) shows that the product $R_F I_S$ is the voltage representing ϕ_o.

If the emitter current of the transistor being represented, with $V_{CB} = 0$, is expressed in the form:

$$I_E = I_{EBS}\,(\exp qV_{EB}/kT - 1) \quad \ldots\ldots\ldots (5)$$

then, from equations (4) and (5),

$$\alpha_{oN} R_D\, I_{EBS} = R_F I_S \quad \ldots\ldots\ldots\ldots\ldots (6)$$

The ratio of the diode current I_D to the emitter current I_E with $V_{CB} = 0$ is clearly independent of V_{EB} and is given by $\alpha_{oN} R_D/R_F$.

The quantities R_F, I_S and R_D must be chosen to satisfy equation (6) for each transistor being represented. A convenient approach is to choose the maximum value of $V_{E'B}$ equal to say 5V, so that small fractional changes are not swamped by amplifier drift. The maximum value of I_E is determined by the dynamic range of the practical amplifier used, and this value determines R_D, from equation (4). R_F can be varied over a limited range—say a factor of two—without affecting the accuracy of the feedback amplifier. The diode current I_S can be chosen from a wide range of diodes, but cannot, of course, be adjusted. Thus in a practical instrument, switching between different diodes or combinations of diodes in parallel will give a coarse control of I_{EBS}, while a continuously variable R_F provides a fine control.

As an example, consider a transistor for which the emitter current I_E is 10mA when V_{EB} is 300mV. Taking $\alpha_{oN} = 0{\cdot}98$ and $V_{E'B(max)} = 5$V, then $R_D = 510\Omega$.

Now, if it is convenient to choose $R_F = 200\Omega$

$$I_S = \alpha_{oN}\, I_{EBS} \,.\, R_D/R_F = 5/2\,.\,I_{EBS}$$

Thus a suitable diode is one for which the current is 25mA when it is forward biased by 300mV. The best diode available is selected, and final adjustment made by varying R_F.

The above equations do not hold for reverse bias, because the chord resistance of the diode rapidly becomes large compared with R_L. When I_D becomes negligible the output voltage is given by

$$V_{EE'} = V_{E'B} \,.\, R_L/R_F$$

and therefore

$$V_{EB} = V_{EE'} + V_{E'B} = (1 + R_L/R_F)\,V_{E'B} \quad \ldots\ldots (7)$$

The emitter current of the analogue (with V_{CB} zero) is given by

$$I_E = V_{E'B}/\alpha_{oN} R_D = V_{EB} R_F/\alpha_{oN} R_D (R_F + R_L)$$

Thus the emitter leakage current is directly proportional to the reverse bias voltage V_{EB}, due to the effect of R_L. The slope resistance of the collector junction with appreciable reverse bias is determined in the same way by the value of R_L in the second amplifier. This slope resistance is independent of emitter current when the emitter is forward biased so that the analogue is operated in the normal active region of its characteristics.

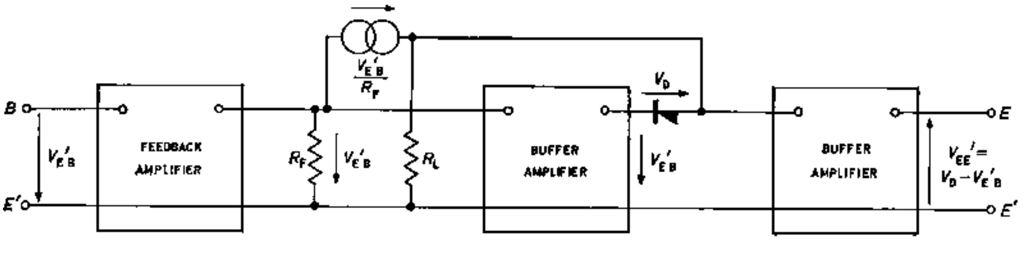
The basic model being represented does not allow for base-width modulation (the Early effect) which causes the slope conductance of the grounded base output characteristics to vary, roughly in proportion to the emitter current. The slope is magnified in the grounded emitter configuration, becoming an important parameter of the transistor, but the same current dependence appears, as shown in Fig. 8. For this Figure the value of R_L has been chosen so that the constant slope of the analogue is made equal to the average slope for the transistor being represented, over the range of interest. The conductance is then too low at high currents and too high at low currents, magnifying the leakage current. However, taken as a whole, the representation is better than that of the basic theoretical model. When R_L is chosen in this way, feedback to the input is also introduced which simulates roughly the voltage feedback effect of base-width modulation: this is analysed in the Appendix.

A General Purpose Analogue

An analogue has been constructed with controls calibrated so that a wide range of transistors can be represented. This section deals with the design and operation of the controls used when the emitter is to be forward biased, and the collector reverse biased; those for inverse operation are exactly similar.

The static characteristic of each junction may be set to simulate any given transistor by means of a coarse control which switches between different diodes, and a fine control which gives a continuous variation in R_F. As the

Fig. 7. Arrangement of non-linear amplifier



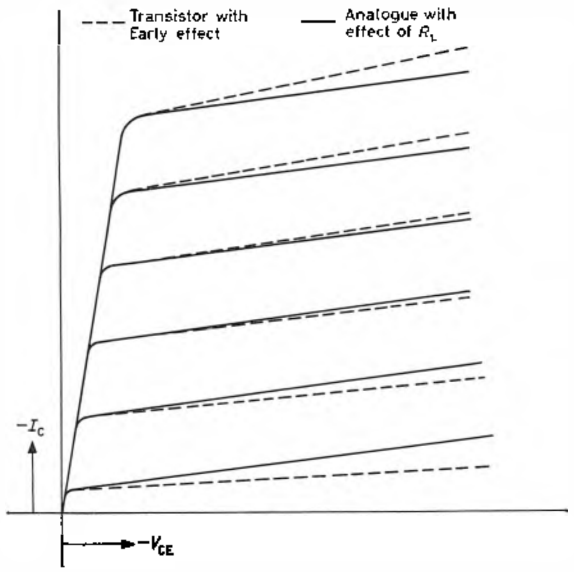


Fig. 8. Output characteristics with grounded-emitter, for equal steps of base current

normal range of forward bias voltages used is much greater for silicon than for germanium devices, silicon diodes are used to generate the static characteristics for silicon transistors, and germanium diodes for germanium transistors. A master switch is provided which changes the entire set of diodes according to the transistor type required.

If the diodes are properly chosen, it is possible to calibrate the coarse and fine controls so that any required emitter current can be selected for a given emitter-base voltage. The voltage chosen for germanium devices is 300mV, and for silicon 600mV. In each case it is assumed that if the analogue is set up to give the correct current with this emitter-base voltage applied to it externally, the diode characteristic will generate an analogue input characteristic of satisfactory shape. Each position of the coarse control switch could be labelled with the current carried by the appropriate diode at the reference voltage, and the fine control, which gives for example a variation in R_F from 125Ω to 250Ω, could be calibrated in terms of the multiplying factor R_F/R_D; here R_D is 500Ω, and so this factor varies from $\frac{1}{4}$ to $\frac{1}{2}$. In practice it is more convenient to calibrate the continuously variable control from one to two, and to label the coarse control witl one-quarter of each diode current. The current set by these controls is then strictly $\alpha_{oN} I_E$ (from equation (4)) a the appropriate voltage. However, the error is small i the factor α_{oN} is close to unity, and a suitable correctior can be applied to the controls if it is not. The reference voltages have been chosen quite arbitrarily and, i required, external meters can be used to set the emitte current accurately to any required value at any othe chosen voltage.

The resistance R_E is given by $\beta_{oN} R_D$ where $\beta_{oN} = \alpha_{oN}/$ $(1 - \alpha_{oN})$. As the value of R_D is fixed, a variable resist ance for R_E can be calibrated directly in terms of curren gain.

The choice of the effective cut-off frequency for the analogue is not critical, but a suitable value is abou 100c/s; this is well within the bandwidth of the amplifier described in the next section and yet sufficiently fast fo convenient repetitive display of the transient waveforms The effective cut-off frequency of a transistor depend not only on the device itself, but also on its configuration and the source and load impedances. The same is true of the analogue, and so the capacitances within it mus be variable if the cut-off frequency is to be held clos to 100c/s, even when representing the same transistor i different circuit configurations.

The required values of the capacitances C_E and C_d are determined as follows. From the cut-off frequency f_{IN} and collector capacitance of the appropriate tran sistor, a rough estimate is made of the effective cut-of frequency f_c in the circuit of interest.

The scaling factor $f_c/100$ is to be applied to externa circuit reactances and to the capacitances within th analogue. Using the additional suffix A to distinguish th values for the analogue, the required capacitances ar given (from equation (2)) by

$$\frac{1}{2\pi\, \alpha_{oN} R_D C_E} = \omega_{INA}/2\pi = f_{INA} = (f_{IN}/f_c)\ .\ 100 \quad \ldots (8$$

Thus:

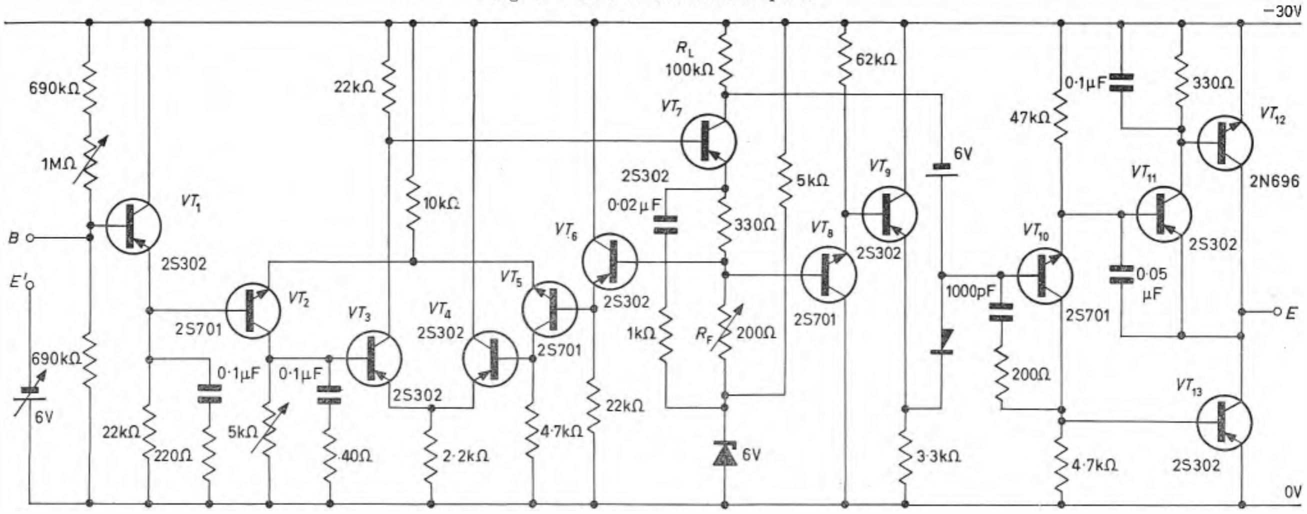
$$C_E = f_c/100\ .\ \frac{1}{2\pi\, f_{IN} R_D \alpha_{oN}},$$

and also

$$C_{dCA} = \frac{C_{dC}\ .\ f_c}{100} \quad \ldots\ldots\ldots\ldots\ldots\ldots (9$$

The switch used to select C_E can be calibrated directl in terms of the cut-off frequency f_{INA}, from equation (8) because R_D is constant and variations in α_{oN} are hardl

Fig. 9. Circuit of non-linear amplifier



significant compared with variations in practical capacitances. Normally f_c falls within the range f_{IN} to $f_{IN}/100$, and so f_{INA}' is made variable from 100c/s to 10kc/s. Then from equation (8) C_E must vary between 3·3μF and 0·03μF approximately. The depletion capacitance C_{dCA} can also be switched between several values, each representing a different value of the product $C_{dC} . f_c$. This product is greatest for a fast transistor, for which, say, C_{dC} is 5pF and f_c is 10^8c/s, and least for a very slow transistor, current driven with grounded-emitter, for which C_{dC} might be 100pF and f_c approximately 10^4c/s. Substituting these values into equation (9) gives limiting values of 5μF and 0·01μF for the range of C_{dCA}; this range is divided into suitable steps, each one labelled with the actual value used for C_{dCA}.

The Amplifier Circuit

An amplifier circuit which has been used for each junction is shown in Fig. 9. This is suitable for operation up to currents of 10mA, and reverse voltages of 10V, and might be developed for operation at higher power levels.

The feedback amplifier consists of two balanced amplifier stages preceded by a grounded collector input stage with an input resistance of about 250kΩ. This is much greater than the maximum value of R_E and so does not affect the π-network. The differential signal is amplified, and applied to the transistor VT_7 and the series-current feedback from R_F in the emitter lead ensures that the voltage across this resistor closely follows the input voltage.

The current feedback also increases the collector resistance of VT_7. At low currents this resistance is about 400kΩ, and the effective value of R_L, when the diode is reverse biased, is given by the parallel combination of 400kΩ and 100kΩ, that is by 80kΩ. The quiescent current in VT_7 is about 170μA, which gives 6V between the base of VT_6 and the collector of VT_7. The diode is connected to the same voltage as the base of VT_6, so that $V_{E'B}$ is subtracted from V_D, and a 6V dry battery is connected in series so that V_D is zero for zero signal.

The first buffer amplifier consists of an npn-pnp pair of transistors cascaded with grounded collector and so arranged that the base-emitter voltages cancel for zero signal. The input resistance is over 100kΩ, and the output resistance, which varies with the current injected from the diode, is typically a few ohms, and can be neglected compared with the diode chord resistance. When the input voltage is varied over a wide range, the output voltage does not follow exactly because when the current in VT_8 decreases, the current in VT_9 increases, and so the base-emitter voltages diverge and no longer cancel.

The resultant error in $V_{EE'}$ is normally negligible, but may reach 20mV when the diode current is 5mA. This distortion of the effective diode characteristic can be considerably reduced, if required, by replacing this buffer amplifier by a duplicate of the final amplifier.

The final buffer amplifier must have a maximum error between input and output voltages of 5mV for an output current range of 10mA, and a voltage range of at least 10V. Unless this specification is satisfied the complete analogue acts incorrectly in reverse bias, since the currents flowing within the π-network caused by the error at the reverse biased junction become significant. The amplifier[3] formed by VT_{10} to VT_{13} satisfies these conditions. The output voltage level is clamped closely by making the currents in VT_{10} and VT_{11} increase and decrease together when the input voltage varies, while the majority of the current load is taken by VT_{12}. The resistances connected between base and emitter of VT_{12} and VT_{13} are chosen to compensate for unmatched transistors, and are preset to balance the base-emitter voltages of VT_{10} and VT_{11}. The output resistance is about 0·5Ω, and the input resistance exceeds 1MΩ.

The capacitive correcting networks shown in Fig. 9 are used to prevent oscillation in the amplifiers, without reducing the bandwidth of the complete system below 1kc/s. The resistor R_F may be varied from 100Ω to 300Ω without causing oscillation. Drift due to changes in temperature is reduced by using silicon transistors throughout, and arranging them in balanced pairs in the feedback amplifier, and complementary pairs in the buffer amplifiers. The transistors are all mounted in a substantial heat sink to reduce short term temperature variations, and over a period of an hour the zero setting is maintained within a few millivolts.

Each amplifier must initially be set so that E, E' and B are all at the same voltage, and the diode current is zero. A current meter is switched temporarily in series with the diode, and the following procedure is used. The terminals E' and B are shorted together, and the voltage at this point is adjusted to set the diode current to zero. The corresponding voltage at E is noted, and the voltage at $E'B$ is reset to this value. By varying the resistance in series with the collector of VT_2 to alter the balance of the differential amplifier the diode current is again set to zero, and the voltage at E is made equal to that at $E'B$. This equality is checked by direct measurement of the difference. Finally the connexion between E' and B is removed, and the base bias resistance of VT_1 is adjusted to reset the diode current to zero again. The current meter is then removed. This procedure avoids the connexion of a voltmeter to the high resistance terminal B, and an ordinary moving-coil instrument can be used.

The voltage supply at E' should be of very low impedance, because the output voltage generated between E and E' must be independent of current. A suitable supply is the output of a buffer amplifier similar to that producing the voltage at E, with an input voltage which can be adjusted slightly using a variable resistor. Alternatively, a commercial stabilized supply may be used at E'.

Applications of the Analogue

In the teaching of transistor electronics, the analogue may be used to show the basic characteristics of the device either in a classroom demonstration by the lecturer, or in the laboratory as part of a practical exercise by the student himself. The significance of the various parameters is more readily appreciated when they can be introduced and varied separately, without interference from stray reactances or the many marginal effects which distinguish individual transistors.

This aid should be particularly valuable if the course is so arranged that the model chosen as the intermediary between the initial semiconductor equations and the final graphical characteristics and small-signal equivalent circuits is the particular large-signal equivalent circuit used here. The π-section is engraved on the face of the instrument, and the variable components are arranged accordingly, so that when the student varies, say, β_N, he is obviously controlling the recombination current drawn from the emitter. A special advantage is the availability of ϕ_e and ϕ_c as voltages which can be monitored, since in this way the transient behaviour of the junctions and of the base can be studied separately. Thus the analogue not only simulates the transistor in a circuit, but at the same time enables the student to investigate the internal mechanism.

The analogue described here has also been used in a detailed study[4] of the transient response of analogue gates using as switches transistors driven heavily into saturation. Additional components were added to the π-network to allow for charge storage in the remote regions of the base. With very little calculation it was possible to establish the chain of events during transient changes, and the relative importance of the various parameters, so that a suitable type of transistor could be chosen and evaluated. A similar economy of effort should be possible in other investigations.

Acknowledgments

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Appendix

Approximate Simulation of the Early Effect

It will be shown that the effect of the finite value of R_L in the amplifier generating each junction voltage is to simulate the effects of base-width modulation correctly in magnitude and sign over a limited operating range.

Let $V_{EB}/V_{E'B} \equiv X$; and $V_{CB}/V_{C'B} \equiv Y$

In terms of the non-linear functions X and Y, the large-signal hybrid parameters of the analogue with grounded base (from Fig. 2, ignoring capacitances) are as follows:

$$V_{CB} = 0; \quad \begin{cases} V_{EB}/I_E = H_{11} = \dfrac{XR_ER_D}{R_E + R_D} \\ I_C/I_E = H_{21} = \dfrac{-R_E}{R_E + R_D} \end{cases} \qquad \ldots\ldots\ldots\ldots (10)$$

$$I_E = 0; \quad \begin{cases} V_{EB}/V_{CB} = H_{12} = \dfrac{XR_E}{Y(R_E + R_D)} \\ I_C/V_{CB} = H_{22} = \dfrac{R_E + R_C + R_D}{YR_C(R_E + R_D)} \end{cases}$$

If the collector is reverse biased, and the current in the diode is negligible compared with that in R_L, then from equation (7)

$$Y = V_{CB}/V_{C'B} = 1 + R_L/R_F \simeq R_L/R_F$$

(typically $80\text{k}\Omega/200\Omega$)

where R_L and R_F now refer to components in the collector junction amplifier.

If the emitter is forward-biased, and the current in R_L in this amplifier is negligible, then from equation (4)

$$V_{E'B} = I_ER_D\,\alpha_{oN} = R_FI_S\,(\exp pV_{EB}/kT - 1)$$
$$\simeq R_FI_S\,(\exp qV_{EB}/kT)$$

Thus

$$XV_{E'B} = V_{EB} = (kT/q)\ln(V_{E'B}/R_FI_S)$$

To obtain the small-signal h-parameters, the non-linear function X must be replaced in equation (10) by the incremental slope x given by

$$x = \frac{dV_{EB}}{dV_{E'B}} = \frac{kT}{qV_{E'B}} = \frac{kT}{qI_ER_D\,\alpha_{oN}}$$

The current gain $\alpha_{oN} = -h_{21} = R_E/(R_E + R_D)$ is independent of I_E, and it is convenient to express h_{11} and h_{12} in terms of this ratio:

$$h_{11} = xR_D\alpha_{oN} = kT/qI_E$$
$$h_{12} = x\alpha_{oN}/Y = kT/qI_ER_D \;.\; R_F/R_L$$
$$h_{22} = \frac{R_E + R_C + R_D}{R_C(R_E + R_D)} \;.\; R_F/R_L$$

h_{11} $(= r_e)$ obeys the well-known formula correctly. As stated previously h_{22} is independent of I_E and V_{CB} in the analogue, whereas in real transistors it is given approximately by an expression of the form $h_{22} = (aI_E/\sqrt{V_{CB}}) + b$, where the first term is the Early effect and the second, which represents leakage effects, is significant only at low current levels. The feedback factor h_{12} in the analogue varies with I_E and is independent of V_{CB}, whereas the Early feedback factor μ is inversely proportional to $\sqrt{V_{CB}}$, and is independent of I_E. Thus the variations of both factors with operating conditions are incorrect. However, for typical conditions the magnitudes may be made roughly correct, whereas the theoretical large-scale equivalent circuit is seriously wrong because base-width modulation is disregarded. The effect of R_L is therefore beneficial and its magnitude important.

For example, suppose $R_F = 200\Omega$, $R_D = 500\Omega$

If $\alpha_{oN} = 0{\cdot}95$, then $R_E = 500 \times 0{\cdot}95/0{\cdot}05 = 9500\Omega$ and if $\alpha_{oI} = 0{\cdot}87$, then $R_C = 500 \times 0{\cdot}87/0{\cdot}13 = 3350\Omega$.

Therefore:

$$h_{22}R_L = \frac{13350 \times 200}{3350 \times 10000} = 0{\cdot}08$$

If h_{22} is to be 10^{-6}mho, say, a typical value for operation at 1mA, R_L must be $80\text{k}\Omega$, as in the circuit given earlier. At 1mA, $kT/qI_E \simeq 25\Omega$, and therefore the voltage feedback ratio becomes $h_{12} = \dfrac{25 \times 200}{500 \times 80 \times 10^3} = 1{\cdot}25 \times 10^{-4}$.

This value is also quite typical. As h_{22} and h_{12} are not set independently it is clearly desirable that whichever is set by the choice of R_L, the other should automatically assume a value which is reasonable according to the theory of the Early effect.

For the analogue

$$h_{22}/h_{12} = \frac{qI_ER_D}{kT}\left((1/R_C) + \frac{1}{R_E + R_D}\right)$$
$$= \left(\frac{1 - \alpha_{oI}}{\alpha_{oI}} + 1 - \alpha_{oN}\right)qI_E/kT$$
$$= ((1/\alpha_{oI}) - \alpha_{oN})\,\frac{qI_E}{kT} \qquad \ldots\ldots\ldots\ldots\ldots\ldots (11)$$

For a transistor with uniform base doping, a one-dimensional analysis[5] of the Early effect gives

$$h_{22} = 2\mu(1-\alpha_{oN})\,qI_E/kT \text{ where } \mu \equiv h_{12} \quad \ldots\ldots (12)$$

For this simple case a comparison of equations (11) and (12) shows that the value of h_{22}/h_{12} due to R_L in the analogue differs from that due to base-width modulation only insofar as $1/\alpha_{oI}$ differs from $(2-\alpha_{oN})$. As α_{oI} and α_{oN} are near to unity for an alloyed-junction (uniform base) transistor this difference is small. For graded-base transistors the factor 2 in equation (12) is considerably increased, but at the same time α_{oI} becomes appreciably less than unity and so, from equation (11), the ratio h_{22}/h_{12} also increases for the analogue. In general, the ratio is correct within a small numerical factor. The inverse gain α_{oI}, determined by R_C, is not in itself important when the transistor is operated only as an amplifier in the normal direction and so R_C can be adjusted to vary h_{22} independently of h_{12} if the analogue is not to be driven into saturation or inverse operation.

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The Harwell 2000 Series Linear Ratemeter/Integrator

By J. W. H. Wilkins* and D. J. Marsh*

Because of improved accuracy, digital techniques have superseded analogue ones, over the last few years, in most rate and count measurements. This article describes circuit improvements to the original analogue system resulting in a simpler dual-purpose precision instrument, using low-voltage supplies.

(Voir page 493 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 500)

RATEMETER circuit designs so far have not succeeded in providing a linearity between d.c. output and pulse-rate input of better than a few per cent over the range 1 to 10^4 p/s, and at the same time achieving acceptable statistical accuracy below 5p/s. This has resulted in scalers being used in preference to ratemeters whenever these features have been desirable. In these instances where rate information, in an analogue form, is essential, little or no provision is made for very low ($\sim$ few p/m) rate measurement and when a wide range of input rates is to be experienced, logarithmic ratemeters are used to avoid range changing, despite the practical difficulty of achieving an accuracy better than ± 10 per cent.

Both low and high rate measurements are possible by integration techniques, i.e. accumulating charge proportional to the number of pulses received, but again non-linearity due to limited charge storage periods and dependence on rate of input pulses has precluded this useful technique.

The need for ratemeter circuits providing better linearity was largely unjustified with the practice of metering the d.c. output, the combined calibration and reading error of most inexpensive meters of the type used being not better than ± 3 per cent. However, the development of accurate data loggers and analogue to digital convertors has made it necessary to improve ratemeter linearity to about 1 per cent over the widest possible range. A solution to this problem and that of providing better integration has therefore been sought and the successful outcome now gives the prospect of a far wider use of a ratemeter/integrator.

The circuit of the instrument, which embodies the improved facilities, has been kept as simple as possible and operates from low-voltage supplies. This allows this dual function therefore to be incorporated into both portable and transportable equipment, thereby bridging the gap which has often previously existed between field and laboratory instruments, mainly in respect of accuracy and low count rate measurement.

Basic Circuit

Giannelli[1] and Vincent[2] have exploited ratemeter circuits, which differ from the now classic diode-pump circuit of Cooke-Yarborough[3], in their attempt to produce precision ratemeters. Only the last circuit (Fig. 1) lends itself readily to the operation of integration and therefore this circuit was re-investigated, with solid state diodes, to resolve some of its basic limitations. These were the deviation from linearity at high count-rates and errors arising, from d.c. amplifier drift and reverse leakage of semiconductor diodes, when associated with integration.

The diode-pump circuit depends for its linear operation on a constant charge per pulse being supplied, via a feed capacitor (C_F) and forward biased diode MR_1, into a tank capacitor (C_T), the complete discharge of C_F before the arrival of the next pulse and no leakage of the charge on C_T, connected for integration, between pulses.

On the application of a positive going front-edge of a pulse to the pump, assuming a virtual earth d.c. amplifier, the series diode MR_1 is forward biased and conducts to

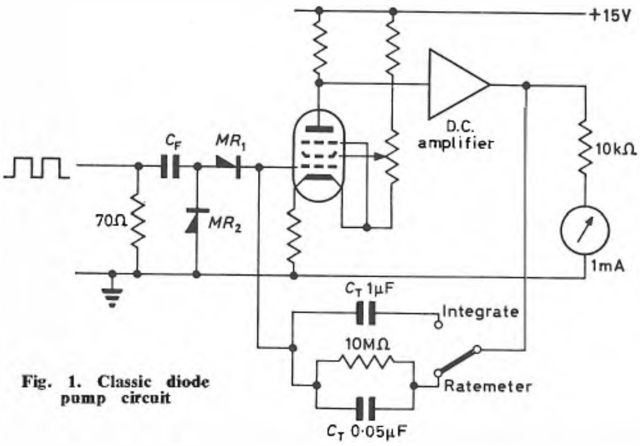

Fig. 1. Classic diode pump circuit

charge C_T while the shunt diode MR_2 is reversed biased and cut off. The negative going back-edge of the pulse puts MR_1 into a reversed bias condition and MR_2 in a forward biased condition to discharge C_F and this condition exists until the next pulse arrives. Ideally the switch over from conducting to non-conducting should be instantaneous and furthermore the impedance of the diodes in these conditions should be zero and infinite respectively. This is not so in practice; a forward biased semiconductor diode does not return to an infinite impedance state immediately a reverse bias is applied. A flow of reverse current persists for some time and is caused by some of the excess minority carriers, present during conduction, being caught on the wrong side of the diode junction at the time of change of applied voltage. These carriers must either recombine or be extracted as a reverse current and in practice both processes take place during the turn-off period.

The amount of charge recovered depends on the impedance of the circuit associated with the diode. Fig. 2(a) shows the order of recovered charge in a low impedance circuit and this is very much greater than that shown in Fig. 2(b) for a high impedance circuit, since recombination of excess carriers takes place to a greater extent in this latter case. Thus for an infinite impedance circuit no reverse charge will be extracted and only recombination will be possible. The time for completion of recombination can be as large as 7μsec and a forward voltage applied during this period, as from a succeeding pulse, finds the diode in a slightly forward biased condition and

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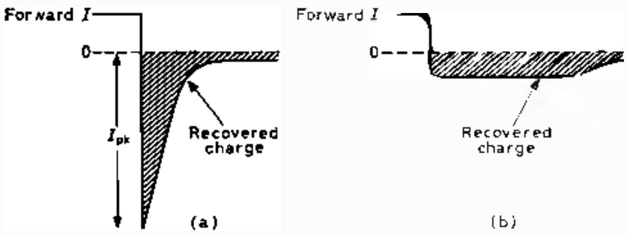


Fig. 2. Recovered charge from high and low impedance circuits
(a) Low impedance circuit
(b) High impedance circuit

of course results in an improved speed of turn-on.

Any charge recovered, however, in the case of MR_1 reduces the amount of charge which should have been put into C_T, and introduces non-linearity both in the indicated count rate or integrated count.

In both instances, the error is reduced by increasing the reverse bias condition impedance of the diode circuit, i.e. by introducing a low leakage diode.

Unfortunately diodes which have a fast turn-on time do not show a high reverse bias impedance but have small carrier storage at turn-off and on the other hand diodes having a reverse bias impedance have a slow turn-on and a high carrier storage at turn-off. In other words, a single diode in the series arm will not meet the ideal conditions.

Requirement on Charge Storage for Count Integration

It can be assumed that the value desired for the reverse current, set by the requirement for negligible error on integration, will be more than sufficient to reduce the error in rate indication to a negligible amount also. Further that the reverse grid current of the d.c. amplifier will normally be less than this value.

Thus considering integration only, two different errors arise:

(a) A random statistical error equivalent to $\pm\sqrt{p}$, where p is the total number of pulses accumulated.

(b) A systematic error due to loss of charge by leakage.

The random error (a) decreases as the integrating time T increases, but error (b) increases rapidly with increasing T and beyond a certain point becomes dominant. The problem is to make both errors small simultaneously.

Assume the integrator circuit consists effectively of a simple CR parallel combination on which a fixed charge q is placed every time a pulse arrives. For satisfactory operation as an integrator (as opposed to a ratemeter), one must have $T \ll CR$. More accurately there will be a fractional loss of charge of amount $\beta \simeq T/2CR$.

This can be proved as follows:

If p pulses, each having charge q, are accumulated, then a voltage increment $\Delta V \simeq pq/C$ occurs, which is approximately a linear increase during the period T. Associated with this is a similar linear increase of leakage current in R from zero to a final value pq/CR. The mean current during this period is $\simeq pq/2CR$ and the total loss of charge by leakage $pqT/2CR$.

$$\therefore \beta = T/2CR \quad \ldots (1)$$

Now the integrating time T is determined by the tolerable statistical error (α, expressed as a fraction) in accordance with:

$$\alpha = 1/\sqrt{p} = 1/\sqrt{nT} \quad \ldots\ldots\ldots\ldots (2)$$

where n = pulse rate, p = total number of pulses.

Eliminating T between equations (1) and (2), the required time-constant CR is obtained:

$$1/\alpha^2 n = T = 2\beta CR$$

$$\therefore CR = 1/2\alpha^2\beta n \quad \ldots\ldots\ldots\ldots (3)$$

$$\text{and } T = 1/\alpha^2 n \quad \ldots\ldots\ldots\ldots (4)$$

In practice, α and β will often be equal and certainly comparable, so that the required value of CR varies inversely as the cube of the permissible total error. A numerical example where n = 5p/m (a typical α-counter background), α = 1 per cent, β = 2 per cent gives:

$$CR = \frac{1}{2 \times 10^{-4} \times 2 \times 10^{-2} \times 5} \text{ min}$$

$$= 50\,000\text{min}$$

$$= 3 \times 10^6\text{sec}$$

Hence, if $C = 1\mu$F

$$R = 3 \times 10^{12}\Omega$$

The corresponding integrating time T is

$$1/\alpha^2 n = 2\,000\text{min} = 33{\cdot}3\text{h}.$$

Very low leakage silicon diodes (1S144) have a reverse impedance of $2 \times 10^{11}\Omega$ at the maximum p.i.v. of 400V. Under circuit conditions, however, the maximum p.i.v. is only momentarily 10V, before feedback reduces it to $\simeq$ 10mV. The reverse impedance of diodes with this low order of applied voltage is not generally known. Several low leakage diodes were measured at larger p.i.v's and the ones with lowest reverse current put into a typical integration circuit and the charge storage time measured. With a p.i.v. of 100V, the following general pattern emerged:

Reverse Current (nA)	1000	100	10
Storage time T (h)	< 1	< 10	< 20

The leakage current, as in all semiconductor diodes will increase with temperature, so that it was estimated that a suitable diode must have less than 5nA leakage current at a p.i.v. of 100V, at room temperature.

Requirements on Switching Speed

Diodes most suitable for linear operation must return to their high impedance state as rapidly as possible after reverse bias has been applied in order to reduce the flow of recovered charge and equally so the shunt diode must switch on as fast as possible to ensure that C_F is fully discharged before the next pulse arrives. This topic is not readily amenable to calculation so that a simple experimental rig was used to measure the extent of non-linearity introduced by a variety of diodes.

A double pulse generator was connected to the circuit of Fig. 1 via a constant amplitude pulse former and a low impedance (70Ω) output circuit. C_F was chosen as 22pF so that it imposed no limitations on discharge time and would reveal the effect of diode turn-on time. Single pulses were supplied at a p.r.f. of 5kc/s and the indicated count-rate set to 50 per cent f.s.d. on an accurate d.c. meter. The pulse generator was then switched to give double pulses, the pulse separation being 1μsec and the rise and fall time of the pulse to the pump being 100nsec

TABLE 1
Count-rate Linearity Tests

	SWITCHING DIODES					LOW LEAKAGE DIODES		
Diode Type	CV4073	ZS40	ZS42	HD5004	BAY31	ZS8	OA202	IS144
Recovery Time (nsec)	—	15	40	0·5	2·5			
Reverse Current (μA)	0·1	0·5	0·5	1·0	0·01	0·005	0·1	0·002
P.I.V. (V)	60	25	100	5	10	30	150	400
Percentage error of f.s.d.	−15	−11	−12	−1·0	−2·5	−22	−56	−85

The increase in the indicated count-rate and the results are shown in Table 1. For a perfect diode, the rate reading should have gone to 100 per cent f.s.d. These results indicate that the least error is introduced by the HD5004 of the switching diodes and the ZS8 of the low leakage diodes, the IS144 has the lowest reverse bias leakage current.

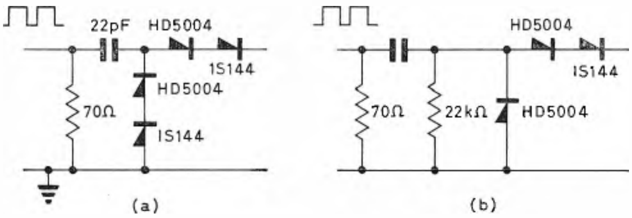


Fig. 3. Double diode pump circuit

(a) Double diode pump circuit initially used
(b) Double diode pump circuit finally used

Double Diode Pump Circuit

From what has gone before, the ideal diode should have a fast switch over, a low charge carrier storage effect and a very high reverse bias condition impedance, although for the shunt diode speed and reverse bias impedance are the two most important characteristics. No single diode meets this requirement but a series combination of a fast diode and a low leakage diode in each arm can go some way towards it. With the circuit arrangement shown in Fig. 3(a), the flow of recovered charge in the low-leakage diode is shut off by the fast return to a high impedance state of the fast diode and during the reverse bias condition the leakage current is limited to that of the low leakage diode. Even this does not produce the ideal characteristics because some proportion of the high recovered charge (55 000pC) of the low leakage diode will still flow during the finite, though short, turn-off time of the fast diode, and this will add to the low recovered charge (30pC) from the fast diode. This effect was examined using the same test as in the previous section and with HD5004 and IS144 diodes in series in each arm. The measured error was now $-1{\cdot}5$ per cent and compared favourably with that obtained using the fastest switching diode alone. On a storage test, however, the value of T obtained, for $\beta = 2$ per cent, was 36 hours which showed a considerable improvement over earlier pump circuits.

Final Diode-Pump Circuit

The differential rate error, using the circuit of Fig. 3(a) was -1 per cent at full scale deflexion and $+2$ per cent at quarter f.s.d., when set-up correctly at half f.s.d. using regular pulses at a p.r.f. of 10kc/s. Using regular pulses, the accuracy of integration was good, as the value of T obtained would suggest, but it was still dependent on the rate at which pulses were supplied. This suggested that C_F was still not being completely discharged between pulses when these had a short separation time. An investigation showed that the forward impedance of the shunt arm diode circuit increased considerably as C_F became more nearly discharged so that the discharge characteristic was not a simple exponential since the resistive component was a function of applied forward bias. The time to fully discharge C_F could be hundreds of microseconds.

A modification[4] to the double diode circuit, which was to put a resistor (R) in parallel with the shunt diodes ensures that the effective impedance of the shunt arm is never greater than R and therefore puts a limit to the time needed to discharge C_F. The value of R must be chosen high enough so as not to shunt appreciably the series arm when C_T is being charged but yet not so high that C_F is not fully discharged between successive pulses separated by the shortest interval to be experienced. With $C_F = 22$pF and $R = 22$kΩ the complete time to discharge will be not greater than 2μsec which is short enough for mean count rates of 10^4p/s since the probability of two pulses occurring within 2μsec is only 1 per cent.

The addition of this resistor nullifies the high reverse bias impedance of the shunt diode MR_3 so that MR_3 can be omitted from the circuit (Fig. 3(b)).

The Ratemeter/Integrator 95/2134-1/6

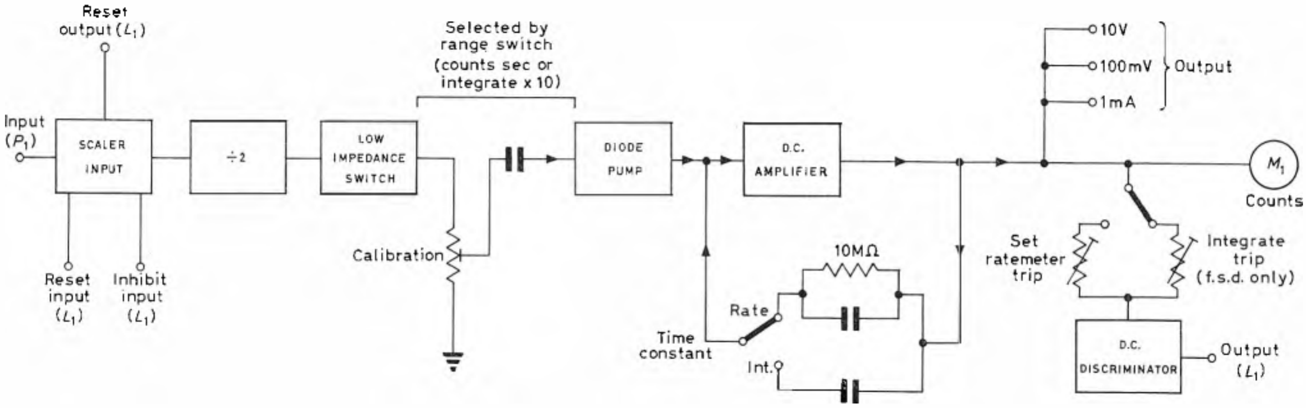
The final diode-pump circuit[5] was incorporated in a prototype model. The diodes in the pump circuit and the d.c. amplifier were contained in a sealed copper box to the tie-bars of the Harwell 2000 Series[6] single-width unit. The box had an easily removed lid. A block diagram of the instrument (now the 95/2134-1/6 is given in Fig. 4.

The following are some of the facilities provided:

Ratemeter

Ranges:	0·1, 2·5, 10 . . . . 10^4c/s
Input	Standard Type 1 pulse*, 700nsec
Resolution:	Limited by input pulse, otherwise 150nsec
Time-constants:	0·2, 1, 5, 25, 125sec
Indication:	2½in (240°) circ-scale meter
Analogue outputs:	1mA, 100mV, 10V.
High alarm:	Adjustable level. Standard Type 1 Binary level (BCL)* output when operated
Self test:	Push-button, nominal 1kc/s
Local Reset:	Push-button
Count on-off	Switch

Fig. 4. Arrangement of linear ratemeter/integrator



Integrator

Ranges: 0-10, 25, 100 . . . . 10^5 counts

Input, resolution, indication, analogue output, self-test, local reset, count on/off as for ratemeter.

High alarm: Adjustable but preset at f.s.d. Standard Type 1 B.C.L. output when operated.

Reset input, Inhibit input, Reset output: Standard Type 1 B.C.L.

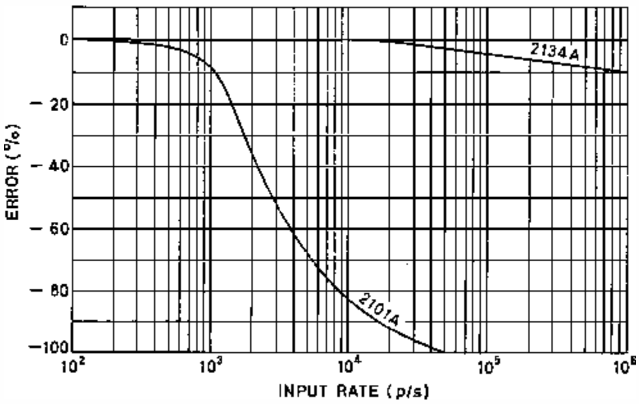


Fig. 5. Input rate dependence integrating 1 000 counts

**Standard Type 1 Pulse*

Polarity: Negative

Amplitude: Excedeing 5V for at least 200nsec but not exceeding 10V.

Duration: Between 400nsec and 700nsec measured at an amplpitude of 0·75V.

Impedance: Delivered from an impedance of not more than 330Ω.

**Standard Type 1 B.C.L.*

'No signal' level: ±0·75V

Duration: −5Vmin to −10V

'Signal level' Minimum of 200nsec at −5V

Change-over time Not greater than 250nsec between no signal and −5V, or vice versa.

The standard deviation of the output for random inputs, using the longest time-constant, is ±2 per cent at 10p/s and ±9 per cent at 0·5p/s. The integration mode will normally therefore, be used on input rates less than 10p/s. To simplify the unit, only the three lowest ranges 10, 25 and 100 counts are accurately set up for the integration and the other ranges are set up accurately for rate measurement. This results in possible errors of ±1 per cent at f.s.d. when the three lowest ranges are used for rate measurement and the same error when the highest ranges are used for integration.

A d.c. discriminator provides a fast leading edge to the level generated when the high alarm is operated. The changeover time between 'signal absent' and 'signal present' is 250nsec and the signal duration at −5V is 500μsec. This signal is suitable for resetting the integrator, the 500μsec reset time being negligible at low input rates, and feeding to 2000 Series scalers, etc.

Measurements on the Prototype Model

All measurements were made using an external B.S.I. 1mA meter having 10μA sub-divisions over a scale length of 6in.

Zero Drift

This was checked after 4 weeks from an initial zero-setting, during which time the unit was switched on and off many times for undetermined periods and was subjected to temperature cycling experiments (0 to 45°C). The final drift was +10mV corresponding to 0·1 per cent of f.s.d. Over a subsequent 3 week check no further drift occurred.

Ratemeter Resolution

Using the test outlined earlier, this time with the double pulse separation at 150nsec, the error was less than 2 per cent.

Ratemeter Linearity

The ratemeter ranges above 10p/s were set up accurately at full scale using a crystal controlled pulse generator. Pulses from the same generator were supplied at 10, 25, 50, 100, 250, 500, 1 000, 2 500, 5 000 and 10 000p/s, and no measurable errors were obtained at 0·25 and 0·5 f.s.d. on the 100, 1 000 and 10 000p/s ranges.

Ratemeter Temperature Tests

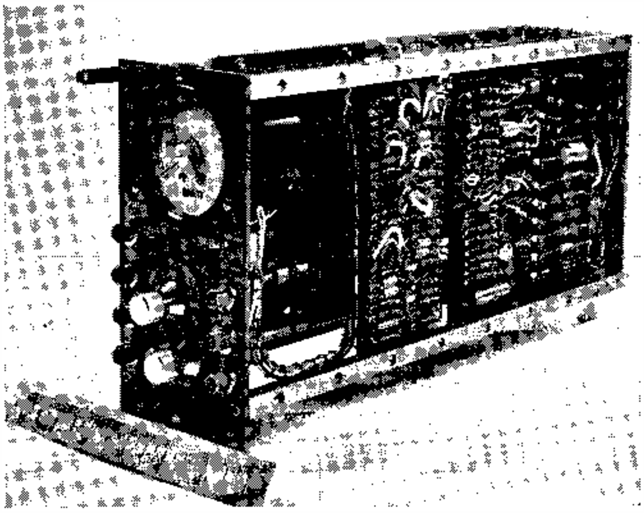
At an input pulse rate of 1 000p/s, time-constant 5sec (percentage s.d. = ±1 per cent), 0·3 per cent reduction in indicated rate occurred when the temperature was raised from 20 to 30°C and a further reduction of 0·7 per cent from 30 to 45°C.

Integrator Linearity

The three lowest ranges were set-up using a scaler to read 10, 25 and 100 counts at full scale respectively at a pulse p.r.f. of 1p/s. No deviation from linearity could be detected on any of the three ranges.

Integrator Dependence on Input Pulse Rate

1 000 regular pulses were integrated at spot pulse recurrence frequencies in the range 10^2 to 10^6p/s. Fig. 5 shows the percentage error observed for the present integrator as compared with a valve type integrator. Thus at 10^4p/s the error was less than −1 per cent and even at 10^6 was only −9 per cent.



The ratemeter/integrator

Integrator Charge Storage Time

With the output at full scale deflexion, the times taken for the output to fall by 2 per cent were 20 hours at 20°C, 6 hours at 35°C and 1 hour at 45°C.

Acknowledgements

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A Pulse Height Measuring Unit

By C. I. Sach*, B.E.E.

A unit which measures the peak value of a single negative polarity pulse is described. It was originally developed for the case of a pulse having a sharp negative front edge of magnitude between 1V and 5V followed by an exponential decay back to zero with a 2sec time-constant. The value of the peak signal is displayed with an error of less than 5 per cent on a d.c. voltmeter providing the reading is taken within 10sec of the event. The characteristics of the unit make it suitable for reading the peak value of single rectangular pulses with durations as short as 3msec.

(Voir page 493 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 500)

As part of a laser investigation programme, a device was developed for measuring the energy output of a single laser discharge. Basically, this consists of a black body absorber whose temperature is measured by an attached thermocouple. On the application of a light flash, the thermocouple generates a voltage step of 1mV which then decays to zero in approximately 2sec. With suitable calibration, the amplitude of the pulse front can be used as a measure of the absorbed energy. A method for recording the peak value of the pulse was required which would not use versatile high speed recorders, so this unit was designed to permit the peak value to be displayed on a normal d.c. voltmeter. The operation of the device is based on a short term storage unit which charges up quickly to the peak value and then holds it long enough for an operator to take a reading. As direct storage and read-out at the millivolt level presents difficulties, an amplifier with a range of stabilized gains around 5 000 is used to bring the pulse level up to about 5V. Owing to excessive drift, d.c. amplification, which would normally be used to preserve the sag characteristic of the pulse, is not possible without some complication. Hence, a.c. amplification with a low frequency response such as not to degrade the pulse shape appreciably is used. The details of the amplifier will not be dealt with in this article; attention being confined to the peak reading facility.

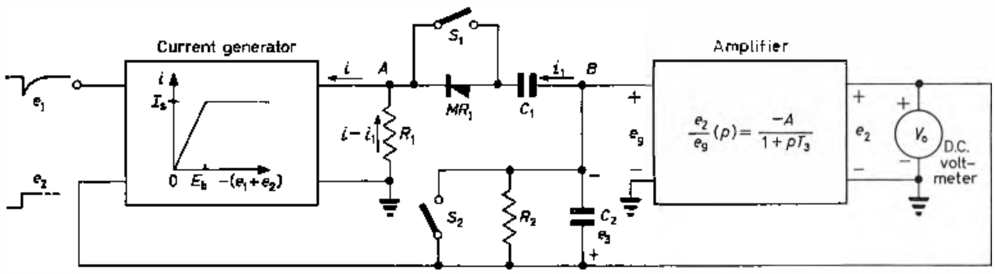


Fig. 1. Arrangement of the device

Peak Reading Unit

A schematic of the device is shown in Fig. 1. Here a current generator stage under the control of the input e_1 and the output e_2 of the amplifier produces a current i. A component of this, i_1, is passed by the diode MR_1 and capacitance C_1 and charges up the integrating capacitance C_2 connected around the output amplifier. To a first approximation C_2 will charge up until the output signal e_2 is positive and equal in magnitude to the peak negative value of e_1. When this condition occurs, the charging current is reduced to zero. As the value of e_1 decays, the diode isolates the current stage from the integrating amplifier which thus holds its output signal (subject to leakage effects) and allows the output to be read on a voltmeter.

Prior to receiving the input pulse, e_1 is at 0 volts so e_2 is also adjusted to this value. This is done by closing the normally open switches S_1 and S_2. This allows C_2 to discharge to zero and C_1 to take up the voltage existing between the amplifier input B and the output A of the current generator. A 'set zero' control is then used to bring e_3 (and of course e_g) to zero. When the switches are released the output remains at 0 volts and the diode is returned to the circuit. As it has no voltage across it, there is no leakage current tending to pass through C_2 and therefore the output level remains steady. The circuit is thus primed to receive a negative going input pulse.

The current control stage has a saturating characteristic such that when the difference between e_2 and $-e_1$ exceeds a value E_b there is a constant charging current I_S. For values below E_b the current stage is linear as indicated in Fig. 1. The resistance R_1 is included to represent a biasing resistance which must be used in any practical circuit.

The amplifier uses an electrometer input valve so that the current drawn from C_2 during the storage phase is minimized. In addition, the gain between e_g and e_2 is made sufficiently large that the input B can be regarded as a virtual earth point at all times, i.e. the output voltage is equal to the voltage across C_2.

The operation of the circuit will be analysed in detail so that expressions for charging and decay time, accuracy etc can be formulated in terms of the circuit parameters. These will then form the design equations for the particular application described.

Circuit Analysis

The amplifier equations, neglecting its time-constant T_3, are

$$e_2 = -Ae_g \quad (1)$$
$$e_g = e_2 - e_3 \quad (2)$$

Here e_3 is the voltage across the capacitance C_2.

Eliminating e_g between these equations gives

$$e_2/e_3 = A/(1+A) \quad (3)$$

Thus if A is made sufficiently large, one may take $e_2 = e_3$ which implies $e_g = 0$. This supports the use of

* *Australian Defence Scientific Service.*

point B as a virtual earth. The method used for calculating A and T_3 is given in the Appendix.

The general equation for the current generator is

$$i = f(e_1+e_2) \quad \dots\dots\dots\dots\dots\dots \quad (4)$$

where the function is defined graphically in Fig. 1.

Since point B is at earth potential two necessarily equal expressions are obtained for the voltage at point A if the forward voltage drop across MR_1 is neglected. These give

$$-R_1(i-i_1) = -(1/C_1 \int i_1 dt$$

On differentiation this gives

$$i_1 = R_1C_1[(di/dt)-(di_1/dt)] \quad \dots\dots\dots\dots \quad (5)$$

Also, since

$$e_2 = e_3$$

$$e_2 = (1/C_2) \int i_1 dt$$

$$\text{i.e.} \quad i_1 = C_2(de_2/dt) \quad \dots\dots\dots\dots\dots \quad (6)$$

Equations (6) and (7) can be used to eliminate i_1 to give

$$(d^2e_2/dt^2) + (1/R_1C_1)(de_2/dt) = (1/C_2)(di/dt) \quad .. \quad (7)$$

Equations (4) and (7) then become the defining ones for the system. To obtain solutions consider the response to an input signal $e_1 = -E_1H(t)$ where $E_1 > E_b$. This means that both the saturation and linear region of the current stage will be used. In the saturation region which lasts from $t = 0$ to $t = t_1$ (which is yet to be established),

$$i = I_SH(t) \quad \dots\dots\dots\dots\dots\dots\dots \quad (8)$$

Integrating equation (7)

$$(de_2/dt) + (e_2/R_1C_1) = (i/C_2) + K_1 \quad \dots\dots \quad (9)$$

where K_1 is constant. At $t = 0$, $e_2 = 0$, $i = I_S$.

In addition it is necessary to know the value of de_2/dt at this point.

From equation (6), it follows that

$$de_2/dt\,|_{t=0} = i_1(0)/C_2 \quad \dots\dots\dots\dots \quad (10)$$

But since at the step front C_1 presents a much lower impedance to the current source than R_1 then all the current I_S will initially flow through it.

$$\therefore\ i_1(0) = I_S \quad \dots\dots\dots\dots\dots\dots \quad (11)$$

Substituting this in equation (10), the necessary initial condition is obtained

$$de_2/dt\,|_{t=0} = I_S/C_2 \quad \dots\dots\dots\dots \quad (12)$$

Substituting the initial conditions in equation (9) it is found that $K_1 = 0$. Therefore the system equation becomes

$$(de_2/dt) + (e_2/R_1C_1) = I_S/C_2 \quad \dots\dots\dots \quad (13)$$

Solving this together with the initial condition $e_2 = 0$, gives

$$e_2 = (I_SR_1C_1/C_2)[1 - \exp(-t/R_1C_1)] \quad \dots\dots \quad (14)$$

Thus the first response of the circuit consists of an exponential rise toward a value $I_SR_1C_1/C_2$ with a time-constant R_1C_1. This response phase is terminated when e_2 reaches $E_1 - E_b$ where the current stage becomes linear. The time t_1 at which the transition occurs is obtained by putting $e_2 = E_1 - E_b$ in equation (14) and solving for t. This gives

$$t_1 = R_1C_1 \ln \frac{1}{1 - \dfrac{(E_1 - E_b)C_2}{I_SR_1C_1}} \quad \dots\dots\dots \quad (15)$$

This condition for minimizing t_1 is

$$I_SR_1C_1/C_2 \gg E_1 - E_b \quad \dots\dots\dots\dots \quad (16)$$

If equation (16) is satisfied then

$$t_1 \simeq \frac{(E_1 - E_b)C_2}{I_S} \quad \dots\dots\dots\dots\dots\dots \quad (17)$$

The value of de_2/dt at $t = t_1$ is denoted by α and is found to be

$$\alpha = I_S/C_2\left[1 - \frac{(E_1 - E_b)C_2}{I_SR_1C_1}\right] \quad \dots\dots\dots\dots \quad (18)$$

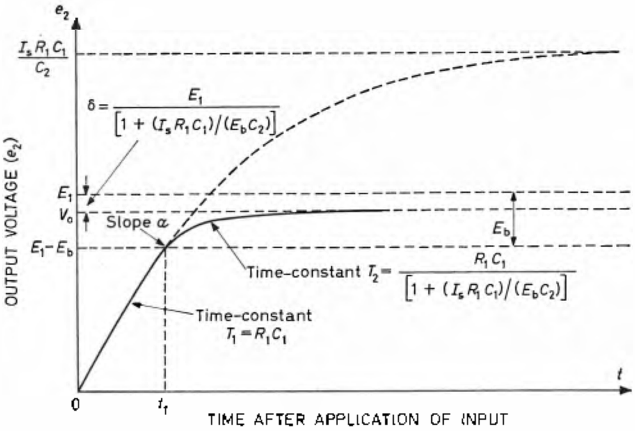


Fig. 2. Idealized output waveform

$$t_1 = R_1C_1 \ln \frac{1}{1 - \dfrac{(E_1 - E_b)C_2}{I R_1 C_1}} \simeq \frac{(E_1 - E_b)C_2}{I_s}$$

This expression is required for solving the system response for $t > t_1$. In this region the current stage is in its proportional or linear region defined by

$$i = -(I_S/E_b)[e_1 + e_2] \quad \dots\dots\dots\dots \quad (19)$$

If the input pulse time-constant is much greater than the charging time, e_1 may be considered as being constant and equal to $-E_1$.

$$\therefore\ i = (I_S/E_b)[E_1 - e_2] \quad \dots\dots\dots\dots\dots \quad (20)$$

This equation, together with equation (7) must be solved in conjunction with the $t = t_1$ conditions viz.

$$e_2 = E_1 - E_b,\ (de_2/dt) = \alpha \text{ and } i = I_S.$$

Integrating equation (7) gives

$$(de_2/dt) + (e_2/R_1C_1) = (i/C_2) + K_2 \quad \dots\dots \quad (21)$$

On substituting the known values at $t = t_1$, it is found that

$$K_2 = \frac{C_2(E_1 - E_b)}{C_1} + R_1C_2\alpha - R_1I_S \quad \dots\dots \quad (22)$$

Using equations (18), (20) and (22) it may then be shown that equation (21) simplifies to

$$(de_2/dt) + e_2[(1/R_1C_1) + (1/C_2)(I_S/E_b)] = I_SE_1/C_2E_b \quad \dots\dots \quad (23)$$

The final value of e_2 which will be that read by the voltmeter (V_o) is obtained by putting $de_2/dt = 0$ in equation 23.

$$\therefore\ V_o = \frac{E_1}{1 + \dfrac{E_bC_2}{I_SR_1C_1}} \quad \dots\dots\dots\dots\dots\dots \quad (24)$$

This shows that the final value reached is not equal to the input pulse amplitude E_1 as suggested in the preliminary description but in fact falls short of it, giving rise to an error,

$$\delta = \frac{E_1}{1 + \dfrac{I_sR_1C_1}{E_bC_2}} \quad \dots\dots\dots\dots\dots \quad (25)$$

This error can be reduced to any desired extent by arranging that $I_sR_1C_1/E_bC_2$ is made sufficiently large. This condition is then in agreement with equation (17).

The cause of the error can be explained as follows. The output voltage V_o is closely equal to the voltage produced on C_2 by the current i_1. This current also charges up C_1 to to a voltage such that the changes in charge on each capacitance due to the input signal are equal. Thus the potential at point A drops by an amount V_oC_2/C_1 and this requires that a residual current flows through R_1

to maintain this potential. This in turn requires a difference in potential of $-e_1$ and e_2.

Clearly the error can be reduced by making $C_1 \gg C_2$ in which case the potential at A is decreased so that the residual current requirement is reduced.

Alternatively, R_1 can be increased.

The solution of equation (23) is

$$e_2 = V_o - (V_o + E_b - E_1) \exp\left[\frac{-(t-t_1)}{T_2}\right] \quad \ldots\ldots (26)$$

where

$$T_2 = \frac{R_1C_1}{1 + \dfrac{I_sR_1C_1}{E_bC_2}} \quad \ldots\ldots\ldots\ldots\ldots\ldots (27)$$

The results of this analysis are summarized in the ideal response shown in Fig. 2.

Charging Time

It is not possible to obtain a formula independent of E_1 for the time taken for the output to rise within a certain fraction of its final value. This follows because of the two distinct regions of operation. However, it is possible to obtain a limiting value which will not be exceeded under any condition.

Firstly if $E_1 \gg E_b$ the charging time can be approximated by t_1.

$$\therefore T_{c1} = \frac{(E_1 - E_b)C_2}{I_s} \ldots\ldots\ldots\ldots\ldots (28)$$

On the other hand, if an input pulse of amplitude E_b is assumed then the system operates completely within its linear region and the response is

$$e_2 = V_o[1 - \exp(-t/T_2)] \quad \ldots\ldots\ldots (29)$$

This is obtained by substituting $t_1 = 0$ and $E_1 = E_b$ in equation (26).

If the storage time is defined as the time required to rise within 1 per cent of the final value V_o, then in this case

$$T_{c2} = 4{\cdot}6\, T_2 \simeq 5T_2$$

$$\therefore T_{c2} \simeq \frac{5R_1C_1}{1 + \dfrac{I_sR_1C_1}{E_bC_2}} \quad \ldots\ldots\ldots\ldots\ldots\ldots (30)$$

Using equation (25), T_{c2} can be expressed in the useful form

$$T_{c2} = 5R_1C_1(\delta/E_1) \quad \ldots\ldots\ldots\ldots\ldots (31)$$

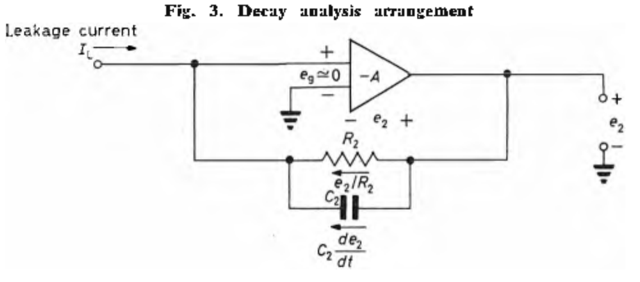
Thus although a simple and exact expression for the charging time is not available, it is clear that it will never significantly exceed $T_{c1} + T_{c2}$ so this will be taken as the design equation for charging time. This approximation errs on the side of safety as it predicts a higher value than would be given by an exact formula. The equation for charging time is therefore taken as

$$T_c = \frac{(E_1 - E_b)C_2}{I_s} + 5(\delta/E_1)R_1C_1 \quad \ldots\ldots\ldots (32)$$

Decay of Peak Signal

The circuit components concerned with the decay of the stored signal are shown in Fig. 3. Here I_l represents the assumed constant leakage current of the reverse biased diode. This current flows despite the series capacitance C_1. The three currents flowing into the virtual earth point B are shown in the diagram and since they must produce a zero resultant, one has the characteristic equation

Fig. 3. Decay analysis arrangement



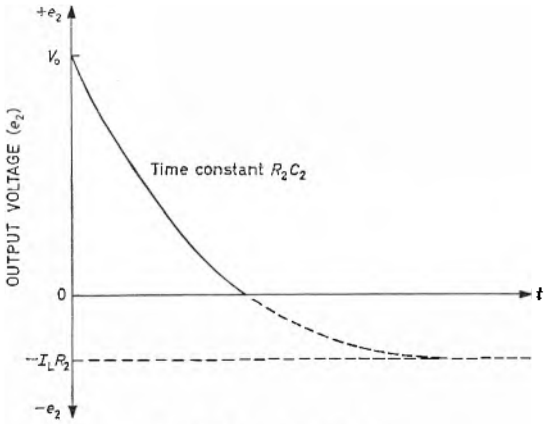


Fig. 4. Decay characteristic

$$I_l + (e_2/R_2) + C_2(de_2/dt) = 0 \quad \ldots\ldots\ldots (33)$$

Solving this equation together with the initial condition $e_2 = V_o$ at $t = 0$ gives

$$e_2 = (V_o + R_2I_l)\exp(-t/R_2C_2) - R_2I_l \quad \ldots\ldots (34)$$

The result is shown in Fig. 4. In practice, of course, e_2 will not fall below zero since this is the value to which e_1 returns. The current generator will then act to clamp the output at this level. The time T_d required for the output signal to decay from 100 per cent to 98 per cent of its peak value can be found from equation (34) to be

$$T_d = 0{\cdot}02R_2C_2\frac{V_o}{V_o + R_2I_l} \quad \ldots\ldots\ldots\ldots\ldots\ldots (35)$$

The value of T_d must be such that the operator has a reasonable time in which to take a reading. Five seconds is considered adequate.

Design Procedure

The previous analysis has shown that the system produces a final output which is in error by an amount δ defined by equation (25). It reaches within 1 per cent of this value in a time T_c given by equation (32). The output thus produced remains within 2 per cent of its peak value for a time T_d given by (35). The amplifier gain A, and time-constant T_3, are given by the equations (43) and (44) in the Appendix.

For the purposes of this design it was required that an input pulse $E_1 = 5$V having a 2sec time-constant should be read with an accuracy of better than 5 per cent for up to 10sec after the pulse.

The following values were nominated

$E_b = 0{\cdot}4$V, $R_2 = 100$MΩ, $I_l = 0{\cdot}01\mu$A and $R_1 = 2{\cdot}7$kΩ.

To obtain the required accuracy a value of $\delta/E_1 = 0{\cdot}005$ was chosen. The decay time T_d was to be 4sec, while T_c was to be 10msec. This allows the input pulse to decay with a time-constant as small as 0·25sec while still recording the peak to better than 5 per cent. The ability to cope with this shorter time-constant allows the specifications on the sag characteristic of the pre-amplifier following the thermocouple to be relaxed.

Using the given values, C_2 is calculated from equation (35) to be 3μF. Equations (32) and (35) are solved simultaneously and give $I_s = 5$mA and $C_1 = 25\mu$F. From equation (43), $A = 200$ and from equation (44), $T_3 = 3$msec. As this value represents a maximum figure

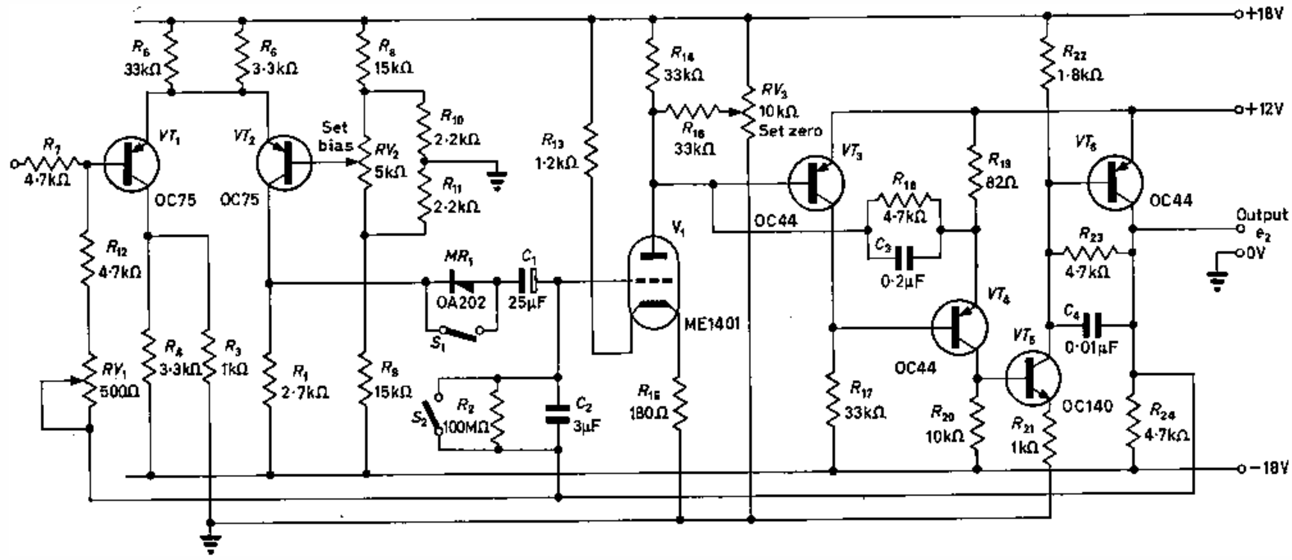


Fig. 5. The complete circuit

and the design is in no ways complicated by a smaller value, then T_3 will be made 1msec. The circuit is shown in Fig. 5. Here transistors VT_1 and VT_2, connected as a long tail pair form the current generator stage. The resistance combination in the emitter circuit supplies 6mA which divides at the emitters so that under normal conditions 5mA passes through VT_2. With e_1 and e_2 at their quiescent values of 0 volts the division of this current is adjusted by the 'set bias' potentiometer RV_2. This is altered until the collector potential of VT_2 is $-4{\cdot}5$V. When the input negative pulse arrives, VT_2 is cut off, thus effectively producing a 5mA current step of the correct polarity. The two signals e_1 and e_2 are added using the resistors R_7 and R_{12}. To help compensate for various errors in the system, a 500Ω trimming resistor RV_1 is used in series with R_{12}.

The charging network consists of R_1, the collector resistance of VT_2, a silicon diode type OA202 for MR_1 and the two capacitors C_1 and C_2. The amplifier consists of an electrometer triode VT_3 and transistors VT_4 to VT_7. The forward gain of the amplifier is given by

$$A = g_m\,(R_{18}/R_{19})\,(R_{20}/R_{21})R_{23} \quad \ldots\ldots\ldots \quad (36)$$

where g_m is the transconductance of the electrometer and is typically about 80µA/V. The parallel combination of R_{18} and C_3 defines the amplifier time-constant of 1msec, while the capacitance C_4 shunting R_{29} is required for amplifier stability. The output stage VT_7 can provide sufficient current for charging C_2 as well as supplying loads R_{24}, R_{12} and the output meter. The output level under quiescent conditions is adjusted to 0 volts using the 'set zero' potentiometer RV_3. During this adjustment S_1 and S_2 are held in the closed position. These switches should be arranged so that on release S_1 'breaks' before S_2.

Capacitor C_2 should have a fairly low leakage current for the voltages to which it will be subjected, i.e. 5V. It is desirable that this be well below 0·01µA so that the diode leakage current and the shunt resistance R_2 become the dominant decay determining components. In the unit constructed a block type was used. The series capacitor C_1 is not so critical and in fact an electrolytic was used in the unit constructed.

Circuit Opertion and Performance

To carry out the initial adjustments, the supplies are switched on and S_1 and S_2 closed. The 'set zero' is then used to bring the output level to 0 volts. The 'set bias' is then used to set the collector potential of VT_2 to 4·5V. The switches are then released. The trimming resistor RV_1 is then set by applying a d.c. voltage of known negative value (desirably -5V) to the input and then adjusting RV_1 until the output reads $+5$V. A brief closure of the switches then restores the circuit to its waiting condition. In practice only the 'set zero' and switches need be available as normal operator controls and even then the zero control will not require much alteration. The other two controls need only occasional adjustment and should not be available for indiscriminate 'twiddling.'

The circuit performed substantially as predicted. To test its charging time capabilities, rectangular pulses of various durations but known fixed amplitude were fed to the input and the corresponding output readings noted. The percentage error as a function of pulse length is shown in Fig. 6, and this indicates that the output reading is still accurate to within 3 per cent for a 3msec input pulse.

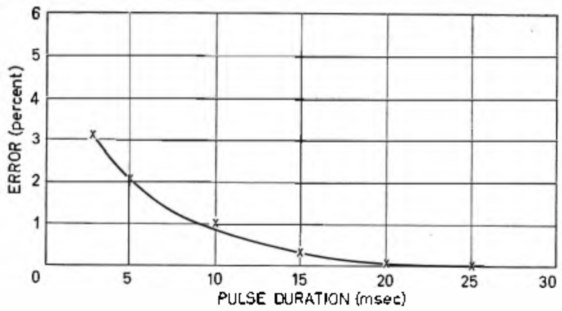
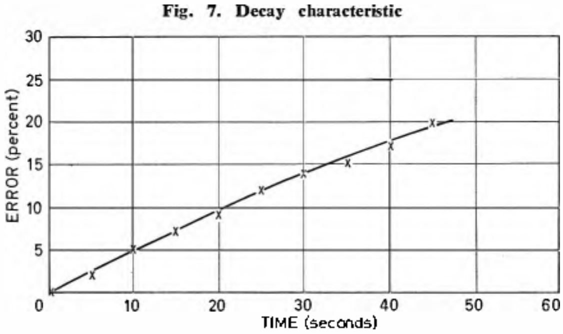


Fig. 6. Charging characteristic

Fig. 7. Decay characteristic



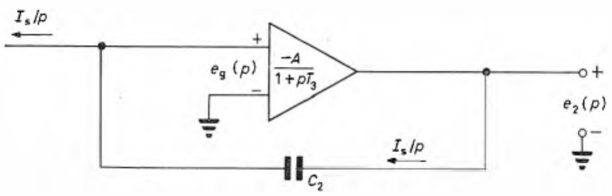


Fig. 8. Amplifier analysis arrangement

The approximate decay characteristic is shown in Fig. 7 where it is seen that the error is 5 per cent, 10sec after the pulse is applied.

For the indicating meter a d.c. voltmeter with critical damping is desirable for ease of reading.

Conclusions

A peak reading unit has been described whose operation agrees well with the simplified analysis presented. It is suitable for the particular application described but has general application for dealing with pulses whose width may be as short as a few milliseconds. It can also be used for reading the peak value of recurrent events, e.g. a pulse train or a sine wave. The principle of the instrument should allow units with smaller charging and longer decay times to be realized.

Acknowledgment

This article is published by permission of the Chief Scientist, Australian Defence Scientific Service, Department of Supply, Melbourne, Victoria, Australia.

Appendix

Evaluation of Amplifier Gain (A) and Time-Constant (T_3)

Realistic estimates of the gain A and time-constant T_3 can be made by considering the output response of the amplifier when the input is subjected to a constant charging current I_s. The circuit for consideration is shown in Fig. 8.

In Laplace transform notation

$$e_2/e_g\,(p) = \frac{-A}{1+pT_3} \quad \ldots\ldots\ldots\ldots (37)$$

Now the input current can be regarded as a step at $t = 0$ whose transform is therefore I_s/p.

From the circuit it is seen that

$$e_2(p) - e_g(p) = 1/pC_2\,.\,(I_s/p) \quad \ldots\ldots\ldots (38)$$

Eliminating $e_g(p)$ between these two equations

$$e_2(p) = \frac{AI_s}{C_2p^2\,(1+A)\,[1+pT_3/(1+A)]} \quad \ldots\ldots\ldots (39)$$

Inverse transforming gives

$$e_2(t) = \frac{A}{1+A}(I_s/C_2)\,t - \frac{AI_sT_3}{C_3\,(1+A)^2}\left[1 - \exp\frac{-t\,(1+A)}{T_3}\right] \quad \ldots\ldots\ldots\ldots (40)$$

For an ideal amplifier the response would be $e_2(t) = I_St/C_2$. Equation (40) shows that the practical system has two components, the first being a constant charging characteristic whose slope is slightly smaller than the ideal case because of the finite gain A. The second is a time dependent term which is zero at $t = 0$ and approaches I_ST_3/C_2A for $t > T_3/(1+A)$. Assuming that $T_0 \gg T_3/(1+A)$ then in the time taken for the ideal system to charge to E_1, a practical one will charge to a value E_2 where

$$E_2 = \frac{A}{1+A}\,E_1 - \frac{I_sT_3}{C_2A} \quad \ldots\ldots\ldots\ldots (41)$$

The fractional error F is defined as

$$F = \frac{E_1 - E_2}{E_1} \quad \ldots\ldots\ldots\ldots\ldots\ldots (42)$$

and gives in this case

$$F = \frac{1}{1+A} + \frac{I_sT_3}{AE_1C_2}$$

The values of A and T_3 are selected so that each component gives an error of 0·5 per cent. In the first case this gives

$$A = 200 \quad \ldots\ldots\ldots\ldots\ldots\ldots (43)$$

while in the second, using equation (42) and (43)

$$T_3 = C_2E_1/I_S \quad \ldots\ldots\ldots\ldots\ldots\ldots (44)$$

A New Air Surveillance Radar

A new high definition general purpose air surveillance radar system has recently been brought into operation at La Villiage Airport, Guernsey in the Channel Islands which deals with an annual traffic movement of some 20 000 and which, at its peak period, has a higher rate of movement than London Airport.

This radar system, the Plessey AR-1 was originally developed by the Ground Radar Division of Decca Radar Ltd which is now part of the Plessey Electronic Group.

The system employs two transmitters operating in frequency diversity, typical frequencies being 2880 ±20Mc/s and 3020 ±20Mc/s, each transmitter having a magnetron peak power of 650kW.

The principal characteristics of the AR-1 system are as follows:

RADIATION

Frequency	Spot frequency in the band 2700-3100Mc/s.
Polarization	Variable from linear through elliptical to circular.

AERIAL

Dimensions	16ft wide × 6·5ft high
Gain	31dB
Radiation pattern ...	Cosec²
Horizontal beamwidth ...	1·5 to half power points
Rotation rate	15 rev/min

TRANSMITTER

Peak power	650kW nominal
Pulse length	1μsec
Pulse recurrence frequency	700p/s

RECEIVER

Noise factor	6dB max.
Intermediate frequency ...	30Mc/s
Overall bandwith	1·2Mc/s
Moving target indicator ...	Uses a three pulse correlation system. No zero response blind speed between 0 and 560 kts.
Diversity operation ...	Two transmitter / receivers operating in frequency diversity. Typical frequencies: 2880 ±20Mc/s and 3020 ±20Mc/s.

A Low-Frequency Pseudo-Random Noise Generator

By C. Kramer*

A circuit is described which generates noise in the sub-audio range by means of a feedback shift register. The noise is nearly random, with a binomial amplitude distribution. The spectrum is flat in the region of interest. The statistical characteristics are highly stable and predictable.

(Voir page 493 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 500)

A NOISE generator in the sub-audio frequency range may sometimes be needed. The most common applications of such an instrument are testing of control systems, vibration testing, and the solution of problems involving noise by means of an analogue computer. It is well known that white noise in this frequency region cannot be generated by simply amplifying the output of a noisy device, as all amplifiers, transistors as well as valves, have a noise spectrum which varies inversely with the frequency ("$1/f$ noise").

Generally, therefore, three different types of low-frequency generator are constructed. In the first type a narrow-band noise signal is heterodyned to the low-frequency range. In a typical realization[1] a 2kc/s noise signal, generated by a thyratron followed by a 2kc/s selective amplifier, is mixed with a 2kc/s sine wave. As the noise amplitude will be directly influenced by the noise properties of the thyratron and by the amplification factors of the amplifiers, stability will always be a problem in noise generators of this kind.

In a second type[2] the noise, generated by a physical source, is converted into a random step function by means of low-frequency sampling. This type of noise generator suffers from the same stability difficulties. These can be partially alleviated by feedback control of the noise amplitude, as the authors suggest, but this affects the noise spectrum in such a way that the lowest frequencies are attenuated.

In the third type[3] the original noise is first clipped and then, by means of sampling, converted into a random binary sequence (also called random telegraph wave). After filtering in a low-pass filter this gives a low-frequency noise signal, in accordance with a theorem by S. O. Rice[4]. Though this type of noise generator is less liable to stability problems than the others, it is still not fully independent of the properties of the noise source.

In the present case it was decided to generate a random binary sequence, not with the aid of a physical noise source, but by a mathematical method. For this new type of noise generator use is made of the properties of a so-called feedback shift register.

The Feedback Shift Register as a Pseudo-Random Sequence Generator

A feedback shift register is a binary circuit, the properties of which were first described by D. A. Huffmann[5]. It consists of a number of flip-flops and 'exclusive OR' gates (Fig. 1). Upon application of a trigger pulse the information contained by the flip-flop chain is shifted one step to the right. At the same time a bit, obtained by a parity check on the content of a number of the flip-flops, is shifted into the first flip-flop. In this way a cyclic code will circulate through the shift register. If the number and positions of the feedback connexions are properly chosen, this code will contain all possible combinations of N bits, except that in which all bits are positive. For instance the circuit shown in Fig. 1 generates the code shown in Fig. 3. These maximal length shift regis-

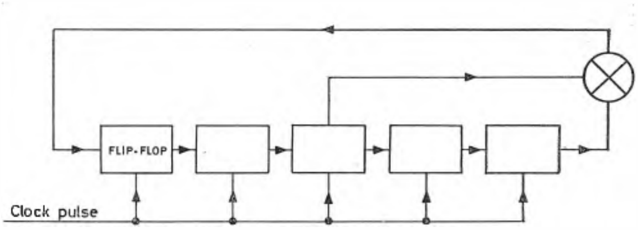


Fig. 1. Feedback shift register

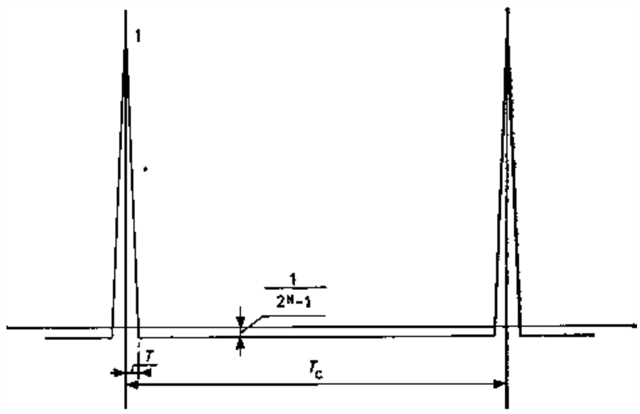


Fig. 2. Autocorrelation function of shift register code

ter codes have a length of $2^N - 1$ bits, where N is the number of flip-flops.

Now these codes have the property that, when such a code is multiplied by a time-shifted replica of itself, the result is again the same code with a different time shift. This property can be used to calculate their autocorrelation function. Because the number of negative bits exceeds the number of positive bits by one, the autocorrelation function $1/T_c \int_0^{T_c} f(t)\, f(t-\tau)dt$ will be $-T/T_c = -1/2^N-1$ for all values of $\tau \neq 0$, T_c, $2T_c$ . . . For $\tau = 0, T_c$ . . . the autocorrelation function is 1. So the autocorrelation function has the form sketched in Fig. 2.

This autocorrelation function is the same as that of a real random binary sequence, except for two differences: this sequence is not purely random but periodic, and the autocorrelation between the peaks is not exactly zero but

* *N.V. Philips' Gloeilampenfabrieken Eindhoven-Netherlands*

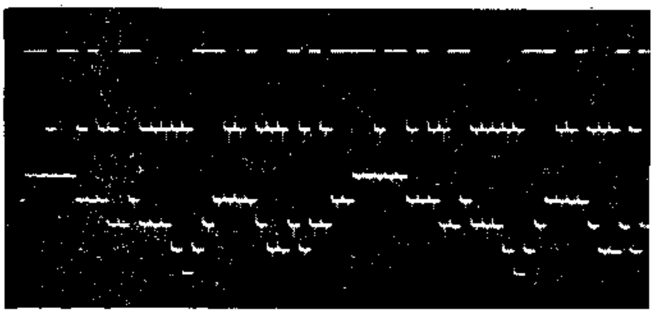


Fig. 3. 31-bit code and step signal

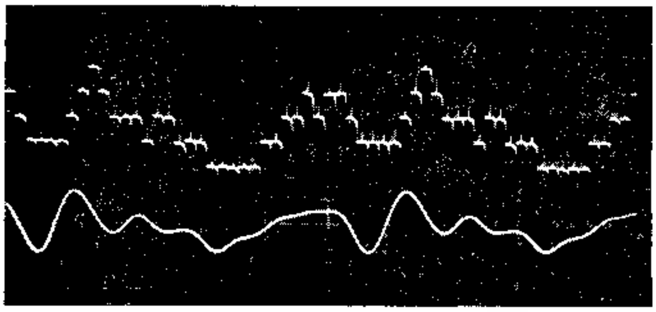


Fig. 4. Filtered code and step signal

has a small negative value. Both differences, however, can be made as small as desired by making N sufficiently large. For these reasons a maximal length shift register code can be designated as a pseudo-random binary sequence. The number N which is required, depends on the problem for which the noise generator is applied. Generally it can be stated that the period $(2^N-1)T_c$ of the code has to be large compared with the response time of the tested circuit.

Pseudo-Random Step Function

Now this binary sequence might be converted into a noiselike signal by applying it to a low-pass filter. The structure of the codes, however, suggests another, more elegant method of conversion. Because all possible combinations of positive and negative bits, except all bits positive, occur once in a cycle in the content of the shift-register, the number of positive bits will have a binomial probability distribution, with the one known exception. So when a signal is constructed, proportional in amplitude to the number of positive bits in the shift register, this signal will have a binomial amplitude distribution. As is well-known, this distribution is the best possible approximation to a Gaussian distribution when the number of possible states is finite. Now this pseudo-random step function can be generated very simply by adding together the contents of all the shift register stages by means of a network of N equal resistors. As an example, the code and the random step function generated by the circuit in Fig. 1 are shown in Fig. 3. The random step function has the form 0 1 2 3 3 3 3 2 1 1 2 1 2 2 3 3 4 4 4 4 4 3 3 3 2 2 3 2 2 2 1 etc. The distribution function is 1, 5, 10, 10, 5 which is nearly binomial.

The random step method is not too different from the low-pass filter method. In fact the action of a low-pass filter can be simulated by adding together a number of delayed versions of the same signal. As these delayed signals are already present in the shift register, it is rather obvious to make use of them. That the shift register-adder combination acts as a low-pass filter is shown in Fig. 4, where a filtered binary sequence and the step signal are compared. It can also be shown by calculating the spectrum of the step signal.

This could be done in a straightforward manner, but it is most easily solved by means of an analogy. It will be noted that the autocorrelation function, as shown in Fig. 2, is the same as that of a positive pulse of width T, repeated after a period T_c. The autocorrelation function of the step signal is therefore the same as that of the pulse after passing the shift register-adder combination. This of course is a pulse of width NT. Now, because the spectrum is fully determined by the autocorrelation function by means of the Wiener-Khintchine relation, the power spec-

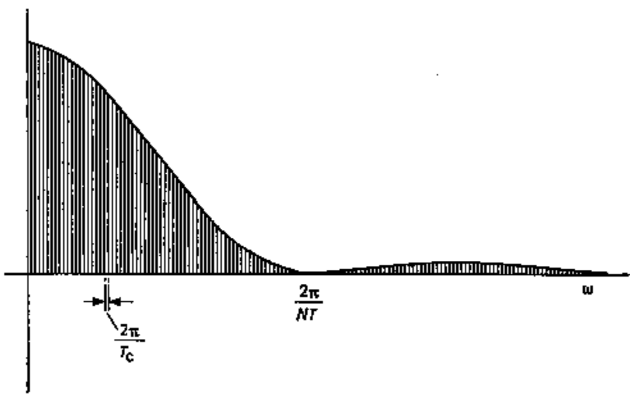
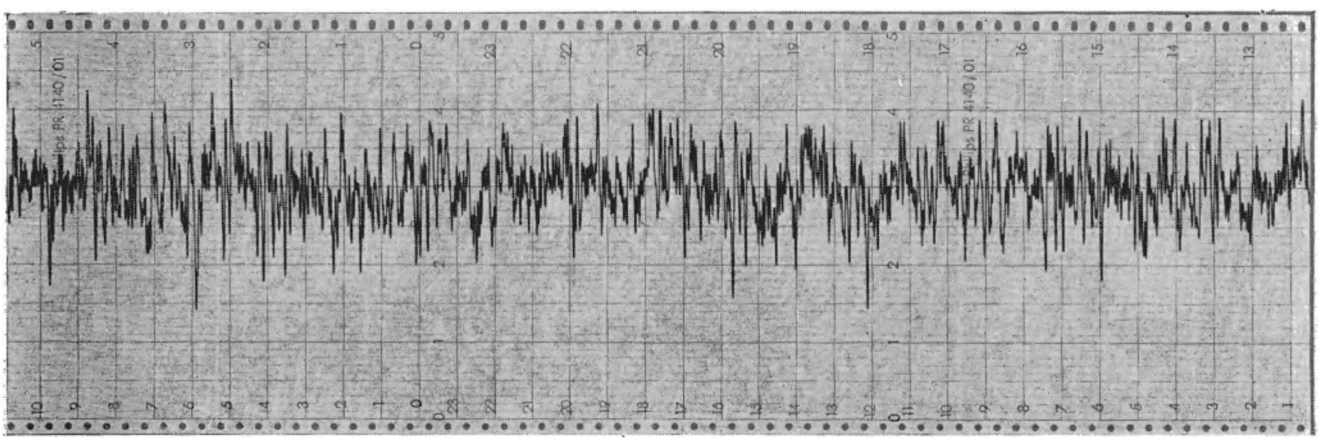


Fig. 6. Noise output from the generator

Fig. 5. Spectrum of the step signal



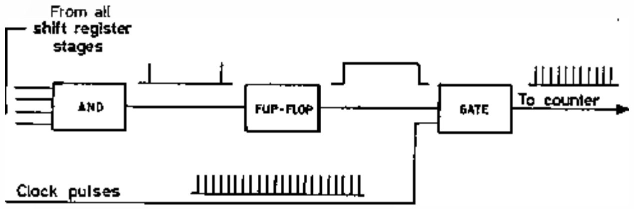


Fig. 7. Circuit for checking the code length

trum of the step signal is the same as that of a pulse of width NT, repeated with a period T_c. This spectrum is a line spectrum with an envelope $\left\{\frac{\sin \omega NT/2}{\omega NT/2}\right\}^2$ and a line spacing of $2\pi/T_c$ (Fig. 5).

Realization and Results

A noise generator, based on the principles described above, has been constructed, with a shift register of 16 stages and a clock frequency of 100c/s. The length of the generated code is $2^{16} - 1 = 65535$, so that the generated pseudo-noise has a period of nearly eleven minutes. This noise is shown in Fig. 6. The bandwidth of the noise to the first zero is $1/NT = 100/16$ c/s $= 6{\cdot}25$c/s. Higher frequencies are filtered out by means of a simple low-pass filter to smooth the signal.

A difficulty which occurs when one tries to build such a code generator is that it has to be determined whether indeed the code has the maximum length. This can be determined by the circuit given in Fig. 7.

The AND gate determines the moment that all shift register stages are in the negative state, which occurs once in every period. By the output of the AND gate a flip-flop is triggered, so this will give off a '1' output on alternate periods. The output of the flip-flop controls a gate which directs clock-pulses to a counter. Thus the counter directly indicates the length of the code. By this means it was determined that in the 16 flip-flop code generator a code of the maximum length 65535 was indeed generated when the inputs to the parity check network were taken from the flip-flops 1, 3, 12 and 16. The determination could be accelerated by running the shift register on a much higher clock frequency.

The completed code generator is constructed of standard digital circuit blocks, mounted on printed wiring boards. The generator has been applied to the solution of a problem on a non-linear noisy oscillator by an analogue computer, and in this application it gave satisfactory results.

Conclusion

In the foregoing it has been shown that a nearly random noiselike signal with exactly known and stable characteristics can be generated by standard digital techniques. Even with only a limited number of components a very good approximation to real noise can be realized. The distribution function is exactly binomial.

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An Automatic Tablet Check Weigher

The weight variation in the output of tabletting machines for the pharmaceutical and chemical industries has to be closely controlled by law. In the most prosaic cases individual tablets are required not to vary in weight by more than a given percentage, and the overall weight of any batch of tablets must be correct to rather closer limits. In the case of dangerous drugs close control is to extremely fine limits.

Continuous monitoring of tabletting machines is neither necessary nor possible, but large drug houses with batteries of these machines in continuous operation must examine batches of tablets from each machine at regular intervals to detect trends in weight variation. An automatic check weigher for performing this task has now been introduced by the Packaging Equipment Division of Soag Machine Tools Ltd, Juxon Street, London S.E.11.

The new machine is known as the Watford tablet check weigher. It comprises a hopper-fed rotary feed table, an electro-mechanical check weighing and sorting balance, and a gating mechanism to classify the tablets and direct them to their allotted containers. These principal components are mounted, together with three containers, in and on the top bay of a metal cabinet. The next bay down contains ten individual electro-mechanical weight recording counters, and two summation counters with their associated switchgear. The third and fourth bays house electronic control equipment for the gating mechanism and pulse-shaping networks for the recording counters. The fifth (bottom) bay houses the power pack and main control switches.

In use, a batch of tablets is taken at regular intervals from each tabletting machine and fed into the hopper at the top of the check weigher. The tablets drop through holes in a rotary feed table on to a chute attached to the underside of the hopper, from where they slide at the rate of up to 70 per minute on to the weigh pan. The weight is checked and classified, as a result of which one of two gates may open, giving access to container No. 1 or No. 2. Alternatively, both gates remain closed, giving access to container No. 3. A jet of air blows the tablet from the weigh pan to the gate unit, removing not only the tablet but also any powder from the weigh pan, which might otherwise collect and falsify subsequent readings.

Since this type of equipment is of rather specialized interest Soag undertake to modify the standard unit to suit specific requirements. In the standard version the weighing mechanism can be set to give up to ten separate weight categories, each having its individual counter. The mechanism can be set to direct tablets of any category into any one of the three containers. For example, tablets in categories 1 to 4 (underweight) can be directed to container No. 1, tablets in category 5 (correct weight) can be directed to container No. 2, and tablets in categories 6 to 10 (overweight) to container No. 3. The weight limit of each category is recorded on its counter for reference. One summation counter gives a record of the number of tablets in container No. 2 (correct weight), the other gives the total in containers No. 1 and 3 (incorrect weight). Deviations in the trend of weight distribution can thus be seen and corrected immediately.

Accuracy of the standard model is $\pm 0{\cdot}003$g within a tolerance range of $\pm 0{\cdot}007$ to $0{\cdot}04$g. Maximum tablet weight is $3{\cdot}0$g. The check weigher is mounted in a standard Imhof cabinet and measures $28 \times 22 \times 60$in high overall.

A Reversible Decade Counter with Symmetrical 1242 Coding

By I. Scollar*, Ph.D.

A binary decade counter using a symmetrical weighted code (1242) is described, overcoming some of the difficulties inherent in a reversible counter previously described by the author. The decade can be readily assembled from available logic modules and is capable of counting speeds up to about one half of that of the binaries employed.

(Voir page 493 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 500)

SOME years ago the author published a simple reversible transistor decade counter in this journal[1]. This counter used pulse feedback and operated in a non-symmetrical version of the 1242 code. Thus, its output could only be displayed by means of meters responding to weighted currents, a feature considered limiting by some[2]. The need for operation with punched tape equipment made it necessary to redesign the counter so that the binary states are symmetrical, regardless of the counting direction, while still retaining the weighted read-out.

A symmetrical version of the 1242 code is:

Count	*Binary Collector* 1 2 3 4 5 6 7 8
0	1 0 1 0 1 0 1 0
1	0 1 1 0 1 0 1 0
2	1 0 0 1 1 0 1 0
3	0 1 0 1 1 0 1 0
4	1 0 1 0 0 1 1 0
5	0 1 0 1 1 0 0 1
6	1 0 1 0 0 1 0 1
7	0 1 1 0 0 1 0 1
8	1 0 0 1 0 1 0 1
9	0 1 0 1 0 1 0 1
Weights	1 2 4 2

Using readily available commercial flip-flops and gates, it is possible to construct a simple counter operating in this code[3]. A decade is shown in Fig. 1. Following the table, all even numbered transistors are initially reset to zero, a state which will be taken as corresponding to conduction. A one will be equal to the negative supply voltage, approximately. A three terminal AND gate is driven from the second, sixth and seventh transistors. For the first three impulses, the gate is closed since transistor six is in conduction. At the end of the fourth impulse, transistor six blocks. The gate is still closed because transistor two is conducting. At the fifth impulse, transistor two blocks and all three gate lines are now negative. A negative voltage appears at the output of the gate and is applied, via isolating diodes, to the bases of transistors three and six. This causes them to conduct. The transistor pair 3,4 changes state as does the pair 5,6. The change of state of 6 produces a carry impulse which causes the pair 7,8 also to change state. With transistors six and seven now conducting, the AND gate is closed. The count has been advanced from the equivalent of binary four to binary eleven. Further counts proceed in a normal binary fashion until the ninth count is completed. The tenth count produces carries in all binaries and the decade returns to zero.

The state of the decade can be observed in a number of ways. Since the sum of the ones times the weighting factors shown in the table is equal to the decimal count, a simple current meter can be used. Alternatively, a diode matrix which derives suitable voltages to drive lamps can be connected and, using all collectors and twenty-eight diodes plus ten resistors, the job is done.

Looking at the table, it will be observed that every count state is the binary complement of another state (obtained by interchanging zeroes and ones). This other state is, at the same time, the nine's complement (the number subtracted from nine) of the first state. The symmetrical 1242 code shares this property with the excess three code for which a reversible counter has been described[4], but the latter code is not weighted and cannot be used with meter read-out.

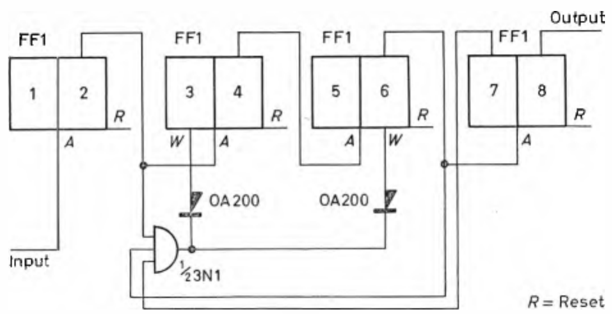


Fig. 1. Decade counter for symmetrical 1242 code

If it is desired to subtract counts from a total stored in the counter, it is merely necessary to take carries from the complementary transistor in each binary and to arrange the pulse advancing scheme so that the counter skips when passing from five to four in the reverse direction in a fashion similar to that used when going from four to five in the forward direction. A scheme which accomplishes this is shown, using commercially available flip-flops and gates, in Fig. 2. When adding, a negative voltage is applied to the add line and zero to the subtract line. When subtracting, these voltages are inverted. The designation add and subtract is of course arbitrary, given the symmetry of the code. The add-subtract voltages may be supplied from a binary or other switching element in another part of the circuit or, if desired, an anti-coincidence gating arrangement, such as that described by Tassinari[4], can be used, giving two separate inputs, one for adding, the other for subtracting.

Each add and each subtract line controls a set of four AND gates which select the even or odd sets of transistors and allow carry pulses to be delivered to the following stages. The pulses are applied to an OR gate and the drive level to the following binary is raised by means of an emitter-follower. With the gates shown, an additional resistor is needed to bring the drive voltage up to the manufacturer's specifications. If binaries similar to those shown in the author's earlier article are used, the emitter-follower and the OR gate may be dispensed with.

* *Reinisches Landemuseum, Bonn.*

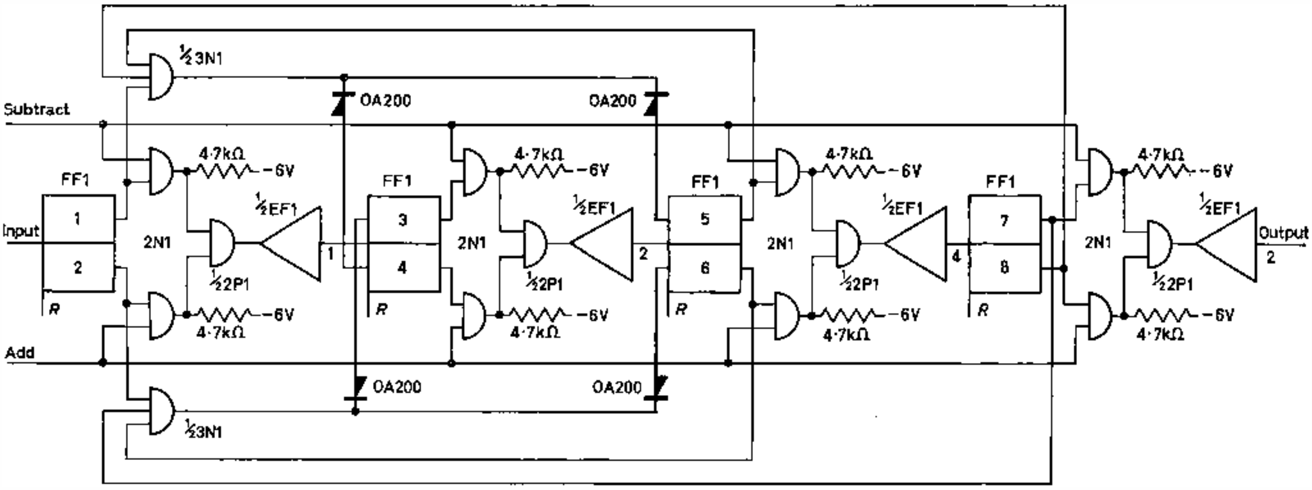


Fig. 2. Reversible decade counter for symmetrical 1242 code

The three terminal AND gate described in connexion with the counter shown in Fig. 1 is also used in that of Fig. 2. One is needed for each counting direction. When adding, the enabling voltages are derived, as before, from the second, sixth and seventh transistors and applied to the third and sixth bases. When subtracting, the enabling voltages are derived from the complementary collectors, one, five and eight and applied to the complementary bases, four and five. The two loops are not interacting for, as can be seen from the table, the states governing the operation of the AND gates are unique for each counting direction.

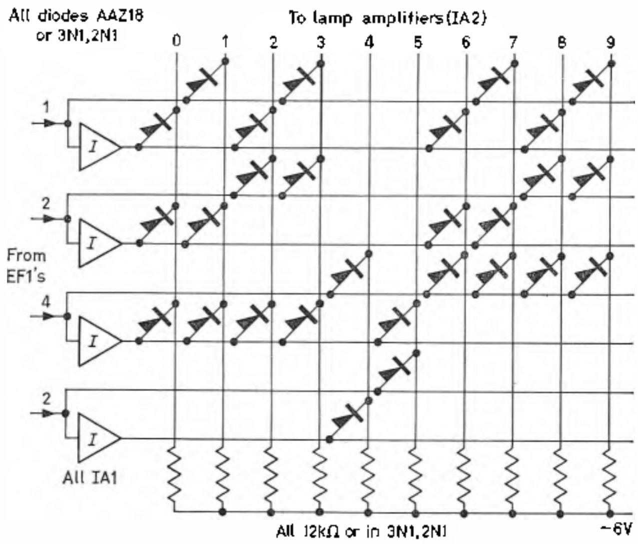
The rapidity with which the add-subtract lines can be inverted depends on the sensitivities of the flip-flops used. If they are such that the half voltage pulse, which sometimes occurs when switching, does not trip the following binary, then the counter can be reversed at a rate equal to roughly half the maximum counting rate in either direction. With the units shown, the maximum single direction rate observed at room temperature was 240kc/s, using input pulses of 6V amplitude, 2μsec width and 0·2μsec rise time. At other temperatures, this rate will not be achieved but performance will still be well above 100kc/s. In a given application, a somewhat higher single direction counting rate can be obtained at the expense of rapidity of add-subtract switch-over if, instead of limiting the sensitivity to the half voltage pulse, the rise time of the change is limited, as suggested in the author's earlier article. Using this method, the full maximum counting rate in a single direction can be achieved with a switch-over rate equal to about a tenth of the counting rate. In principle, the circuit can be applied using much faster flip-flops than those used in the experimental set-up.

Display of the output of a reversible decade counter poses some problems. If the weighted output is used to drive current meters, then it is necessary to switch the weighting lines to the complementary collectors if a count should go negative, in order to avoid the presentation of negative values as nines complements. In the earlier counter described by the author this was, in fact, the only practical method. In the counter shown in Fig. 2 it is possible to obtain an in-line read-out with correctly displayed negative numbers. There are several possible schemes for doing this, one of which is shown in Fig. 3. In this arrangement a diode matrix, in practice made up of a set of two and three terminal AND gates, is used to transform the 1242 indication into decimal values. If indication of a negative value by means of a nines complement is satisfactory, then it is merely necessary to connect the eight diode input lines to the eight collectors of the counter. If complement presentation is to be avoided, then the lines are connected, via alternate invertors, to the emitter-followers located between the stages. When a negative count total obtains, as observed when the last counter in a chain produces an output pulse going through zero to nine, it is merely necessary to interchange the state of the add-subtract lines following the end of the count and the complement of the stored count will be displayed by the lamp matrix. Thus a negative count is displayed as a positive value. One additional count, an end-around-carry pulse, will have to be given to the input of the counting chain so that there is no ambiguity about the sign of zero.

Another arrangement, usable if there is no need for speed in display, simply makes use of a four pole relay, connecting the matrix to either one or the other set of collectors as needed. The relay is actuated by an amplifier and flip-flop which changes state on receipt of the going-through-zero pulse.

A more elegant possibility involves the use of a set of four intermediate storage flip-flops with gate control

Fig. 3. Lamp matrix for symmetrical 1242 code



lines driven from the binaries in the counter. On receipt of a transfer pulse, the storage flip-flops take up a position analogous to that of the counter flip-flops. The diode matrix is attached directly to the storage flip-flops without invertors. This method has the advantage of providing a steady display. To obtain the complement of a negative count, a delayed complementing pulse is introduced via separate inputs to the storage binaries.

In the application for which this reversible counter was designed, a portable differential proton precession magnetometer operated from batteries, lamp output is not practical due to the high current required. Hence meter indication is used. At the same time, the results of the measurements are recorded on punched tape using the binary states to derive signals for the tape punch magnets.

Negatives are left in the nine complement form for, in the application in question, no value exceeds 500. Hence there is no confusion between negative and positive values and no need to prepare special sign indications for the computer. The sign is automatically indicated by the value of the highest digit which is less than five for all positive numbers and five or more for all negative values.

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Harmonic Linearization of the Wien Bridge Oscillator

By M. A. Kaaz*, B.Sc

In order to obtain stable operation of an oscillator a certain non-linearity must be present in the system: this must be sufficient to enable adequate automatic control of amplitude but should introduce the least possible harmonic distortion. An analysis is given for this aspect of the Wien bridge oscillator using the concept of the describing function.

(Voir page 493 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 500)

OF prime interest in the design of oscillators is the stability of the amplitude and frequency of oscillation; in sinusoidal oscillators one assumes naturally a good measure of purity of the waveform. For special applications, there are also noise generation problems to consider[1,2].

In general amplitude, frequency and waveform will all be interdependent quantities and, usually, a practical compromise will have to be reached between them. An analysis concerned with the general behaviour of an oscillator must necessarily consider the non-linearity (or non-linearities) of the system without which stable oscillation is not possible. In a sinusoidal oscillator the non-linear characteristic should, however, be barely sufficient to produce adequate automatic control of the amplitude, yet introduce least harmonic distortion. Such a system will, nevertheless, require the solution of a set of non-linear differential equations or the application of the eminently suitable concept of the describing function, as in this case.

Harmonic Linearization

In what follows, it will be assumed that the oscillator is linear except for the slightly non-linear characteristic of the amplifier given by

$$x = ay - by^3 \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (1)$$

It is further desired and assumed that

$$y = A \sin\psi,\; A = \text{const} \quad \ldots\ldots\ldots\ldots \quad (2)$$

Since x in equation (1) is an odd function it will, if expanded in a Fourier series, contain only the fundamental and 3rd harmonic.

$$x = aA \sin\psi - bA^3 \sin^3\psi \quad \ldots\ldots\ldots\ldots \quad (3)$$

By definition, the describing function of $x = f(y)$ neglects all but the fundamental part which is denoted as

$$x_1 = A_1 \sin\psi \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (4)$$

The coefficient A_1 is calculated, in conformity with Fourier analysis, from

$$A_1 = 1/\pi \int_0^{2\pi} f(y) \sin\psi \, d\psi$$

$$= aA/\pi \int_0^{2\pi} \sin^2\psi \, d\psi - bA^3/\pi \int_0^{2\pi} \sin^4\psi \, d\psi$$

Applying the identity $\sin^2\psi = \dfrac{1-\cos 2\psi}{2}$ twice, yields

$$A_1 = (a - \tfrac{3}{4} bA^2) A \quad \ldots\ldots\ldots\ldots\ldots \quad (5)$$

The describing function or the linearized gain of the non-linear amplifier is, accordingly, denoted*

$$T_1 = A_1/A = a - \tfrac{3}{4} bA^2 \quad \ldots\ldots\ldots\ldots \quad (6)$$

and equation (4) now reads

$$x \simeq x_1 = T_1 y \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (7)$$

This is a linear relationship, although T_1 is a function of the amplitude of the input y and it is now legitimate to use linear analysis and the very useful Laplace transformation.

The Feedback Filter

In the Wien-bridge-oscillator, this consists of an RC four-pole (Fig. 1)

It is selective in the sense that y is the derivative of x

* *It should be borne in mind that A is a function of initial conditions only. In fact*
$y'(t) = (dy/dx)\,(dx/dt) = A_1\omega/T_1 \cos\omega t$; thus $y'(o) = A_1\omega/T_1$ or $A = y'(o)/\omega$ (see ref. 7).

* Honeywell GMBH.

at low frequencies and the integral of x at high frequencies. Its transfer function

$$G(s) = Y(s)/X(s) = \frac{\omega s}{s^2 + 3\omega s + \omega^2} \quad \ldots\ldots\ldots (8)$$

where

$$\omega = 1/RC \quad \ldots\ldots\ldots\ldots\ldots\ldots (9)$$

This filter is stable by itself. Initial and final limit tests confirm this.

$$\lim_{t \to 0} g(t) = \lim_{s \to \infty} sG(s) \to \omega$$

$$\text{and} \lim_{t \to \infty} g(t) = \lim_{s \to 0} sG(s) \to 0$$

The Oscillator

It is now a simple matter to calculate the stable amplitude of oscillation (the frequency is given by equation (9)). The closed loop equation for Fig. 2 is

$$Y \,.\, G \,.\, T_1 = Y$$

$$\text{or} \quad Y(1 - T_1 G) = 0 \quad \ldots\ldots\ldots\ldots (10)$$

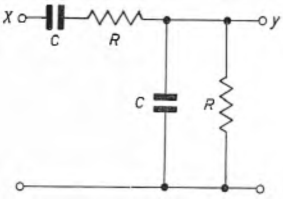

Fig. 1. RC feedback filter

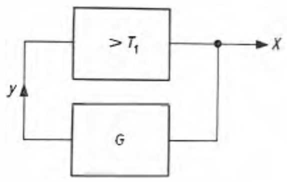

Fig. 2. The oscillator loop

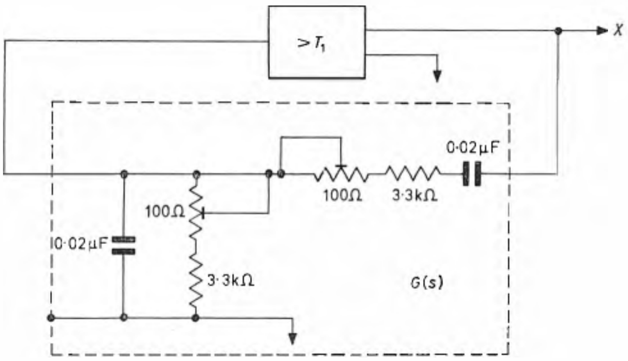

Fig. 3. A practical example

Aside from the trivial case $Y = 0$, equation (10) is satisfied if

$$1 - T_1 G = 0 \quad \ldots\ldots\ldots\ldots\ldots\ldots (11)$$

This is the well-known Nyquist instability condition. Inserting equation (8) into equation (11) yields

$$s^2 + \omega(3 - T_1)s + \omega^2 = 0 \quad \ldots\ldots\ldots\ldots (12)$$

Stable oscillation is possible if the damping coefficient vanishes, i.e.

$$T_1 = 3 \text{ or } A = 2\sqrt{\left(\frac{a-3}{3b}\right)} \quad \ldots\ldots\ldots (13)$$

$$\text{and } s = \pm j\omega.$$

The conclusion from relation (13) is that the amplitude of y, and consequently also that of x, depends only on the characteristic of the amplifier*. This should, therefore, be determined very accurately and remain constant during operation.

* *This conclusion is in no way conflicting with the previous footnote. They are mathematically* ***equivalent.***

The amplitude stability problem has an exact analogy in mechanics. It will be recalled that a mechanical system is absolutely stable at a point (x_0, y_0) if the potential energy there is least and if its gradient changes sign at (x_0, y_0). The gradient is analogous to damping, which is capable of changing sign and vanishes when the amplitude is stable. The change of sign of the damping is in effect the movement of poles along the real axis in the complex plane. They must initially lie in the right-half plane for the oscillation to build-up and must then fall on the j-axis for stable oscillation. Equation (11) yields another interesting result if it is transformed into the time domain after re-writing, thus

$$T_1 G = 1$$

This leads to:

$$\omega T_1 \, e^{-(3/2)\omega t} \left[\cosh \Omega t - (3/2)(\omega/\Omega) \sinh \Omega t\right] = \delta(t-0) \quad \ldots\ldots\ldots\ldots (14)$$

where $\Omega = (\sqrt{5}/2)\,\omega$

and

$\delta(t - 0)$ is the Dirac impulse at $t = 0$

Particularly, at $t = 0$

$$\delta(t) = \omega T_1 \quad \ldots\ldots\ldots\ldots\ldots\ldots (15)$$

i.e. a frequency distribution ωT_1 is necessary at the instant of switching on in order for the oscillator to reach its stable amplitude at once*.

A Practical Example (Fig. 3)

The measured characteristic of the amplifier is given as

$$x = 3{\cdot}03y - 0{\cdot}01y^3$$

The desired frequency $\omega/2\pi = 2400$c/s.

Therefore

$$RC = 1/4800\pi$$

Suitable values are:

$R = 3{\cdot}3\text{k}\Omega$ metal film resistor + 100Ω wirewound potentiometer

$C = 0{\cdot}02\mu\text{F}$ mica capacitor.

From equation (13) $A = 2$V.

Hence $T_1 = 2{\cdot}73$ and $x_1(t) = 5{\cdot}46 \sin(4800\pi t)$ V.

A carefully designed transistorized unit with adequate thermal control is capable of operating stably within the following specification:

$$A = 2\text{V} \pm 1 \text{ per cent}$$

$$\omega/2\pi = 2400\text{c/s} \pm 0{\cdot}005 \text{ per cent}$$

Harmonic distortion $\leqslant 0{\cdot}5$ per cent.

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* *ωT_1 is an infinitely large quantity, impossible to realize. It may be written as $\omega T_1 = T_1/A\,y'(0)$ in which T_1 and $y'(0)$ are finite, but A is zero at the start. The amplitude A will, therefore, take some time to establish itself.*

Improved Technique for Photoelectric Measurements

By M. D. Matthews*, E. A. Faulkner*, and G. W. Green*

In the measurement of photoelectric thresholds it has been found necessary to measure photocurrents down to 10^{-15}A. This is best done by chopping the incident exciting radiation and using an a.c. system in which a very narrow bandwidth is obtained by the use of a phase-sensitive rectifier. This article describes in detail how such an a.c. system may be assembled from commercially available electronic components and discusses the advantages that this method has over the more conventional arrangement employing a d.c. electrometer.

(Voir page 493 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 500)

FOR many years, the conventional method of measuring small direct currents in the range 10^{-8} to 10^{-15}A has been by d.c. amplifiers having an electrometer valve input[1]. By very careful design[2], these instruments can be used to detect currents as low as 5.10^{-16}A, but their notorious susceptibility to drifts makes them unsuitable for general laboratory use.

The problem of zero drift can be partially overcome by converting the signal to a.c. and using conventional tuned amplification and detection. The vibrating reed capacitor is the most commonly used modulator[1]. This system, however, still suffers from a number of disadvantages.

These originate because the electrometer detects all voltage or current signals having periods greater than the response time of the instrument, which means that any spurious d.c. signals arising in the input between source and modulator are detected and will contribute to drift at the output. Stray signals may arise from unstable contact potential differences or from thermoelectric and thermionic effects. These unwanted signals, which are often difficult to eliminate in practice, considerably reduce the signal-to-noise ratio of the instrument.

Furthermore, in order to develop sufficient voltages (10^{-3}V) across the modulator for small currents in the 10^{-15}A range, very high input resistors ($10^{12}\Omega$) are necessary. This presents problems of source insulation which must be greater than $10^{12}\Omega$; particular care must be taken to prevent surface contamination of insulators especially by water vapour. The high input resistance means that input capacitance must be reduced to a minimum by placing the electrometer as close as possible to the current source in order to avoid long electrical time-constants. With the electrometer close to experimental apparatus, it often becomes susceptible to mechanical, electrical and thermal shocks. It is commonly found that such disturbances can cause 'noise' currents of 10^{-12}A or more which decay with very long time-constants and therefore appear as zero drift.

The above problems of zero shift set a practical limit to the usefulness of the instrument and the theoretical performance (governed by Johnson noise in the input resistor) is difficult to achieve.

These difficulties may be overcome in the particular case of photoelectric measurements where the d.c. to a.c. conversion can be performed by chopping the incident ultra-violet radiation. The signal is then a.c. at the outset and since stray signals are not modulated much improved signal-to-noise ratios result. This superior sensitivity enables currents in the 10^{-15}A range to be detected using a smaller input resistance ($10^{9}\Omega$) so that leakage problems become unimportant. A detector bandwidth of 0·1c/s may easily be achieved by the use of phase-sensitive signal rectification giving a signal to Johnson noise ratio only a little less than 10dB for a current of 10^{-14}A peak-to-peak and an input resistance of only $10^{9}\Omega$.

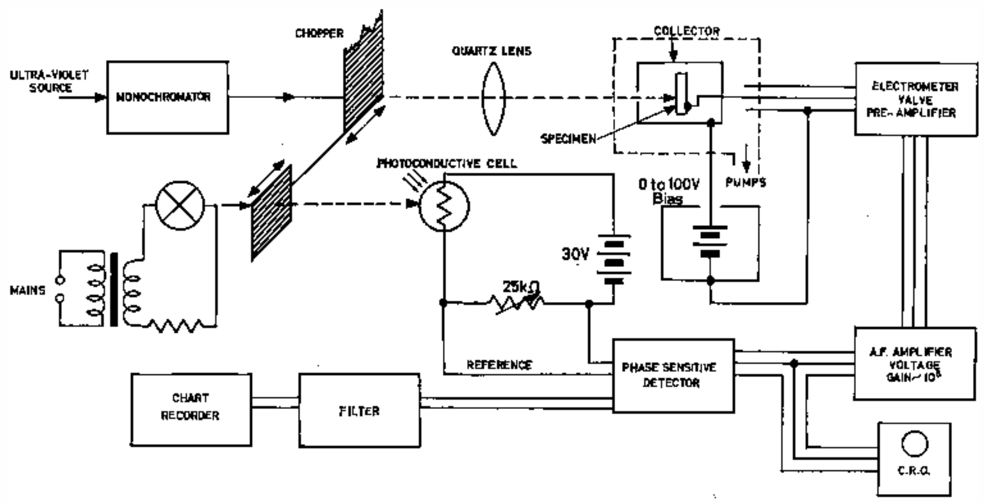


Fig. 1. Circuit for the measurement of photo-emission currents

Despite the obvious advantages of this a.c. system only a few workers[3] have adopted it because of the difficulties experienced in constructing phase-sensitive detectors of suitably high performance. Now, however, the development in this laboratory of a general purpose phase-sensitive detector[4]* has made the use of the a.c. system a much more attractive proposition. This article describes in detail how this detection system can be easily and cheaply set up.

* *J. J. Thomson Physical Laboratory, University of Reading*

* *Commercial versions of the phase detector and main amplifier used in this work are supplied by Brookdeal Electronics Ltd.*

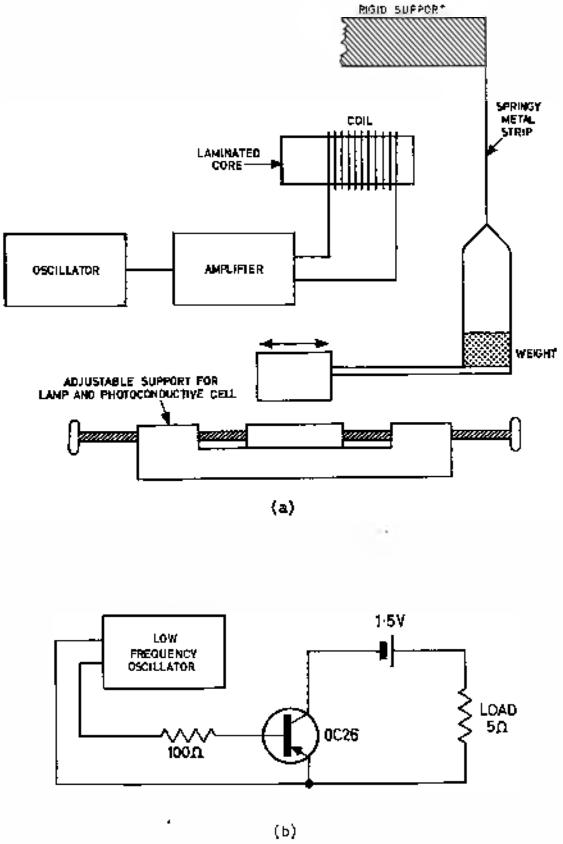


Fig. 2. (a) The chopper system
(b) Chopper drive circuit

Circuit Details

The arrangement of the circuit is shown in Fig. 1. The essential components are the chopper, pre-amplifier, main amplifier and phase-sensitive detector.

The chopper, shown in Fig. 2(a), is mounted immediately next to the exit slit of the monochromator used to select the wavelength of the incident radiation. The essential feature of the chopper is lightness of construction to avoid microphonic pick-up resulting from vibrations set up in the experimental tube. A springy mild steel arm (a 4·5in length of 0·25in × 0·15in packing case strip) is caused to vibrate at its natural frequency (~10c/s) by being placed in a periodically varying magnetic field supplied by a small exciter coil. This coil was made by winding about 50 turns of 32 gauge copper wire on a piece of laminated transformer core (approximate dimensions 1·0in × 0·5in × 0·25in). The energizing signal is supplied by a transistorized low frequency oscillator†. As this oscillator delivers little power, a simple amplifier (Fig. 2(b)) was made. An OC26 transistor connected into the exciter coil circuit is used as a switch, being controlled by the oscillator signal. The power dissipated in the coil is then determined purely by the battery voltage selected. In this case, with a coil resistance of about 5Ω, a voltage of 1·5V gives ~0·5W.

A low frequency of operation is desirable because of the time-constant of the detector system with $R_{in} = 10^9\Omega$, and $C_{in} = 10$pF. It is also necessary, in order to avoid microphonic pick-up, to select an operating frequency well away from any natural frequencies of input components. The chopper frequency is readily adjusted by either changing the length of the vibrator arm or by adding a suitab weight.

The reference signal for the phase-sensitive detect is obtained by connecting a second vane to the vibrat arm and causing this to interrupt a light beam falling c a photoconductive cell (ORP60) as shown in Fig. 1. . voltage of suitable level is tapped off across a variab resistor in series with the photoconductive cell. No refe ence phase shift is necessary because no phase shift introduced by the amplifier and phase detector.

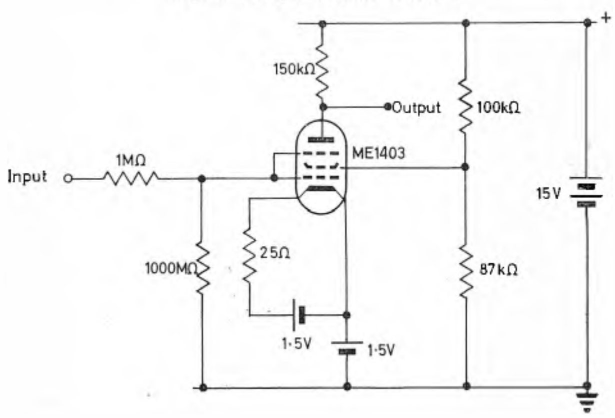
The ultra-violet beam is focused on to the specimen wi quartz optics. The photoelectric emission current collected on a screen biased at a minimum positive volta of 20V with respect to the specimen. To avoid spurio currents originating from surfaces other than the or being studied, the current through the specimen rath than the collected current is measured. Good electric shielding of the current source has been achieved by coa ing the Pyrex glass vacuum envelope with a conducti layer of graphite (Aquadag). The electrical leads a brought out through the vacuum envelope by means tungsten-C9 glass seals.

The signal is amplified before being fed into the phas sensitive detector. The pre-amplifier is a battery-operat electrometer valve circuit as shown in Fig. 3; there is n need to compensate for drifts of battery voltages becau only a.c. signals are involved. The pre-amplifier, its mai function being as an impedance convertor (10^9 to 10^5Ω gives only a small voltage gain ~3. The circuit an batteries are mounted in a strong light-tight metal bc forming a compact robust unit, which can readily be co nected directly to the experimental tube to avoid cab capacitances. A 3ft length of good quality low noise cab is used to connect the pre-amplifier to the main amplifie

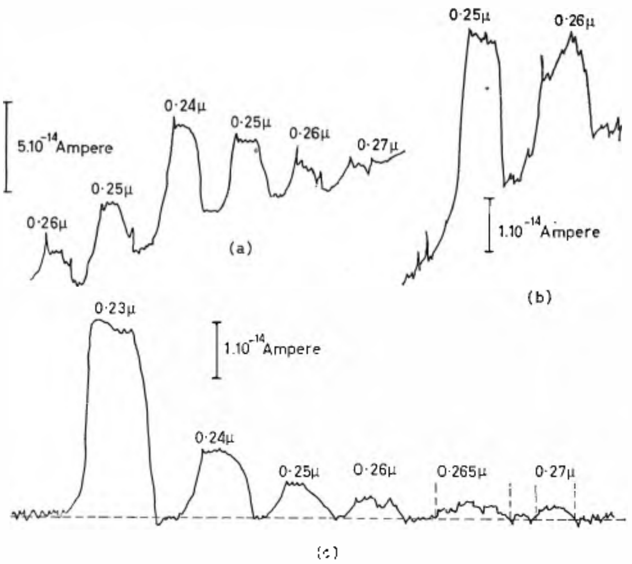
The main amplifier is a general purpose instrume which has been developed in this laboratory and is no available commercially. The circuit is a conventional o employing three ECC83 double triodes giving a max mum gain of 95dB and a distributed attenuator whic enables the gain to be reduced to 20dB in steps of 5d

The signal is simultaneously fed into the phase detect and displayed on a c.r.o. so that the noise level can monitored; the input signal is below the noise level f currents less than 10^{-13}A. The integration time-consta of the detector is brought up to about 5sec by 5 000μ capacitors across the output[4], which is connected to chart recorder so that changes in photoelectric curre with wavelength can be recorded. The filter (Fig. 1) r moves 10c/s components from the signal to avoid p vibrations; a multi-stage filter is used rather than a sing *RC* circuit to achieve sufficient filtering without undu increasing the time-constant.

Fig. 3. The pre-amplifier circuit



† *Manufactured by Levell Electronics Ltd (Type TG 150).*



ig. 4. **Spectral response curves for cleaved germanium surfaces of area about 1 mm²**

) and (b) d.c. method (c) chopped-beam method (different surface) he monochromator is set at intervals of 0·01 micron. Each setting is aintained for 30sec, with 30sec intervals during which the monochromator is shut off to establish the zero line

'onclusions

A performance very near to the theoretical expecta- on was achieved in practice. Typical chart traces of hotoelectric currents and their variation with wavelength are shown in Fig. 4 for both the d.c. (electrometer having $R_{in} = 10^{13}\Omega$) and a.c. methods. The zero stability of the a.c. method is clearly apparent.

The advantage of the a.c. detection system may be summarized as follows.

(1) Much improved base-line stability; small d.c. signals of thermionic or thermoelectric origin are not detected.

(2) Satisfactory sensitivity (10^{-15}A) achieved using a comparatively low ($10^{9}\Omega$) input resistance.

(3) Rapid recovery time (~10sec) after mechanical, electrical or thermal shocks.

(4) Short warm-up time; the system could be used only a few seconds after being switched on.

(5) Main components i.e. main amplifier and phase detector commercially available making the circuit straightforward to construct.

The a.c. method has now been used successfully in this laboratory for a number of measurements of photoelectric thresholds of clean germanium surfaces produced by cleavage in ultra-high vacuum.

Acknowledgments

The authors wish to thank Professor R. W. Ditchburn for the provision of research facilities and the Admiralty for their support.

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A New Doppler Navigator

A new Doppler navigator, type AD570, employing a small ghtweight aerial system designed specifically to fit the range f smaller airliners in service today, has been developed by he Marconi Co. Ltd. Precise measurement of aircraft ground peed, drift angle and distance flown can be made with this ew system.

The AD570 is a self-contained Doppler velocity sensor quipment. It comprises an aerial unit type AA-5701, a trans- nitter/receiver type AA-5702, a tracker unit type AA-5703 and 'ground speed and drift' indicator type AA-5604. A 'dis- nce flown' indicator can also be supplied as an optional xtra.

The equipment is fully transistorized and employs sealed odular construction techniques. It meets all the operational quirements of ARINC Characteristic No. 540 and provides e outputs necessary for a navigation computer conforming ith ARINC Characteristic No. 543. The speed range 90 to 000 knots is offered as standard, although other ranges can e made available to cover the requirements of STOL machines nd supersonic aircraft.

The aerial is considerably smaller than those used in other larconi Doppler systems. This has been achieved by selecting higher operating frequency. The aerial dimensions are 17½in 11in × 5 9/16in. All the servo mechanisms have been in- orporated within the body of the aerial itself and, although rugged construction, the complete assembly weighs only 12lb.

The transmitter has a nominal power output of 1W and e system has been designed to operate up to a height of 0 000ft. The 1W output is time-shared to each of the four eams in turn, which ensures that the maximum energy is diated by each beam.

By careful aerial design, the effect of 'water bias' on the AD570 system has been reduced to an average of only 0·25 per cent. 'Water bias' is an effect suffered by all Doppler systems when flying over water, and is caused by unequal amounts of energy being scattered back from different parts of the beam. This effect varies according to the nature of the water surface. Due to the very low 'water bias' of the AD570 system there is no need for a land/sea switch.

The AD570 has been designed to operate at 13·3Gc/s and has a frequency modulated continuous wave output of 1W. It employs a Doppler stabilized moving aerial, containing a linear array of slotted waveguides producing four separate beams. A null tracking system is used, whereby the aerial orientation is continuously adjusted to produce equal Doppler shifts on all of the four beams.

The system is entirely automatic in operation and only requires a one minute warm-up period. Once a usable signal has been received, the tracking system will align the aerial along the aircraft's flight path and indications of ground speed and drift angle will be shown on the appropriate indicators.

If the signal-to-noise ratio falls below a workable level, the system will automatically preserve the last accurately determined values of ground speed and drift angle until the quality of the signal improves. This is known as memory operation and a small flag, mounted in the 'ground speed and drift angle' indicator, will be shown when the equipment is operating in this state. Only when three of the beam signals are being correctly tracked and the speed, drift and pitch servos are correctly locked-on, will the memory flag be with- drawn and the system revert to normal operation.

A comprehensive check of the system can be made during both normal and memory operation, and this is achieved by manually operating a 'confidence' test button.

LETTERS TO THE EDITOR

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

Matrix Analysis

DEAR SIR,—Some time ago, out of sheer curiosity, a friend of the writer suggested that it would be of interest to analyse the circuit of Fig. 1 by matrix methods.

It was noted that the output terminals 2 and 3 are 'floating', and it is, therefore, not possible to treat the system as a three-terminal network. Readers may find the following method, which is briefly outlined, of academic interest. For brevity, use of the method will be confined to finding the output impedance of the circuit (looking into the terminals 2 and 3). See Fig. 1.

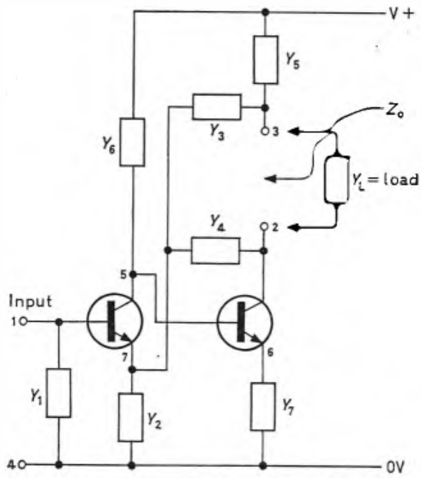

Fig. 1. Two-stage feedback amplifier

Assuming the two transistors are identical, the indefinite admittance matrix of the circuit of Fig. 1 may be represented as:

$$[Y]_i = \left[\begin{array}{cccc|ccc} Y_1+y_i & 0 & 0 & -Y_1 & y_r & 0 & -y_i-y_r \\ 0 & Y_4+y_o & 0 & 0 & +y_f & -y_f-y_o & -Y_4 \\ 0 & 0 & Y_3+Y_5 & Y_5 & 0 & 0 & -Y_3 \\ -Y_1 & 0 & -Y_5 & +Y_t & -Y_6 & -Y_7 & -Y_2 \\ \hline y_f & y_r & 0 & -Y_6 & Y_6+y_i+y_o & -y_i-y_r & -y_f-y_o \\ 0 & -y_r-y_o & 0 & -Y_7 & -y_i-y_f & Y_7+D_y & 0 \\ -y_i-y_f & -Y_4 & -Y_3 & -Y_2 & -y_r-y_o & 0 & Y_2+Y_3+Y_4+D_y \end{array}\right]$$

where $Y_t = Y_1+Y_2+Y_5+Y_6+Y_7$

$D_y = y_i+y_r+y_f+y_o$

Partitioning as indicated by the dashed lines, the matrix is in the form

$$[Y]_i = \left[\begin{array}{c|c} Y_{11} & Y_{12} \\ \hline Y_{21} & Y_{22} \end{array}\right]$$

Since the input and output quantities are referred only to terminals 1, 2, 3 and 4, the 7×7 indefinite matrix may be reduced to a 4×4 indefinite matrix $[Y']_i$ as follows:

$$[Y']_i = [Y_{11} - Y_{12} \,.\, Y_{22}^{-1} \,.\, Y_{21}]$$

Now, this matrix corresponds to the condition shown in Fig. 2. Here, it is seen that $V_i = V_1 - V_4$, $V_o = V_2 - V_3$, then, $I_2 = -I_3$. Similarly, $I_1 = -I_4$.

If terminal 4 is brought to the reference level, i.e. putting $V_4 = 0$, the fourth row and fourth column of $[Y']_i$ may be deleted, giving the following 3×3 definite Y' matrix: (Here, the individual elements of the resultant matrix are represented as y_{ij}'s for conciseness)

$$[Y'] = \begin{bmatrix} y_{11} & y_{12} & y_{13} \\ y_{21} & y_{22} & y_{23} \\ y_{31} & y_{32} & y_{33} \end{bmatrix}$$

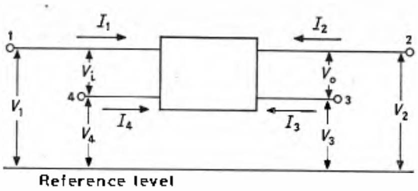

Fig. 2. Four terminal representation of Fig. 1.

Inverting this matrix gives the corresponding definite impedance matrix $[Z']$,

$$[Z'] = \begin{bmatrix} z_{11} & z_{12} & z_{13} \\ z_{21} & z_{22} & z_{23} \\ z_{31} & z_{32} & z_{33} \end{bmatrix}$$

Therefore

$$\begin{bmatrix} V_1 \\ V_2 \\ V_3 \end{bmatrix} = \begin{bmatrix} z_{11} & z_{12} & z_{13} \\ z_{21} & z_{22} & z_{23} \\ z_{31} & z_{32} & z_{33} \end{bmatrix} \begin{bmatrix} I_1 \\ I_2 \\ I_3 \end{bmatrix}$$

$$= \begin{bmatrix} z_{11} & z_{12} & z_{13} \\ z_{21} & z_{22} & z_{23} \\ z_{31} & z_{32} & z_{33} \end{bmatrix} \begin{bmatrix} I_1 \\ I_2 \\ -I_2 \end{bmatrix}$$

and

$$\begin{bmatrix} V_i \\ V_o \end{bmatrix} = \begin{bmatrix} V_1 \\ V_2 - V_3 \end{bmatrix} =$$

$$\begin{bmatrix} z_{11} & z_{12} - z_{13} \\ z_{21} - z_{31} & z_{22} + z_{33} - z_{23} - z_{32} \end{bmatrix} \begin{bmatrix} I_1 \\ I_2 \end{bmatrix}$$

The resultant Z matrix may now be cor verted to the cascade 'a' matrix to ob tain the output impedance of the circuit Putting z_g as source impedance, and I to zero, gives

$$Z_o = \frac{a_{22}z_g + a_{12}}{a_{21}z_g + a_{11}}$$

K. K. SUM,
Northern Polytechni
London.

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The Laddic in All-Magnetic Logic Circuits

DEAR SIR,—With reference to th article on the 'Laddic' by D. J. Morr in May, 1965, I wish to criticize a su gestion in his conclusion that the systei is 'fail-safe'.

When a control rung becomes d energized, the rung has an m.m.f. acro it which is alternating by the drive 1 an drive 2 supplies. In fact in the trippe state the flux tends to switch in th decontrolled rung. If a fault develops i part of the magnetic core, or a shor circuit occurs on the winding of the d controlled rung, it is not possible guarantee that the output reduces to sufficiently low level in the tripped stat with this winding configuration.

Yours faithfully,

R. HUTCHINS,
Associated Electrical Industries Lt
Leicester.

The Author Replies:

DEAR SIR,—I agree with Mr. Hutchi that the laddic is not fail-safe for th given wiring configuration.

May I point out that it was not stat in the article that the laddic is fail-saf It was stated that All-Magnetic Circu may be used in fail-safe applications. was also stated that the laddic, being magnetic device, may be used only as input element to all-magnetic circuits. could not be part of the all-magne circuits since the input must come fro a relative high impedance and thus t input to the laddic may not be of wi only.

Yours faithfully,

D. J. MORRIS,
English Electric-Leo-Marcc
Computers Ltd,
Stoke-on-Trent.

BOOK REVIEWS

Computing Methods

y I. Berezin and N. Zhidkov. Vol. 1 464 pp. ol. 2 679 pp. Med. 8vo. Pergamon Press. 1965. Price £5 (set of 2 vols.)

THIS book, which is in two volumes, has been developed from lecture ›urses at Moscow State University for udents in the mechanical-mathematical ıculty specializing in computer matheıatics. Volume 1 contains chapters on ıe rules governing operations with ›propriate quantities, methods of ›proximating functions and their ›plications and on numerical methods f differentiation and integration; olume 2 is devoted to the solution f sets of linear algebraic equations, the version of matrices, the calculation of genvalues and eigenvectors, the soluon of polynomial and transcendental juations, ordinary and partial differenıl equations and integral equations.

The work has a distinctly Russian ıvour and many methods described are ss familiar in the West. Western ıthors receive scant acknowledgment, ıt the well-known books by Milne and ouseholder are mentioned.

The range of problems covered in ıe two volumes is unusually large, ıt there is one common idea undering all the methods. This is the con›pt of functional analysis, which is ›ncerned with abstract spaces and the ınctions which points of the spaces present. The concept of distance can introduced in many ways into these ›aces and to each definition corre›onds a particular kind of approximaon (e.g. mean or least square). The ajority of problems tackled can be ‹pressed in the form: find y from x here $y = A(x)$ and x, y belong to ‹ed spaces. In general there are two ethods of tackling this problem, ›stly by changing the spaces (e.g. by placing a general function by a polyımial approximation to it) or by ıanging A (e.g. by replacing integration y summation).

One complete chapter is devoted to e theory of approximations and in ımerous other places appeal is made general theory as a basis for the ›velopment which follows. These eoretical excursions are liable to prove ther heavy going for those without a ›und mathematical background. The athematical development of the main pics is generally elementary but often ımbersome. The emphasis is someıes misplaced. For example, one will ad through a lengthy derivation to ıd that it is not suitable for prac›al applications. Again, the square›ot method of solving sets of simulneous equations is illustrated briefly and only for the case of a symmetric matrix, whereas this method can be applied generally and is believed to be the most convenient in terms of the number of operations and the accuracy of the solution. The question of the evaluation of the accuracy of methods is not expounded on the grounds that a full solution has still to be found.

There are a number of textual misprints (the translator has corrected a few in the original text) and in some chapters the brackets round equation numbers cited in the text are omitted.

As a general compendium of methods the book is bound to be useful, but readers should be aware that later and better methods of tackling some problems are available.

J. CALDWELL

Physics of Magnetism

By S. Chikazumi and S. Charap. 554 pp. Med. 8vo. John Wiley & Sons. 1964. Price 120s.

PROFESSOR CHIKAZUMI'S textbook is admirably suited for workers who require a comprehensive background in the physics of magnetism. Theoretical expositions are limited to the mathematics of a physics graduate and a correct emphasis is given to the explanation of physical phenomena. Throughout the book a balance is achieved between the descriptions of physical phenomena, the methods used for their observation and measurement, the presentation of experimental data, and the theory. The text is well illustrated and some good photographs of domain configurations are included.

The book is in six parts. The first introduces magnetostatics and some phenomenalogical effects of magnetizations, in particular the important demagnetization effect. The second part discusses the origins of magnetism in terms of orbital and spin momentum and the classical theories of paramagnetism and ferromagnetism are presented. Antiferromagnetism and ferrimagnetism are also discussed in some detail.

The third part describes magnetic domain structure. The concept of the domain is introduced and separate chapters are used to describe crystalline anisotropy, magnetostriction and magnetostatic energy. The influence of these on domain structure is described and the properties of domain walls outlined.

Part four discusses the magnetization process in terms of domain structure and the reversible and irreversible changes by rotation and wall movement. Time effect variations of magnetization such as disaccommodation and dynamic effects such as ferromagnetic resonance are also described.

Under the heading of 'Special Topics', part five covers induced anisotropy, the properties of thin films, special spin configurations, and neutron diffraction and Mossbauer techniques for determining magnetic structure. Galvanomagnetic, magnetothermal and magnetomechanical effects are also described. The concluding sixth part deals briefly with some engineering aspects of magnetic materials.

The reader is left with an overall impression of the importance of domain theory, and the clarity with which this and the other topics listed are dealt with makes this book highly recommendable.

D. C. STAPLETON

The Design and Use of Electronic Analogue Computers

By C. Gilbert. 528 pp. Med. 8vo. Chapman & Hall Ltd. 1964. Price 105s.

SEVERAL years ago when the subject of digital computing was novel and the machines awe-inspiring in their ingenuity and complexity, discussion always included the relative merits of the well established analogue computers and the new digital computers. In the circles noted for a flat emotion-response, feeling was known to run high over the issues involved.

Digital computers moved onwards and upwards. Caught in the grip of their own versatility they became hard worked; economic necessity ensured that every working second was fully occupied. The anomalous situation arose, and still exists, that a computer capable of providing a solution in a matter of seconds frequently has to be booked days beforehand and a user after weeks of preparation finally achieves access to it then retires with a scrap of paper bearing the legend:

C O F 15 READ X J
NO GO

To anyone with a certain type of what may loosely be termed a small problem, this is a frustrating situation. One is inevitably tempted to reconsider the use of an analogue machine where the pressure of economic pay load is less and there is the rewarding experience of being closer to the physical nature of the problem.

Mr. Gilbert's book is excellent in that it reminds the reader of what is possible with analogue computers; the list is impressive. He treats the whole subject in great detail and the text layout and diagrams are exemplary.

In his first chapter he establishes firmly the analogue philosophy, an approach which is maintained throughout. For example, diagrams where appropriate are in three parts, viz., the physical system, the electrical analogue and the behaviour in graph form.

He draws an interesting distinction in Chapter 4 between the analogue simulator and the analogue computer illustrating the different program techniques in the case of a second order system. He later shows how, in the case of impedance networks, complex transfer functions can be obtained from single amplifiers. The point is illustrated by examples of single and high order transfer functions. This is followed by sections on amplitude scaling factors, dynamic errors and time-scale factors.

Chapter 5 is devoted to a detailed study of auxiliary computing equipment such as diode circuits for generation of continuous and non-continuous function, non-linear resistive devices, relay circuits and multipliers.

Chapter 6 deals with hardware aspects but emphasis is placed, sensibly, on the control system and output facilities. Chapter 7 covers the general subjects of the process of problem solution, problem checking and fault investigation and miscellaneous calibration techniques.

The weakest part of the book is Chapter 3 which covers the subject of amplifier design. The content is almost entirely in terms of obsolescent hard valve techniques. It is unfortunate that Mr. Gilbert has made a case for doing this by suggesting not only that solid state amplifiers are not of comparable performance to valves but that, in any case, design techniques are so similar that the latter suffice for both. The statements, which are open to serious doubt, introduce a note of controversy which is not necessary to the main purpose of the book.

The strength lies in the firm grasp the author has of physical systems and the importance he attaches to a proper understanding of physical processes. This point is well illustrated by Chapter 8 which covers a wide range of applications from studies of bouncing springs, through temperature control of a nuclear reactor to linear programming. A part of a sentence in the section describing the optimization of a factory process is of striking relevance. The author writes '. . . if coal is not delivered on time, diode *A* conducts.' The reader is left with the impression that nothing could be more system-orientated.

T. R. H. SIZER

Impedanzuntersuchungen an breitbandigen abgeschirmten Spezialschlitzantennen im dm/cm-Wellenbereich (Impedance Investigations on Wide-band Screened Spiral Slot Aerials in the S.H.F. and E.H.F. Ranges)

By G. Eckart and D. Putzer. 81 pp. Med. 8vo. Westdeutscher Verlag. 1964. Price DM.53

UP to now wide frequency bands with little change of impedance have been reported as the result of experimental and theoretical work on spiral slot aerials for the special case of unlimited space on both sides of the aerial plane. These special space conditions are no longer available for practical applications in communications, e.g. the transmissions of signals to missiles into the wall of which a slot aerial has been incorporated. The missile itself screens the aerial to one side. This poses the problem of the retention of wide-band frequency bands with little change of impedance under these conditions. The report covers measurements of impedances on logarithmic spiral slot aerials screened from one side over approximately three octaves in the s.h.f. and e.h.f. ranges. Cylindrical and conical reflectors are employed for screening. The dimensions of these hollow bodies are varied systematically to study their influence on the input impedance of various spiral slot aerials. The results prove to be very illuminating.

E. R. FRIEDLAENDER

High-Intensity Ultrasonics Industrial Applications

By B. Brown and J. Goodman. 235 pp. Demy 8vo. Iliffe Books Ltd. 1965. Price 55s.

This book is devoted to the applications of high-intensity ultrasonics to industrial processes. It is divided into two main parts, the first dealing with the principles and properties of ultrasonic waves and the various applications such as cleaning, homogenization, welding and fatigue testing.

Dictionnaire Anglais-Francais L'Electrotechnique L'Electronique

By H. Piraux. 362 pp. Med 8vo. Editions Eyrolles. 1965. Price 40.70NF (including postage)

This is the seventh edition of this dictionary originally published in 1952 and now contains over 20 000 terms. Included are not only the correct translations but the idiomatic or slang expressions.

Mechanization and Automation in Radio-Component Manufacture

By N. V. Kashin. 208 pp. Med 8vo. Oldbourne Press. 1965. Price 63s.

This book deals with the mechanization and automation of production processes in the manufacture of radio components in the Soviet Union. Methods in the production of resistors and ceramic, metallized-paper and other types of capacitors are described.

The Transistor

By J. Dosse. 283 pp. Med 8vo. Philips Technical Library. 1965. Price 72s.

The fourth edition of this book includes the latest developments in transistor technology and mention is made of the Planar, POB and MAD transistors.

The section dealing with switching has been revised and new examples have been added. A list of some 300 references to technical publications has also been included.

Worked Examples in Electronics and Telecommunications—Volumes 1 and 2

By B. Holdsworth and Z. Jaworski. 343 pp. Demy 8vo. Iliffe Books Ltd. 1965. Price 25s. Vol. 1, 22s. 6d. Vol. 2

These books, the first in a series of four, have been written specifically to meet the needs of students preparing for the B.Sc. Final examinations in electronics and electrical engineering and of those preparing for the qualifying examinations of the professional bodies. The problems, which have all been carefully selected, are taken from past examination papers—Parts 2 and 3 of the London University Final degree paper and the Graduate Examinations of the Institution of Electrical Engineers. Each chapter deals with one main topic and contains a selection of representative examples which enable the reader to acquire a thorough grasp of the principles involved. Each problem is accompanied by a complete worked solution; not only are all the electrical steps shown, but also the complete mathematical derivation and analysis required to produce the solution.

Recognition: A Study in the Philosophy of Artificial Intelligence

By K. Sayre. 312 pp. Demy 8vo. University of Notre Dame Press. 1965. Price $6.95

Dealing with the increasing field of cybernetics, this book offers to philosophers, social scientists and computer specialists a careful analysis of man's recognitive behaviour and how it may be successfully simulated.

Electronics Data Handbook

By M. Clifford. 158 pp. Demy 8vo. Iliffe Books Ltd. 1965. Price 16s.

The purpose of this handbook, originally published in the U.S.A., is to reduce the task of locating information by providing in one volume the formulae most frequently required.

Theory has been kept to a minimum, since it is the formulae which are the important feature of this book. Where necessary, however, explanatory text is added to clarify the use or derivation of the equations.

Industrial Isotope Techniques

By L. Erwall, H. Forsberg, K. Ljunggren. 338 pp. Med 8vo. Munksgaard Publishers. 1964. Price 120DK

This book, originally published in 1962 and written in Swedish, is based on the work carried out by Danish Isotope Centre, EKONO in Finland, the Norwegian Institute for Atomic Energy and the Isotope Techniques Laboratory in Sweden.

A version written in English has now been published and contains seven chapters, the first three of which deal with the elements of nuclear physics, nuclear chemistry and radiation measuring techniques. The next two chapters review the industrial use of radioactive sources and tracers, based on the authors experience in the Swedish industry. Chapter six deals with radioactive methods of analysis and the last chapter with the aspects of health physics related to the industrial use of radioisotopes.

The Preparation and Production of Technical Handbooks

By H. Keys. 183 pp. Crown 8vo. Issac Pitman. 1965. Price 18s.

The three main objects of this book are to give the would-be technical authors an insight into the profession, to provide a reference book for established authors and finally to indicate to manufacturers of engineering products methods of producing handbooks.

Humidity and Moisture Volume Two—Applications

By A. Wexler. 634 pp. Crown 4to. Reinhold Publishing Corporation. 1965. Price $27.50

Humidity and Moisture presents the proceedings of the First International Symposium on this subject held at Washington in 1963 and sponsored by organizations including the National Bureau of Standards, the American Meteorological Society and the Instrument Society of America.

Volume 2 deals with measurements which are unique or special to the various fields of disciplines; studies and investigation in which humidity or moisture is a critical factor.

ELECTRONIC EQUIPMENT

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

(Voir page 487 pour la traduction en français; Deutsche Übersetzung Seite 494)

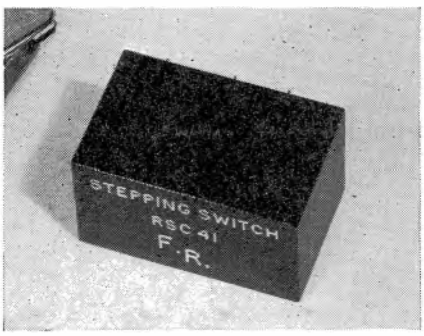
STEPPING SWITCH MODULE

Flight Refuelling Ltd, Wimborne, Dorset

(Illustrated below)

Flight Refuelling Ltd's Industrial Electronics Division now has available the 'RSC-41' stepping switch which is basically a high-speed uniselector incorporating reed inserts as the switching contacts. It is produced in modular form, each unit being 1⅜in long × 0·9in deep × 1 3/32in high, which offers the circuit designer the maximum degree of flexibility with the minimum complication in interconnexion.

In fact, each module is one position of a stepping switch and as there is no



limit to the number of modules that can be connected together, a stepping switch can be constructed having as many, and exactly the number of, positions as the designer requires. Alternatively, two modules can be used together to form a true bistable.

The pin positions on the base of each module conform to the standard 0·1in spacing for printed circuit boards and are so arranged that interconnexion between modules is reduced to simple strapping.

Although primarily designed to meet a need in stepping switch configurations, the 'RSC-41' module can be used in many switching concepts.

EE 83 751 *for further details*

PRIMERLESS T.R. CELL

Ferranti Ltd, Kings Cross Road, Dundee, Scotland

A new S-band t.r. cell type JF.20 has been developed by Ferranti Ltd. The cell—which has been designed to provide protection for travelling wave tubes or parametric amplifiers used in pulsed radar systems—has a radio-active filling and therefore does not need an external primer supply. Within the frequency limits 2 700Mc/s to 3 100Mc/s the v.s.w.r. is less than 1·25 and the insertion loss is typically 0·3dB.

Where crystal receivers are in use, the JF.20 used in conjunction with another recent development—a varactor diode limiter type 104SSS—provides adequate receiver protection (spike $\simeq$0·07 ergs/pulse) for an insertion loss of less than 1·0dB, without any external power supplies being required.

Receiver protection is provided by the JF.20/104SSS combination even when the radar is not operative, thus dispensing with the need for a mechanical waveguide shutter.

Where parametric amplifiers are in use they should be inserted between the JF.20 t.r. cell and the 104SSS limiter so that the cell provides protection for the amplifier with a minimum insertion loss penalty while the limiter reduces to a safe level the break-through reflected by the amplifier into the crystal mixer.

The JF.20 is rated at 100kW peak, 100W mean, and should be preceded by a ferrite circulator or gas duplexer (e.g., Ferranti 10D series) if operation at greater power levels is intended.

EE 83 752 *for further details*

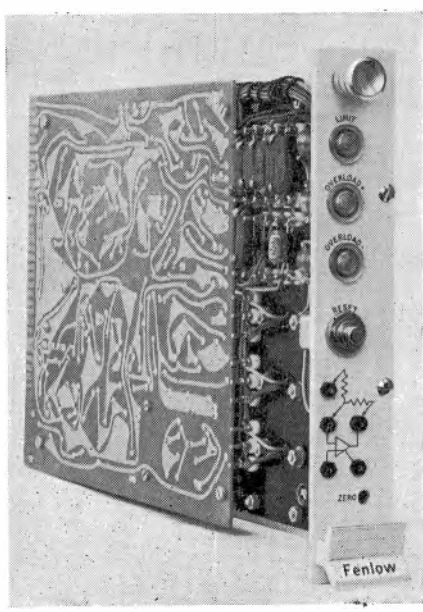
OPERATIONAL AMPLIFIER

Fenlow Electronic Ltd., Springfield Lane, Weybridge, Surrey

(Illustrated below)

The A7 is a photo-resistor stabilized transistor amplifier designed to give an output swing comparable to valve equipment. At the same time it achieves the drift and noise performance of amplifiers operating at much lower levels, thus allowing a considerable increase in computing accuracy.

The amplifier is designed for all the standard analogue and hybrid computer



applications and for instrumentation. Its high output power allows for direct incorporation in quite large servo systems.

Silicon transistors and diodes are used throughout. The amplifier is fully short-circuit proof.

The amplifier provides linear operation up to an output of ±100V at 50mA and full output is available up to 5kc/s. The d.c. gain is greater than 10×10^6 and the gain-bandwidth product is 10Mc/s. The input impedance is 100kΩ, the noise less than 50μV peak-to-peak and the drift less than 1μV/°C.

EE 83 753 *for further details*

CABLE TRUNKING

Insuloid Manufacturing Co. Ltd, Leestone Road, Wythenshawe, Manchester

The Insuloid Manufacturing Co. Ltd has introduced a new range of cable trunking which incorporates many time and cost saving features. This trunking is slotted throughout its length and each slot has a narrow opening which permits cables to be 'led out' at any point. These small openings are so designed that they prevent the cables from breaking free during assembly. Wiring runs can be segregated with snap-in bridging pieces and the cover has a self-retaining 'snap action' fixing. Insuloid cable trunking is grooved in the base to facilitate quick and accurate drilling and has a clean profile for parallel stacking. It is available in four sizes and with three different slot positions. Being manufactured in strong lightweight rigid p.v.c. it is corrosion resistant and requires no attention after it has been installed.

EE 83 754 *for further details*

RELAY INSPECTION TOOLS

P. W. Allen & Co., 253 Liverpool Road, London, N.1

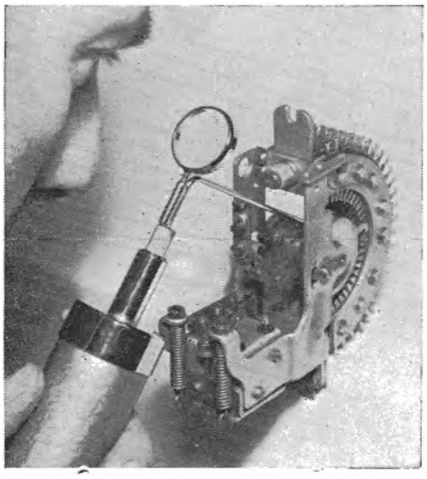
(Illustrated on page 479)

A new set of inspection tools for examining contacts and wipers on telephone step-by-step switchgear has been introduced by P. W. Allen & Co.

This set known as the Allen type TS consists of a battery handle, three light probes and small mirror attachments.

Being able to put a light, and where required a small mirror, right up against a contact even inside multiple rows, the contact face can easily be viewed.

The set is suitable for the inspection of contacts, wipers and mechanisms on all kinds of uniselectors and relays. This selection was made by telephone engineers from the large range of Allen light probes to determine the most economical set to meet all normal inspection



requirements both in the factory and in the exchange.

The small probes are less than $\frac{1}{8}$in in diameter with mirrors of the same width. Magnifiers of either $\times 2$ or $\times 3$ mounted on the probes allow for close viewing with suitable low power magnification.

EE 83 755 *for further details*

DATA LOGGING-RECORDING EQUIPMENT

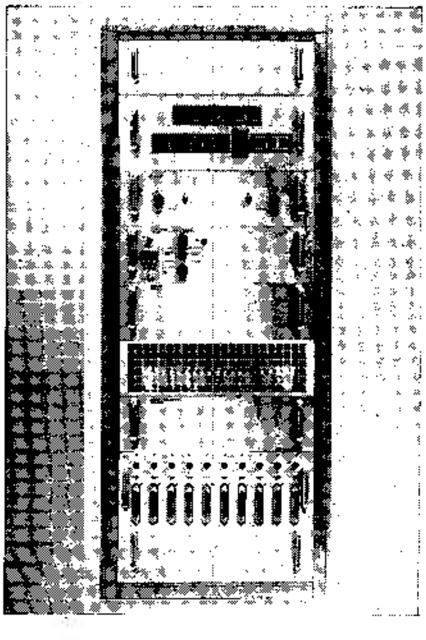
Hawker Siddeley Dynamics Ltd, Whitley Works, Coventry, Warwickshire

(Illustrated below)

In this new data logging/recording system for marine, industrial and scientific applications, full integrated design enables the equipment to be offered as a simple recorder or extended into a sophisticate logging control system with on-line computing facilities.

High or low level analogue signals up to 1 000 in number may be sampled at 20 inputs/sec. Independent guarding on each input and advanced design of the analogue/digital convertor (patent pending) enable the effects of common mode interference to be kept to low limits.

The equipment is designed to operate



continuously within the accuracy limits for at least six months without calibration. Self-calibration occurs each cycle and different and independent limits may be set for each input.

Cycles may be continuously repeated or demanded singly either manually or at intervals under the control of the digital clock.

Printed output information in various formats and an edited punched tape are simultaneously provided, as well as visual displays of input signal values and identities, alarm conditions and time.

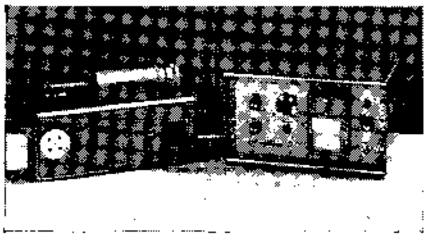
EE 83 756 *for further details*

SURFACE PROFILE MONITOR

G. V. Planer Ltd, Windmill Road, Sunbury-on-Thames, Middlesex

(Illustrated below)

The surface profile monitor is designed for the contour display of surfaces and is suitable for measuring the thickness of thin films within the range 0 to 50 000Å.



The equipment utilizes a displacement transducer to monitor the movement of the stylus, with reference to a datum skid, both of which are made to transverse the surface to be evaluated, two alternative speeds of advance being provided. The resulting measurement is indicated by the recorder pointer or, alternatively, the contour can be recorded on the chart enabling the thickness of films to be determined to an accuracy of better than 100Å.

The equipment is particularly suited for the thickness measurement of vacuum deposited layers as well as for the monitoring of the surface quality of materials.

The equipment comprises two units, the detector unit and the control unit, both units being housed in metal cabinets.

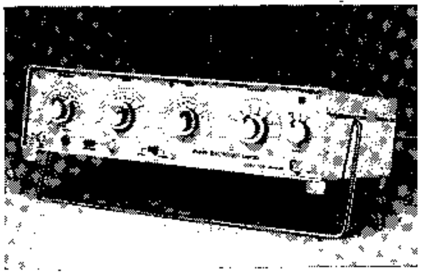
EE 83 757 *for further details*

DOUBLE PULSE GENERATOR

Venner Electronics Ltd, Kingston By-Pass, New Malden, Surrey

(Illustrated above right)

A solid-state double pulse generator, type TSA 628, has been introduced by Venner Electronics Ltd. Using silicon semiconductors throughout, the TSA 628 has a frequency range of from 2·5c/s to 2·5Mc/s used as a single-pulse generator, and from 2·5c/s to 2Mc/s as a double pulse generator. Three methods of operation are provided: free running, triggering from a 2 to 50V external pulse or from contacts, and push-button.



A switched attenuator permits outputs of 20, 10, 5, 2, 1 and 0·5V to be attained, a fine control allowing a minimum setting of 0·2V. Output impedance is 100Ω at 20V and 50Ω elsewhere.

True positive or negative pulses with respect to earth can be selected. Rise and fall times are 10nsec leading edge, 15nsec trailing edge, and pulse width and delay circuits are continuously variable from 100nsec to 10msec. The pre-pulse output is ±5V minimum (unloaded) a.c. coupled, width 100 to 150nsec.

EE 83 758 *for further details*

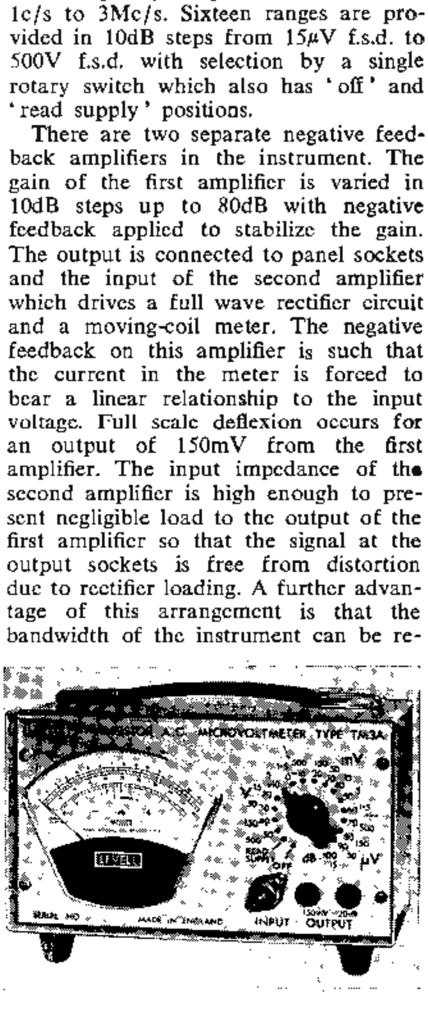
A.C. MICROVOLTMETER

Levell Electronics Ltd, Park Road, High Barnet, Hertfordshire

(Illustrated below)

This portable instrument can be used as an a.c. voltmeter or a.c. amplifier. The frequency response extends from 1c/s to 3Mc/s. Sixteen ranges are provided in 10dB steps from 15μV f.s.d. to 500V f.s.d. with selection by a single rotary switch which also has 'off' and 'read supply' positions.

There are two separate negative feedback amplifiers in the instrument. The gain of the first amplifier is varied in 10dB steps up to 80dB with negative feedback applied to stabilize the gain. The output is connected to panel sockets and the input of the second amplifier which drives a full wave rectifier circuit and a moving-coil meter. The negative feedback on this amplifier is such that the current in the meter is forced to bear a linear relationship to the input voltage. Full scale deflexion occurs for an output of 150mV from the first amplifier. The input impedance of the second amplifier is high enough to present negligible load to the output of the first amplifier so that the signal at the output sockets is free from distortion due to rectifier loading. A further advantage of this arrangement is that the bandwidth of the instrument can be re-



duced by connecting reactive components across the output terminals. This is a useful facility when measuring small signals over a limited frequency band as a reduction of the bandwidth also reduces the input noise level.

The instrument is ideal as an oscilloscope pre-amplifier as it has a low noise level, low microphony, no hum and is independent of a mains supply.

EE 83 759 *for further details*

MATERIALS TESTER

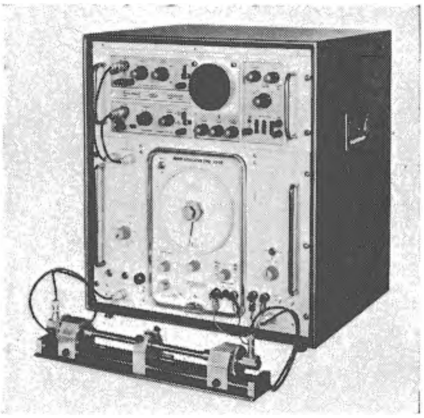
Dawe Instruments Ltd, Western Avenue, Acton, London, W.3

(Illustrated below)

The type 1825 dynamic materials tester is designed primarily for demonstration purposes in educational establishments. It measures Young's modulus, Q and damping capacity of bar specimens excited to resonance.

The equipment comprises an oscillator, step-up transformer and d.c. bias supply, together with a precision test bench, amplifier and oscilloscope. A uniform rod or bar of the test material is clamped at its centre to the test bench, with two small air-gaps (20 to 50 microns) separating its free ends from the electrostatic heads of the test bench. An alternating field applied to the drive head excites the specimen to resonance, the small a.c. signal resulting at the pick-off head being applied via the amplifier to the oscilloscope or some other suitable detector.

With additional equipment, the type 1925 can be used for industrial measurements, but it is not claimed to be accurate enough for all engineering purposes. Young's modulus may be determined with reasonable accuracy from the frequency scale of the oscillator, and with much greater accuracy if an external frequency meter is used. The Q of low-Q specimens may be determined from the oscillator, the Q of medium-Q specimens from an external frequency meter. A different technique, using the logarithmic decrement, is necessary for high-Q specimens, and a variant of the equipment with this additional facility is offered. The oscillator and oscilloscope are self-contained 'modules' and may therefore be used independently for other purposes—an important consideration in college laboratories with limited space or facilities. The standard equipment is fitted with a single-beam oscilloscope, on which Lissajou's figures are obtained. Additional variants are, however, also offered with a double-beam oscilloscope or with no oscilloscope at all if it is intended to use an external oscilloscope or valve-voltmeter as detector.



The standard test bench caters for specimens 150 to 280mm long and 6 to 15mm diameter, a special jig being provided to assist the process of setting up, which takes less than a minute. The oscillator used is a high-power 6W model giving an output from 20c/s to 20kc/s. The electronic equipment measures 20½ × 17 × 23in high.

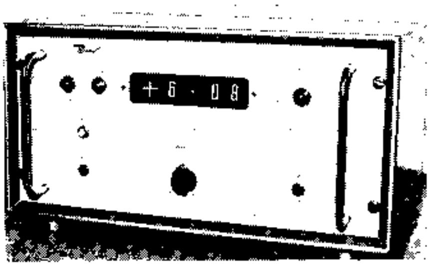
EE 83 760 *for further details*

PERCENTAGE RATIO METER

Bendix Electronics Ltd, High Church Street, New Basford, Nottingham

(Illustrated below)

The percentage ratio meter comprises a series of transistorized instruments



using digital counting techniques to measure and display the percentage difference between two processes, i.e., roller speeds, shaft speeds, etc. The instrument is particularly useful for industrial calendering processes where percentage stretch and shrink need to be measured. Differences at 0·01 per cent can readily be measured.

Consisting basically of two counters, input is applied to both simultaneously, with one counter arranged to inhibit both inputs on completion of 1 000 counts (or whatever count base is chosen). The inputs are connected so that the second counter will be equal to or greater than 1 000 (dependent on the count base chosen). The preferred input transducer is the Rotapulse 191A or the Pulse Generator 491A. Other input circuits can be fitted to the counter to cater for other types of pulse transducers.

If the measured ratio is outside the limits 0 to ±9·9(9) per cent one of two signs will be initiated − for underscale, + for overscale. In this condition the numeral display is inhibited.

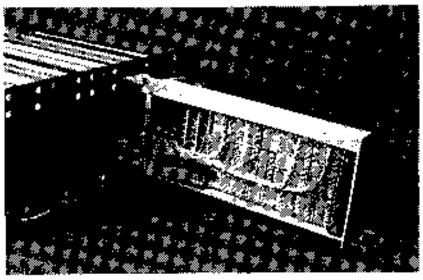
EE 83 761 *for further details*

CIRCUIT MODULES

Morecombe Electrical Equipment Co. Ltd, Westgate Works, Morecombe, Lancashire

(Illustrated above right)

Several new items have been added to



the 'Meecostattic' range of transistorized logic, input, output and circuit modules, designed for industrial control. These new modules consist of an RC timer, a binary to decade de-coding matrix, a two circuit equivalent gate, ramp generator, reversible binary counter and adjustable on level and off level trigger modules.

These additions now bring the total number of modules in the range to 35, all of which are electrically and mechanically compatible one with another.

To ensure satisfactory assembly of these modules a 19in standard rack mounting chassis system has been evolved, a section of which is shown in the photograph. Each deck can contain up to five 3in wide chassis or a proportionately smaller number of 6in or combined 6in and 3in units. Each chassis can house up to sixty ½in modules and is designed for plugging into the deck frame-work. Construction is mainly in aluminium having a high quality instrument finish, the front surfaces which would be externally visible are black anodized and have mounted in them chromium plated handles for easy withdrawal of the chassis from the decks. The photograph illustrates how chassis may be plugged into an 'inter-connect' unit and thus be made live for testing purposes, proud of the front face of the main equipment.

At the front of the chassis can be seen a test probe. This is connected to a small indicating lamp so that a check may easily be made of the 'on' or 'off' conditions in any of the modules in that section of the unit.

Another addition of importance to the range has been a series of power racks, these being in ½A, 1A and 2A sizes. These packs give a −24V regulated output, stabilization ratio being greater than 5 000 : 1 and output d.c. resistance less than ·002Ω, the output impedance being less than 0·1Ω at 100kc/s. Also in these power packs is a +24V supply having a common zero volt connexion. This positive supply is principally used for biasing purposes and employs simple Zener regulation.

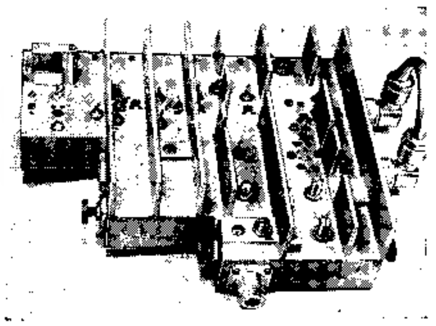
EE 83 762 *for further details*

MICROWAVE EQUIPMENT

G.E.C. (Telecommunications) Ltd, Telephone Works, Coventry

(Illustrated on page 481)

A new range of completely 'semi-



conductored' microwave radio equipment is now being produced by G.E.C. (Telecommunications) Ltd. Previously, 'semiconductored' microwave equipments have used thermionic devices in some part of the circuit, usually travelling wave tubes, klystrons or microwave triodes in the power output stage. Now G.E.C. has successfully produced equipment using semiconductors exclusively. The two main advantages are increased reliability and a substantial reduction in power supply requirements, even over equipment that uses thermionic devices in only one stage. Both these advantages are particularly important in countries where access to remote stations is difficult and both power supplies and maintenance have to be reduced to a minimum.

Of the two equipments that G.E.C. has currently in production, one is suitable for main trunk routes and will carry 960 speech circuits, or a television channel plus associated sound, in the frequency bands 1 700 to 1 900Mc/s and 1 900 to 2 300Mc/s. The transmitter power output of this equipment is 2W obtained solely from semiconductors and the receiver incorporates a low noise input stage, employing a tunnel diode amplifier, having a noise factor of 5½dB. The broadband equipment is associated with a completely semiconductored auxiliary radio channel and multi-r.f. channel protection switching equipment.

The second equipment operates in the frequency band 7425 to 7725Mc/s and, having a capacity of 300 speech channels, is suitable for providing high reliability circuits for the smaller telephone Administrations, and for spurs from high-capacity broadband routes. A twin path bothway repeater is packaged on one rack complete with switching, order wire and remote alarm equipment. A useful feature in this equipment enables any portion of the baseband to be dropped and reinserted at any repeater station without demodulation of through circuits.

EE 83 763 *for further details*

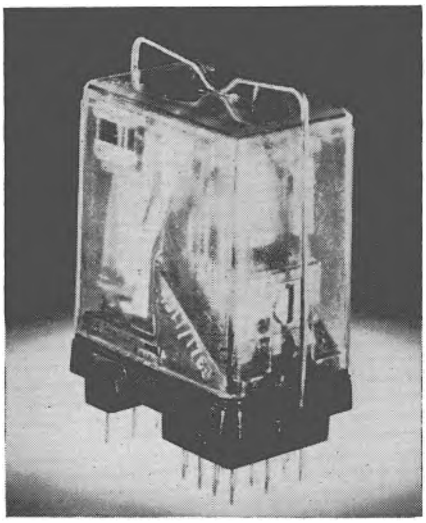
MINIATURE LATCHING RELAY

Distributed by: B & R Relays Ltd, Temple Fields, Harlow, Essex

(Illustrated above right)

B & R Relays Ltd., distributors of Siemens & Halske miniature relays in the U.K., has now introduced to the British market the new Siemens type P magnetic latching relay. The type P is available in two sizes. The smaller version has two changeover light duty contacts. The larger has four changeover, six normally open and six normally closed contacts or two changeover heavy duty contacts. Contacts are available in either silver or gold alloy for most versions. The type P cradle relay (reference number V23003) is designed to operate from a pulse.

The operation is such that when the relay is fitted with a single coil, it can be energized or pulsed—and it will then operate and stay in that state until the



coil is excited in the opposite direction.

Alternatively the relay can be fitted with two coils when the relay can be energized by exciting one coil and de-energized by exciting the other.

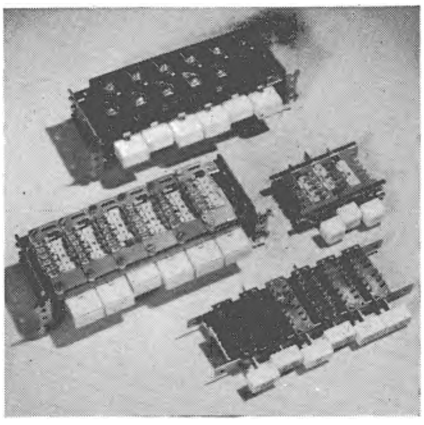
EE 83 764 *for further details*

PUSH-BUTTON SWITCHES

Standard Telephone & Cables Ltd, Electro-Mechanical Division, West Road, Harlow, Essex

(Illustrated below)

The range of push-button assemblies made by Mayr, an STC Associate Company, is now available from STC's Electro-Mechanical Division. These assemblies offer the control engineer a versa-



tility of switching arrangements which is unusual for this type of unit. The flexibility of the system gives multiple combinations of button hold and button release conditions. These include independent and mutual release assemblies with mutual release in groups, if required, or with common release.

As well as varying multiples of hold and release buttons, each, in turn, being able to perform a variety of switching functions, the Mayr assemblies can contain 'pulse' buttons in the same unit. A large number of electrical and mechanical combinations is possible. Switch assemblies of differing construction (high and low power) can be supplied on the same switch-button chassis. Each button can be made to perform a specified number of make, break or changeover operations.

Most of the ranges are available with banks of 12 buttons. The construction allows for very close vertical and horizontal stacking of the banks. The buttons have removable transparent covers for the insertion of identification colour signals or written characters.

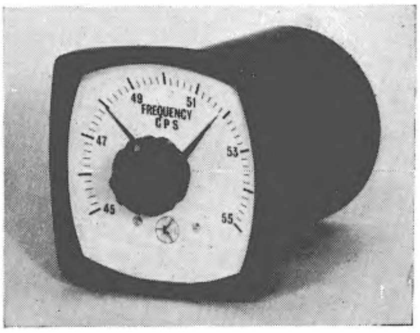
EE 83 765 *for further details*

FREQUENCY CONTROLLER

Kent Precision Electronics Ltd, Vale Road, Tonbridge, Kent

(Illustrated below)

Kent Precision Electronics Ltd has recently developed a new solid state circuit which is fitted within the mechanical assembly of their standard solid state calibrated control instrument for



control, protection or warning in frequency monitoring applications.

The system consists basically of a frequency to d.c. convertor which feeds a d.c. signal proportional to the frequency, accurately over a small range, say 45 to 55c/s. The input frequency is initially converted to a pulsed output of the same frequency which is then fed to a linear integrator, the output of which is fed to the input of a standard KPE solid state calibrated control instrument. The extra frequency to d.c. converter circuit is built as an extra printed circuit card within a standard 'Elite' type instrument case and derives its power supply from that of the rest of the instrument. The auxiliary supply to the instrument is normally the mains a.c. voltage and the instrument can be used to monitor the frequency of its own auxiliary supply, if required. Alternatively it can be used to monitor a secondary input.

Single or dual instruments are available as standard, the dual instrument having concentric control knobs. In each case an internal relay with changeover contacts rated at 5A 240V a.c. is made to operate at the set point. Alternatively, the output may consist of thyristor proportional or on/off control elements.

EE 83 766 *for further details*

DEMOUNTABLE D.C. GAS LASER

Scientifica, 148 St. Dunstan's Avenue, Acton, London, W.3

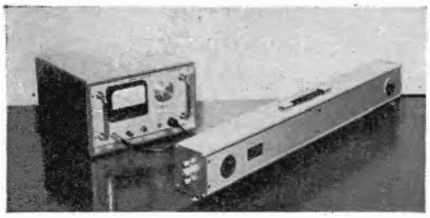
(Illustrated below)

The development of the Scientifica d.c. gas laser is the result of efforts to produce a demountable instrument which will have the precision and stability required for research, and yet which will be sufficiently rugged for general laboratory use. The design allows for the laser to be operated either as a free standing apparatus able to be used in any orientation, or as an optical bench assembly where the resonator mirrors can remain mounted in the laser, or be held by separate optical bench supports. This versatility of arrangement makes it possible for experiments to be performed with different mirror spacings, so enabling a detailed study of stable and unstable resonator configurations to be made, and for standard optical components to be placed inside the resonator cavity.

Further features of the instrument are that it incorporates 17-layer hard coated mirrors which may be cleaned with normal lens tissue, or washed in alcohol, and the specially designed d.c. gas laser tube which has exceptionally long life. The output of the laser tube is a continuous c.w., plane polarized in the red and infra-red region of the spectrum, with typical outputs of 3mW at 6 328Å and 10mW at 11 530Å.

A single cable connects the laser to a d.c. drive unit, where a control is provided for adjusting the light output, the level of excitation being displayed on a 50mA moving-coil meter. The unit is designed for operation at either 110V or 230V a.c. and provides a low ripple d.c. output which produces less than 1 per cent modulation of the laser beam. Both the laser and the drive unit incorporate conservatively rated components of the highest quality which are guaranteed unconditionally to be free from defect for one year.

A manual is supplied with each laser giving details of experiments that may be performed using this new source of coherent light. These include both physical and geometrical experiments such as Fraunhofer diffraction, Abbé theory, Doppler shift due to a moving mirror, collimation of the laser beam giving a spread of a few inches per mile, and so on.



With the high power monochromatic beam provided by this instrument, the procedure for setting up any interference experiment is considerably simplified. Furthermore, the interference patterns which are obtained are several orders of magnitude greater in intensity than those obtained by previous methods, and are therefore the more striking. This enables such experiments to be performed in non-darkened rooms, and for large audiences to be able to view the interference effects.

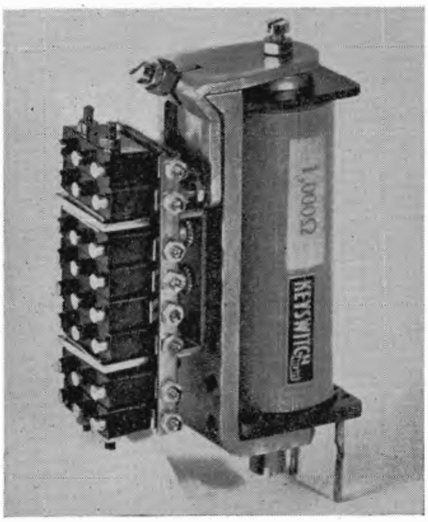
EE 83 767 *for further details*

MICROSWITCH RELAYS

Keyswitch Relays Ltd, 120-132 Cricklewood Lane, London, N.W.2

(Illustrated below)

A range of microswitch relays, based on the GPO 3 000 and 600 standard types, and particularly designed for temperature control and sensing circuit uses, has been introduced by Keyswitch Relays Ltd. Features include high contact



ratings, low actuating forces, a high (snap action) switching rate. Operating contacts are fully enclosed for suitability in hazardous atmospheres, and currents of up to 20A at 480V a.c. can be switched. A power transfer ratio of 1 : 100 000 can be achieved.

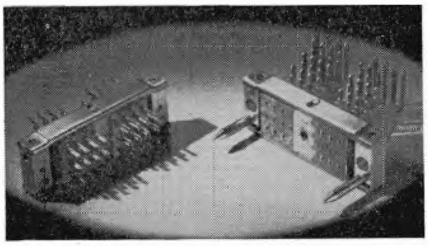
The three basic types available are GPO 3 000 (multiple and single switch), and a GPO 600 type single switch. Working voltage (3 000 types) is from less than 1V d.c. to 300V d.c., and up to 440 V a.c. with added components; on the 600 type it is from less than 1V d.c. to 175V d.c., and up to 440V a.c.

D.C. resistance is from less than 1Ω to 50kΩ for the 3 000 types, and from less than 1Ω to 10kΩ for the GPO 600 types.

Available variations on the standard range include double armatures on the 3 000 types, to allow for independent operation of the microswitches.

Adjustable operating bias can be fitted to the 3 000 single switch and 600 types, and manual-reset latching to both the 3 000 units. Coils can be tropicalized if required, and either screw or solder switch connexions can be supplied.

EE 83 768 *for further details*



MULTIPLE PLUGS AND SOCKETS

Oxley Developments Ltd, Priory Park, Ulverston, Lancashire

(Illustrated above)

Oxley Developments Co. Ltd has recently introduced a novel type of chassis mounting multiple plug and socket. The connectors are provided with 24 independent gold plated contacts and the location method makes them very suitable for equipment drawers in rack panels and consoles. The connectors are provided with a built in mechanical locking device (screw held) which simultaneously applies pressure to the contacts thus preserving low contact resistance over long periods under arduous climatic conditions.

EE 83 769 *for further details*

DRILLS FOR PRINTED CIRCUITS

Drill Service (Horley) Ltd, 89 Albert Road, Horley, Surrey

Nearly all printed circuit boards are made from a material which is so abrasive that normal HSS drills wear very rapidly. Not only do the points blunt very quickly, but also the cylindrical drill land wears causing undersize holes.

Starbide solid tungsten carbide circuit board drills have been developed specifically for printed board drilling. The use of tungsten carbide completely eliminates land wear and gives a greatly extended point life. Sixty thousand holes between re-grinds would appear to be a normal accomplishment.

It is also essential to produce burr free holes in circuit boards. This requires a high rake angle and high spiral. Also to remove the swarf at the very high feed rates employed, a very thin web is necessary.

Starbide printed circuit board drills have a very high spiral and thin web with flutes ground to a high micro finish. The diameter tolerances are to ·0003in. This is necessary if precision holes to ·0005in tolerances are to be produced in quantity. The overall length is the same for all sizes at 1½in., with a flute length of ½in.

EE 83 770 *for further details*

VIBRATION MEASURING EQUIPMENT

Associated Engineering Ltd, Cawston, Rugby, Warwickshire

(Illustrated above)

Associated Engineering Ltd has made additions to its range of displacement and vibration measuring equipment. The additions consist of three items of electronic equipment designed for use with its subminiature inductive displacement transducers. The equipment consists of a stable oscillator with five switched frequency ranges between 50kc/s and 150kc/s, and an amplifier unit which when driven by the oscillator provides six very low impedance outputs of 20V peak-to-peak. These two units are capable of energizing six channels of measuring equipment using inductive transducers, each channel working in conjunction with a detector/filter unit which provides a matched load for the transducer, suitable detection facilities, and a specially designed filter capable of separating the modulated signal from the carrier.

With this combination a sensitivity of 10mV per thousandth of an inch movement is obtainable and movements of frequencies up to 25kc/s may be indicated with no measurable distortion.

The photograph shows examples of each of the three units mentioned and also four inductive transducers of the two sizes currently in production.

EE 83 771 *for further details*

DIGITAL WORD GENERATOR

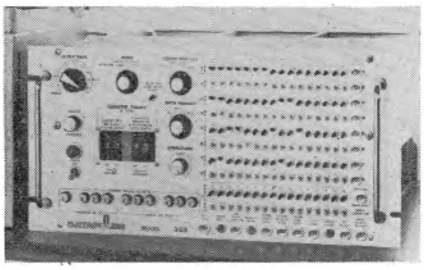
Digital Measurements Ltd, Salisbury Grove, Mychett, Aldershot, Hampshire

(Illustrated below)

The Datapulse model 208 word generator provides flexible control of data format and content for simulation of high-speed, multi-channel digital data.

The unit features 5c/s to 5Mc/s clock rate, 8 bit parallel characters, 1 to 16 serial characters per data block plus character or block repeats, 6V output into 50Ω, and RZ or NRZ formats.

In parallel mode, parallel characters with variable block length and control of blocks-per-frame permit simulation of paper or magnetic tape data. Digitally



controlled repeat is available for generation of longer effective patterns.

In serial mode, eight different sixteen-bit serial words may be commutated on a single line with digital control of words-per-frame for simulation of p.c.m. data or multiplexing systems. Read-out lights provide continuous indication of program status. 50Ω output buffers drive interconnecting cables directly with no need for additional buffering.

Frames may be repeated continuously, recycled manually, or repeated by external signal command. A front panel push-button permits reset to frame-start condition. Provision is made for program stop under external signal control.

Sync output is positive 6V pulse, 50nsec wide with 100Ω source impedance and may be selected to occur coincident with any one or combination of characters. Frame start sync occurs coincident with the first bit or character in each frame. Frame and sync occurs coincident with last bit or character.

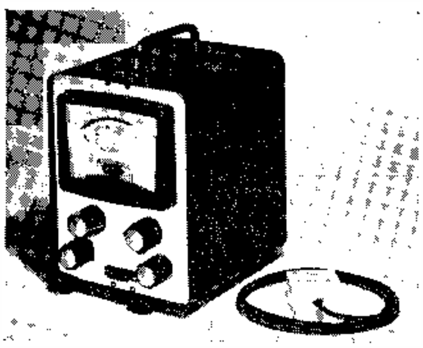
EE 83 772 *for further details*

PORTABLE GAUSSMETER

Preformations Ltd, Cheney Manor, Swindon, Wiltshire

(Illustrated below)

Preformations Ltd, a Plessey sub-



sidiary, has developed a new portable gaussmeter providing linear static measurements of magnetic field strengths over the wide range of 5 000 to 105 000 gauss. Preformations Ltd, whose products are sold under the 'Magloy' trade name, is one of the largest British manufacturers of permanent magnets.

The instrument is believed to be the first to make use of the relatively little known magneto-strictive effect and uses a probe consisting of a small plate of either indium antimonide or indium arsenide. This plate changes its resistance when placed in a magnetic field normal to its surface, the change being monitored by a transistorized ohmmeter, which feeds a d.c. amplifier.

Above 5kG the change in resistance becomes a sensibly linear function of flux density, and this has been confirmed in fields up to 105kG.

In order to dispense with external calibrating fields, internal calibration is provided. The instrument can also be used in conjunction with an oscilloscope, pen recorder or telemetric device.

As the probe can be as thin as 0·15mm it can be used even in very small gaps and the indication is always positive, irrespective of field polarity. An advantage of this type of probe is that it can be used to make detailed surveys of magnetic fields for linearity and homogeneity.

The instrument, originally developed for research into ferromagnetic resonance at microwave frequencies, can also be used in superconducting solenoids and in apparatus for research in proton resonance, mass spectrometry and plasma physics. By linking the instrument with an oscilloscope it is possible to measure very high pulsed currents, for example, those used in laser pump supplies and pulse magnetizers, and display the waveforms produced.

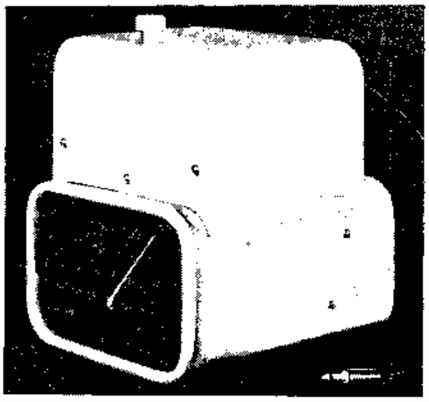
EE 83 773 *for further details*

GRATICULE PROJECTOR

Marconi Instruments Ltd, St. Albans, Hertfordshire

(Illustrated below)

The graticule projector type TM 8098 employs a semi-reflecting mirror to permit normal viewing of oscilloscope displays while allowing the projection of various kinds of reference data into the same plane as the display itself with complete freedom from parallax. Typical uses for this versatile accessory are: the projection of standard graticules for high precision work, special graticules for pulse-and-bar checking, standard waveforms and limit lines for production checking. The data to be projected may be produced by photographic reduction from master drawings by direct photography of reference waveforms or by sketching on 'Scribecoat' sheets. The compact assembly clips simply and quickly on to the rectangular bezel now fitted as standard on all Marconi oscillo-



scopes. Power for the projector lighting is obtained from the jack socket fitted to all current instruments which transfers the supply from that provided for normal graticule illumination. The accessory is also available in another version, the TM 8097, which is suitable for instruments having $5\frac{3}{4}$ in. diameter bezels; this version obtains its 6·3V supply via a MODAC connector which mates with the cathode-follower probe supply outlet on earlier TF 2200s. For other oscilloscopes a separate 6·3V supply is necessary.

EE 83 774 *for further details*

Short News Items

'Engineering Aspects of Computer Technology' is a 13-day course, commencing 11 October 1965 and repeated from 1 November 1965, for graduates who are entering, or who have recently entered, the computer industry.

A special feature of the course is that lecturers from all the major computer companies are taking part. In addition, specialists from R.R.E. and N.P.L. have been invited.

The course is intended for engineering, physics and mathematics graduates and will consist of lectures, discussions and practical sessions as well as a series of introductory lectures on programming. Among the topics covered are cryogenic stores and computers, features of scientific and business machines, integrated circuits and the reliability of components.

Further details are available from, The Principal, College of Technology, Broadway, Letchworth, Hertfordshire.

The Council of the International Federation of Information Processing has appointed the British Computer Society, whose headquarters are in London, as its first permanent secretariat. Mr. John Mackarness, executive secretary of the British Computer Society, will act as administrative secretary to the IFIP Council.

The Council meeting at which this decision was made was held at the same time as the third IFIP Congress and Exhibition, which took place at the New York Hilton Hotel between 24 and 30 May. At the close of the conference the IFIP Council announced that the next Congress and Exhibition will be held in Edinburgh from 5 to 10 August 1968.

The Digital Systems Department of Ferranti Ltd has received a contract from the Ministry of Aviation to supply a Hermes computer and associated peripheral equipment to the Southern Air Traffic Control Centre (SATCC) at West Drayton.

The equipment—known as MINICAP—is scheduled for delivery in May 1966 and will be used as an interim system to automate the major part of the flight progress strip production and distribution system. It will also produce and transmit flight progress strips to be used by Ground Movement Planning at Heathrow airport. Transmission will be through a data link.

The flight progress strips are used by controllers to give information such as aircraft identification, destination, estimated time of arrival etc. The basic information is obtained from the pilots' flight plan and is at present written on the strips by hand.

When the Hermes is installed, the flight plan information will be fed into the computer with the variables such as take-off direction and windspeed. After processing, the flight progress strips will be printed out ready for insertion into the flight progress boards.

The Ferranti Hermes to be used as the central processor is a small, very fast and flexible, transistorized computer, which has been developed for real-time on-line applications.

Redifon Ltd has supplied a £55 000 comprehensive h.f. communications system to Malaysia. The system consists of transmitting, receiving and ancillary equipment at each of three stations. Each station has separate transmitting, receiving and operating buildings linked by v.h.f. and/or underground cables and is provided with full remote control facilities for all transmit functions.

The system, operating on r.t., r.t.t. and on c.w. is an i.s.b. duplex network using Redifon G.426B 1·5kW transmitters and R.403 receivers as main equipments. R.T.T. two-tone keying with automatic error detection and r.t. secrecy services are provided by ancillary units. Complete standby equipment is provided at each station.

S.E. Laboratories (Engineering) Ltd is to provide Hindustan Aeronautics Ltd, Bangalore, India, with a £41 000 flight flutter test instrumentation system in connexion with local aircraft production.

S.E. Laboratories will be responsible for the complete system engineering, including the flight instrumentation as well as the ground read-out equipment for flight flutter measurement.

The S.E.L. equipment used will include accelerometers, oscillators, demodulators, integrator units, u.v. recorders and galvanometers.

Other manufacturers' equipment will include voltage controlled oscillators, discriminators, magnetic tape record/replay units for both airborne and ground use, together with XY plotters and resolvers.

Decca Radar Ltd has entered into the field of medical electronics and automation. The Company exhibited a new and advanced automated transfusion unit at the Hospital Equipment Exhibition at Olympia. Based on the Decca Mastercount photo-electric technique, this new unit is capable of administering and controlling a transfusion without supervision. Decca is already developing its medical electronics service in a number of countries with equipment designed to improve the accuracy and efficiency of making and dispensing tablets, pills and capsules.

The Decca Mastercount automated transfusion unit is a monitoring and control system for use with all types of administration set for blood and other liquids. The unit, which is safe and reliable, eliminates the need for constant supervision during a transfusion and itself makes the adjustments usually required to maintain a constant drip rate.

The complete equipment is mounted on a light, mobile stand. It can be quickly adjusted to operate an administration set at any drip rate between 4 and 60 drips/min, and will thereafter maintain the chosen rate regardless of any external changes affecting the system. The Mastercount enables transfusions to be carried out simply, safely and with the minimum of supervision by nursing staff; when required it can be operated manually in the normal way.

Industrial and Trade Fairs Ltd. regret to announce that they have been compelled to cancel The '65 Show mainly because of the decision of many of the largest companies in the British radio industry not to take part.

In these circumstances, it would not be possible to hold a radio and television show which would fulfil its object of presenting the industry in all its aspects both to radio and television dealers at home and overseas and to the general public.

Discussions are now taking place between Industrial and Trade Fairs Ltd and the industry concerning the form which an International Radio and Television Show might take in 1966.

G.E.C. (Telecommunications) Ltd has been awarded further contracts worth over £100 000 to supply automatic telephone exchange equipment to Aden.

The orders were awarded by the Ministry of Posts and Communications of the Federation of Southern Arabia. They include the supply of equipment to extend the new Crater exchange which was successfully put into service last month, an extension to Little Aden exchange and equipment for a new exchange at Al Mansoura. This replaces a rural automatic exchange at Sheikh Othman originally supplied by G.E.C. twenty-five years ago. In addition, equipment will be supplied to convert the Aden Central exchange from 4 to 5-digit working.

The equipment will be manufactured

at the Company's Telephone Works in Coventry and G.E.C. engineers will supervise the installation which is scheduled to begin in the near future. The installations will be complete by the end of 1966 bringing the total number of exchange lines in Aden to over 8 000, all supplied by G.E.C.

The largest closed circuit television system in Europe and probably in the world, has been installed in the Ministry of Defence building in Whitehall. The 'Whitehall Television Network' has been designed to use 32 camera channels feeding 70 different locations, with up to 100 screens. All these are remotely controlled from one main control room in the heart of the Ministry building.

The network was designed and installed to Ministry requirements by Peto Scott Electrical Instruments Ltd, using Hi Q cameras and studio monitors. Both mobile and static cameras are used in the system, which has been designed to speed up and clarify the transmission of information within the Ministry.

The system, although only in operation since the beginning of March and still not at full capacity, has already attracted the attention of other Ministries, and officials from SHAPE HQ in Paris, who are to pay further visits to London to study the system with a view to adapting it for their own purposes.

Corporate members of the Institution of Electrical Engineers—Members and Associate Members—will become Fellows and Members respectively if a group of Council recommendations are endorsed at a Special General Meeting and approved by the Privy Council. The Council's recommendations also propose that the existing Graduate and Associate classes should be combined to form a new non-corporate class of Associate Members.

These proposals follow others announced last year that the recruitment to the class of Associates should end and that a new class of non-corporate membership should be created for persons interested only in the learned society activities of the I.E.E. such as doctors concerned with medical electronics, architects, teachers and technician engineers, who do not qualify for other classes of membership.

The proposal to use the titles 'Fellow' and 'Member' for the corporate membership has the support of the Board of the Engineering Institutions Joint Council.

The Ferranti Inertial Systems Department has received an order from Bolkow-Entwicklungen KG of Munich on behalf of Eldo for the supply of attitude reference systems and ground test equipment for the European Satellite Launch Vehicle.

The system provides guidance and control information derived from a modified Ferranti light-weight inertial platform.

The present contract is worth almost £250 000 and there is an option to purchase further systems for a similar sum.

Close collaboration has existed for some time between Ferranti Ltd and SAGEM (Societe d'Applications Generales d'Electricite et de Mecanique), in France, and the latter will be supplying all the gyroscopes and accelerometers for incorporation in the Ferranti system.

The platform carries three orthogonally-mounted, single-axis, rate-integrating gyros and three force-feedback accelerometers. It is aligned so that two accelerometers have their input axes lying in the plane normal to the vertical and pointing north and east (north/south and east/west navigation channels). The third accelerometer has its input axis aligned with the vertical. The platform is Schuler-tuned (84 minute period) in the two horizontal channels and is processed in azimuth to maintain its orientation as it is moved over the surface of the earth.

The Science Research Council (SRC) announces that efforts are being made to bring back men with considerable experience of industry to academic life for further postgraduate training before returning to industry.

It is felt that there are many men with academic qualifications in British industry today who would welcome the opportunity of furthering their studies, perhaps adding to their present qualifications and bringing themselves up to date with new techniques with a view to increasing their potential to industry.

The basic value of a research or advanced course studentship is £450 p.a. free of tax, plus approved fees, but this value is under review. Allowances may be paid for adult dependants and children and to older students. To encourage men with postgraduate experience in industry, the SRC pays a tax-free allowance of up to £200 per year to students who have this experience. Students are encouraged to supplement their award by undertaking paid demonstrating and teaching within certain limits. A married man with children and with industrial experience may well receive around £1 000 p.a. tax free.

This year 100 studentships are on immediate offer by the SCR between now and October 1965. Applications are already under consideration and men who feel they have the appropriate qualifications, who have been working in industry for two years or longer, are invited to contact the Postgraduate Training Awards section of the Council at State House, High Holborn, London, W.C.1, Mr. Crabtree, CHAncery 1262, Extension 59.

'Precision Sound Level Meters' is a new publication by the International Electrotechnical Commission and is known as IEC Publication 179. It applies to sound level meters for high precision apparatus for laboratory use, or for accurate measurements in which stable, high fidelity and high quality apparatus is required. A sound level meter is, generally, a combination of a microphone, an amplifier, certain weighting networks, an attenuator and an indicating instrument having certain dynamic characteristics. The sound level meter covers the frequency range 10c/s to 20kc/s.

The object of Publication 179 is to specify the characteristics of an apparatus for measuring accurately certain weighted sound pressure levels. The weighting applied to each sinusoidal component of the sound pressure is given as a function of frequency by three standard reference curves called A, B and C.

The instrument described in this Recommendation represents a practical combination of characteristics that will achieve a high degree of stability and accuracy. It is the best compromise at present, the manner of use determining, in large part, the validity of any measurement made with an instrument that meets this specification.

Recommendations applying to sound level meters for general purposes .re given in IEC Publication 123 'Recommendations for sound level meters.'

Copies of IEC Publication 179 may be obtained from the BSI Sales Branch, 2 Park Street, London, W.1. Price £1 5s. each.

The present schedule of 625-line experimental NTSC colour transmissions from Crystal Palace, Channel 33, has been temporarily replaced by a further series of experimental transmissions using the PAL system.

Initially these experimental colour transmissions are being radiated during the BBC-2 trade test periods at the following times.

Monday-Friday inclusive
12.00 to 13.00
15.00 to 16.00

Each one-hour transmission consists of 15min trade test card in black and white, 10min colour bars and 35min colour slides.

The Royal Radar Establishment, (R.R.E.), at Malvern, has placed an order for a Marconi MYRIAD computer for use on advanced radar research projects. This is the first order for this new desk-size computer which uses microelectronic circuits, developed by The Marconi Co Ltd, to provide far greater reliability than has been possible with previous machines. MYRIAD is a very fast, on-line, real-time computer, designed primarily for air traffic control and defence applications, although its high capacity and flexibility make it suitable for use in many other fields.

The MYRIAD computer for R.R.E. will have storage capacity for 16 384 words of information, and will be supplied with full test equipment and spares. A Marconi system analyst will work for R.R.E. for a year, to assist with the programming of the machine, and with the preparation of other 'software' for a variety of tasks.

The computer is due to be delivered in March 1966. A high level language compiler (Algol) will be supplied, together with an extensive library of subroutines, editing, test and diagnostic (fault finding) programmes. A MYRIAD User Code will also be provided for programming on-line applications.

An annual 'Appleton Lecture' is to be inaugurated by the Council of the Institution of Electrical Engineers to commemorate the life and work of Sir Edward Appleton, the eminent physicist and Nobel Prize winner, who died 21 April 1965.

Sir Edward was an Honorary Member of the Institution and was awarded the I.E.E. Faraday Medal in 1946 for 'the conspicuous services rendered by him in the advancement of electrical science, particularly in the field of radio propagation.'

Further details and the date of the first lecture will be announced later.

Marconi television cameras and transmitters will be exported to America in early September for a new v.h.f. television station to be built by Television Chicago, a joint venture controlled by Field Communications Corporation. Field Communications Corporation is a wholly owned subsidiary of Field Enterprises, Inc., whose Newspaper Division publishes the Chicago Sun-Times and the Chicago Daily News.

This is the first time that British television transmitters have ever been sold in the United States, and the contract also includes the first export sale for the new Mark V camera, which was introduced recently.

Morganite Resistors Ltd are now selling a quarter of a million resistors a week to Hong Kong for use in low-cost transistor radios.

The resistors are of greater accuracy than anything that Japan can make within the price range, with a tolerance of better than 10 per cent.

The same company has stepped up its export of resistors to the United States. Sales worth more than a million dollars are expected this year including a major contract with government departments calling for a very high accuracy.

Test equipment for U.K.3 and other earth satellites has been ordered for the Royal Aircraft Establishment from The M.E.L. Equipment Co. Ltd. This follows a design study by M.E.L. for a data sensing and extracting system for a satellite environmental testing facility.

The M.E.L. work is largely connected with temperature sensors and ancilliaries for operation between -180 and $+180°$ in the testing chamber. The chamber is large enough to accommodate full-scale satellites and will produce low pressures and extremes of temperature similar to those encountered in terrestrial orbit.

The satellite under test can be moved in two directions at right-angles to each other in the chamber and rotated continuously for three turns about its own axis. Considerable attention has therefore been given to devising a flexible joint for connecting a large number of sensors on the satellite to the external equipment. The joint has to be of small dimensions and has to operate reliably in a near vacuum. Initially, an R.A.E. system of separate coiled conductors will be used, but among the most likely final solutions to the problem is a specially impregnated glass fibre strip coiled like a clock spring and printed with parallel copper conductors.

Developments in the terminology relating to semiconductors and semiconductor devices have made it necessary to supplement the British Standard 'Glossary of terms used in telecommunication (including radio) and electronics', B.S. 204, although this was fully revised in 1960.

In Supplement No. 2 'Semiconductors and semiconductor devices', now published, it has been found possible to include very many terms which have been agreed internationally since 1960.

The Supplement supersedes subsection 22 of B.S. 204.

Copies of Supplement No. 2 to B.S. 204 may be obtained from the BSI Sales Branch, 2 Park Street, London, W.1. Price 5/- each. (Postage will be charged extra to non-subscribers.)

The BBC's television and v.h.f. sound relay station at Machynlleth, Montgomeryshire, will be brought into service on Monday 28 June. It will transmit BBC Wales Television on Channel 5 with horizontal polarization. The three sound services will be transmitted on v.h.f. with horizontal polarization on the following frequencies:

Welsh Home Service	93·8Mc/s
Light Programme	89·4Mc/s
Third Network	91·6Mc/s

Membership of the Institution of Electrical Engineers now exceeds 54 000 according to the Annual Report of the Council for the year ended 31 March 1965.

Of the total membership of 54 150 just over 50 per cent are Corporate members and over 9 000 of the total live abroad. The increase in membership during the year of 2 025 represents the largest in any year since the post-war influx of 1944-47.

Smiths Aviation Division's series 5 duplex automatic flight control system, a wholly British development with which British European Airways' Hawker Siddeley Trident airliners are fitted, has become the first automatic equipment in the world to be awarded official certification to carry out automatic touchdown during normal airline service.

This certification which has been awarded by the British Air Registration Board, marks an important stage in the evolution of automatic landing.

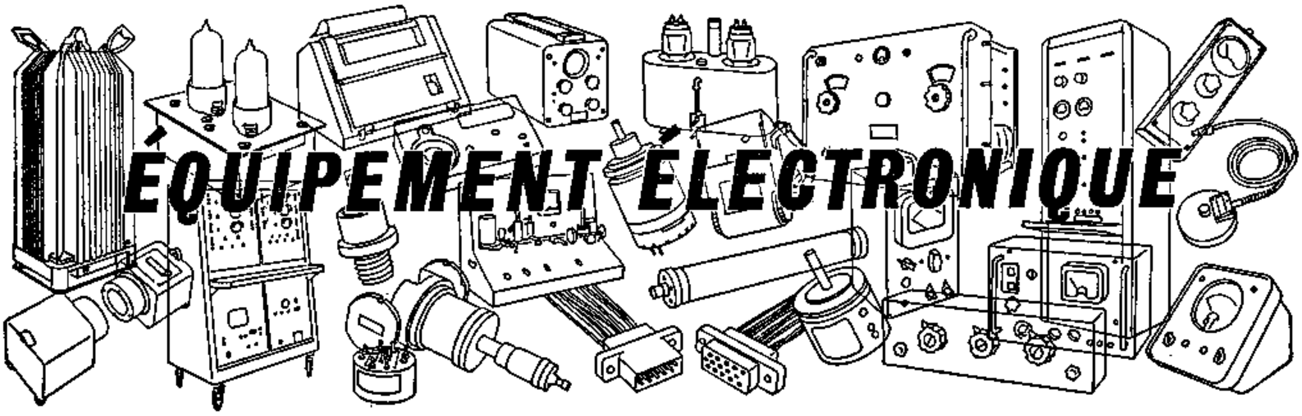
Autoflare as automatic touchdown is known, in which the execution of the difficult landing phase is shared between the pilot and the automatic SEP.5 equipment, will enable the Trident to land more accurately and more precisely than is possible with the human pilot solely in command.

Autoflare means that the automatically-controlled descent to the airfield, which normally reverts to pilot control at a height of about 150ft is extended to touchdown. The complex pattern of aircraft pitch movements and throttle adjustments which make possible a smooth landing is now controlled wholly by the SEP.5 equipment, the pilot's role during the landing being to keep the aircraft aligned with the runway and to monitor the autolanding system. In this way, the automatic flight control system carries out the most onerous part of the Control Manoeuvre in the final landing phase.

A £4·7 M telecommunications contract, awarded by the Government of Nigeria to a British consortium in which Automatic Telephone and Electric Co. Ltd (Plessey Group) and British Insulated Callender's Cables Ltd are the leading participants, has just been signed in Nigeria.

The contract is for the supply of British 'crossbar' telephone equipment to provide four modern exchanges totalling 25 000 lines as part of Nigeria's national telecommunications modernization programme. Subscriber trunk dialling facilities will be provided at Lagos, Port Harcourt and Enugu, together with extensions to existing exchanges, new exchange buildings, associated cable networks and civil engineering works.

Aero Electronics Ltd has completed the development and supply to the Department of Posts and Telecommunications in Malawi of two fully transistorized 3W AEL.507 REP (DS-3) special Duplex Repeater units to offer a v.h.f. radio link between Mzuzu and Likhoma Island, a 66 mile leg over Lake Malawi. These units are being integrated with the Malawi telephone system.



ÉQUIPEMENT ÉLECTRONIQUE

Une description basée sur des renseignements fournis par les fabricants de nouveaux organes, accessoires et instruments d'essai

Traduction des pages 478 à 483

COMMUTATEUR À GRADINS

Flight Refuelling Ltd, Wimborne, Dorset

(*Illustration à la page 478*)

La division d'électronique industrielle de la société Flight Refuelling Ltd a réalisé le commutateur à gradins RSC-41 qui est fondamentalement un unisélecteur à grande vitesse comportant des composants à lame vibrante agisant comme contact de commutation. Le nouveau commutateur est produit sous forme modulaire, chaque élément mesurant 4,12 cm de long × 2,45 cm de profondeur × 2,77 cm de hauteur, ce qui offre aux constructeurs de circuits une souplesse maxima avec un minimum de complications pour les interconnexions.

En fait, chaque module constitue une des positions d'un commutateur à gradins et comme il n'y a pas de limite au nombre de modules pouvant être reliés entre eux, un commutateur à gradins peut être construit avec autant de positions que l'exige le réalisateur. En variante, deux modules peuvent être utilisés pour former un élément bistable.

Les positions des broches sur la base de chaque module sont conformes à l'espacement standard de 0,15 cm pour plaquette de circuit imprimé et sont disposées de manière à ce que l'interconnexion entre les modules soit réduite à une simple courroie. Bien que prévu à l'origine pour répondre à la demande de montage de commutation à gradins, le module RSC-41 peut être utilisé dans de nombreuss applications de commutation.

EE 83 751 *pour plus amples renseignements*

CELLULE T.R.

Ferranti Ltd, Kings Cross Road, Dundee, Scotland

(*Illustration à la page 478*)

La société Ferranti Ltd. a réalisé une nouvelle cellule t.r. à bande S, type JF.20. Cette cellule, conçue pour assurer la protection des tubes à ondes progressives ou des amplificateurs paramétriques utilisés dans les systèmes de radar à impulsions, est à remplissage radioactif et ne nécessite donc pas une alimentation extérieure. Le taux d'ondes stationnaires, dans les limites de fréquence de 2700 MHz à 3100 MHz, est inférieur à 1,25 et la perte par insertion est nominalement de 0,3 dB.

Lorsque des récepteurs à cristal sont utilisés et que la cellule JF.20 est employée en liaison avec un autre composant récent—un limiteur à diode varactor type 104SSS—elle fournit une protection de récepteur suffisante pour une perte par insertion inférieure à 1,0 dB, sans qu'il y ait lieu d'employer des alimentations de puissance extérieures.

La protection du récepteur est assurée par la combinaison JF.20/104SSS même lorsque le radar n'est pas utilisé, ce qui obvie à la nécessité d'employer un obturateur mécannique de guide d'ondes.

Lorsque des amplificateurs paramétriques sont utilisés, ils doivent être insérés entre la cellule T.R. JF.20 et le limiteur 104SSS pour que la cellule puisse protéger l'amplificateur avec une perte par insertion minima cependant que le limiteur réduit à un niveau sûr l'interférence réfléchie par l'amplificateur dans le mélangeur à cristal.

La puissance nominale de la cellule JF.20 est de 100 kW de pointe et de 100 W de moyenne et elle doit être précédée d'un circulateur à ferrite ou d'un duplexeur à gaz (par exemple la série Ferranti 10D) si on envisage le fonctionnement à des niveaux de puissance supérieurs.

EE 83 752 *pour plus amples renseignements*

AMPLIFICATEUR OPÉRATIONNEL

Fenlow Electronic Ltd., Springfield Lane, Weybridge, Surrey

(*Illustration à la page 478*)

L'amplificateur à transistors stabilisé A7 ets un instrument à résistance photoélectrique conçu pour assurer une excursion de sortie comparable à celle du matériel à lampes. En outre, sa performance de dérive et de bruit est égale à celle d'amplificateurs fonctionnant à des niveau nettement inférieurs ce qui assure une amélioration considérable dans la précision du calcul.

L'amplificateur a été conçu pour toutes les applications ordinaires de calcul analogique et hybride ainsi que l'instrumentation. Sa sortie de puissance élevée permet de l'incorporer directement dans des appareillages d'asservissement de grandes dimensions.

Il est entièrement muni de transistors au silicium et de diodes. Il est, en outre, entièrement à l'épreuve des court-circuits.

L'amplificateur assure le fonctionnement linéaire jusqu'à une sortie de ±100 V à 50 mA et la sortie total est prévue jusqu'à 5 kHz. Le gain en courant continu est supérieur à 10×10^6 et le produit de la largeur de bande de gain est de 10 MHz. L'impédance d'entrée est de 100 kΩ, le bruit est inférieur à 50 μV de crête à crête et la dérive est inférieure à 1 μV/°C.

EE 83 753 *pour plus amples renseignements*

FOURREAU DE CÂBLES

Insuloid Manufacturing Co. Ltd, Leestone Road, Wythensbawe, Manchester

La société Insuloid Manufacturing Co. Ltd a réalisé une nouvelle série de fourreaux de câbles dont les caractéristiques permettent d'économiser du temps et de l'argent. Le nouveau type de fourreau comporte des fentes sur toute sa longuer et chacunes de ces fentes est à ouverture étroite ce qui permet de faire sortir des câbles à n'importe quel point. Ces petites ouvertures sont conçues de manière à empêcher les câbles de "s'échapper" durant l'assemblage. On peut ainsi séparer les divers câblages tout en les enchaînant l'un à l'autre au moyen de pièces de connexion. L'enveloppe comprend une pièce de fixation à action rapide retenue par elle même. Les fourreaux Insuloid Cable sont cannelés à la base pour faciliteér le forage rapide et précis et ils peuvent être entassés en parallèle. Quatre formats sont prévus ainsi que trois différentes positions des fentes. Le nouveau fourreau est en chlorure de polivynile rigide et de poids léger et il est donc à l'épreuve de la corrosion et n'exige aucun entretien après son installation.

EE 83 754 *pour plus amples renseignements*

OUTILS DE CONTRÔLE À RELAIS

P. W. Allen & Co., 253 Liverpool Road, London, N.1

(*Illustration à la page 479*)

La société P. W. Allen & Co. vient de mettre au point une nouvelle série d'outils de contrôle pour l'examen des contacts et des cames sur les mécanismes de commutation téléphoniques.

La série des outils Allen type TS comprend une poignée de batterie, trois sondes légères et quelques accessoires à petit miroir.

Il permet d'éclairer et de fixer un petit miroir contre un contact, même à l'intérieur de plusieurs rangées, de manière à ce que la surface de contact puisse être facilement examinée.

Il a été prévu pour le contrôle des contacts, cames et mécanismes d'unisélecteurs de toute sorte. Il est le fruit des travaux des ingénieurs téléphoniques, qui se sont basés sur la gamme étendue des sondes lumineuses Allen pour déterminer le moyen le plus économique de répondre aux conditions normales de contrôle tant dans les fabriques qu'à l'extérieur.

Les petites sondes ont un diamètre inférieur à 0,31 cm et comprennent de miroirs de même largeur. Des loupes à grossissement X2 ou X3 montées sur les sondes permettent la vision de près avec grossissement approprié à faible puissane.

EE 83 755 *pour plus amples renseignements*

EQUIPEMENT DE CONCENTRATION DE DONNÉES

Hawker Siddeley Dynamics Ltd, Whitley Works, Coventry, Warwickshire

(*Illustration à la page 479*)

La conception de ce nouvel équipement de concentration et d'enregistrement de données pour les applications marines, industrielles et scientifiques, permet de l'utiliser soit comme simple enregistreur soit comme système de concentration de données avec facilité de calcul sur ligne.

On peut échantillonner un maximum de mille signaux analogiques de niveau élevé ou de niveau bas à 20 entrées/sec. Un dispositif de protection indépendant à chaque entrée allié à un convertisseur analogique/numérique d'un modèle perfectionné (en voie d'être breveté) permet de réduire à de faibles limites les effets d'interférence de mode commun.

L'équipement a été étudié pour le fonctionnement continu dans des limites de grande précision pendant six mois au moins sans étalonnage. L'étalonnage automatique s'effectue à chaque cycle et des limites différentes et indépendantes peuvent être fixées pour chaque entrée.

Les cycles sont à répétition continue, mais ils peuvent être également mis en oeuvre individuellement, soit à la main soit à certains intervalles par commande de minuterie numérique.

Des données de sortie imprimées sous diverses formes et une bande perforée revue sont prévues simultanément, ainsi que des affichages visuels des valeurs et identités de signaux d'entrée, de condition d'alarme et de temps.

EE 83 756 *pour plus amples renseignements*

CONTRÔLEUR DU PROFIL DE SURFACE

G. V. Planer Ltd, Windmill Road, Sunbury-on-Thames, Middlesex

(*Illustration à la page 479*)

Ce contrôleur du profil de surface a été conçu pour l'indication du contour des surfaces et il peut mesurer l'épaisseur de pellicules minces de 0 à 50 000 Å.

Il utilise un transducteur de déplacement pour contrôler le mouvement du style par rapport à un patin de données, tant l'un que l'autre traversant la surface devant être mesurée. Deux vitesses d'avance sont prévues. La mesure est indiquée par l'aiguille enregistreuse mais le contour peut aussi bien être enregistré sur bandes qui permettent de déterminer l'épaisseur des pellicules à un degré de précision supérieur à 100 Å.

L'appareil est particulièrement indiqué pour mesurer deux couches déposées sous vide, ainsi que pour contrôler la qualité de la surface des matériaux.

L'appareil comprend deux éléments: le détecteur et le contrôleur, tous deux logés dans des coffrets métalliques.

EE 83 757 *pour plus amples renseignements*

GÉNÉRATEUR D'IMPULSIONS DOUBLES

Venner Electronics Ltd, Kingston By-Pass, New Malden, Surrey

(*Illustration à la page 479*)

Le générateur d'impulsions doubles constitué de corps solides, type TSA 628, a été conçu par la société Venner Electronics Ltd. Il est entièrement muni de semi-conducteurs au silicium et sa gamme de fréquence s'étend de 2,5 Hz à 2,5 MHz lorsqu'il est employé comme générateur d'impulsions uniques, et de 2,5 Hz à 2 MHz lorsqu'il est employé comme générateur d'impulsions doubles. Trois méthodes de fonctionnement sont prévues: en marche libre, à déclenchement par impulsions extérieures de 2 à 50 V ou par contacts, et par bouton-poussoir.

Son atténuateur à commutation permet d'atteindre des sorties de 20, 10, 5, 2, 1 et 0,5 V, une commande de réglage précis permettant un calage minimum de 0,2 V. L'impédance de sortie est de 100 Ω à 20 V et 50 Ω par ailleurs.

Des impulsions négatives positives réelles par rapport à la masse peuvent être obtenues. Les temps de montée et de chute sont de 10 nsec de durée d'établissement, de 15 nsec de durée d'affaiblissement, et les circuits de largeur d'impulsion et de retard sont à variation continue de 100 nsec à 10 sec. La sortie de pré-impulsion est de ±5 V au minimum (déchargé) sur courant alternatif et la largeur est de 100 à 150 nsec.

EE 83 758 *pour plus amples renseignements*

MICROVOLTMÈTRE À COURANT ALTERNATIF

Levell Electronics Ltd, Park Road, High Barnet, Hertfordshire

(*Illustration à la page 479*)

Cet instrument portatif peut être utilisé soit comme voltmètre à courant alternatif soit comme amplificateur à courant alternatif. La réponse de fréquence s'étend de 1 Hz à 3 MHz. Seize gammes sont prévues en plots de 10 dB, de 15 μV à 500 V de déviation totale, avec choix de la gamme voulue par simple commutateur rotatif comportant également des position "arrêt" et "lecture".

L'instrument est muni de deux amplificateurs à contre-réaction séparés. Le gain du premier amplificateur peut être varié par plots de 10 dB à 80 dB, la contre-réaction étant appliquée pour stabiliser le gain. La sortie est reliée à des prises de panneau et à l'entrée du deuxième amplificateur qui actionne un circuit redresseur à ondes intégrales et un instrument de mesure à cadre mobile. La contre-réaction de cet amplificateur est telle que le courant dans l'instrument de mesure doit se maintenir en rapport linéaire avec la tension d'entrée. La déviation sur la totalité de l'échelle se produit, pour une sortie de 150 mV, à partir du premier amplificateur. L'impédance d'entrée du deuxième amplificateur est suffisamment élevée pour présenter une charge négligeable à la sortie du premier amplificateur, de sorte que le signal aux prises de sortie est exempt de distorsions dues à la charge du redresseur. Un avantage supplémentaire de ce dispositif est que la largeur de bande de l'instrument peut être réduite en branchant des composants réactifs sur les bornes de sortie. Cet avantage est particulièrement utile lorsqu'on mesure de faibles signaux sur une bande de fréquence limitée, car toute réduction de la largeur de bande entraîne également une réduction du niveau de bruit d'entrée.

L'instrument est idéal comme pré-amplificateur oscilloscopique car il se caractérise par un faible niveau de bruit, une microphonie réduite, l'absence de bourdonnement et il est indépendant du courant de secteur.

EE 83 759 *pour plus amples renseignements*

CONTRÔLEUR DE MATÉRIAUX

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

(*Illustration à la page 480*)

Le contrôleur dynamique de matériaux type 1825 a été conçu principalement pour les démonstrations dans les établissements d'enseignement. Il mesure le module de Young, le facteur Q et la capacité d'amortissement de spécimens de barres excitées de manière à produire une résonance.

Il comprend un oscillateur, un transformateur-élévateur et une alimentation de polarisation à courant continu, ainsi qu'un banc d'essai de précision, un amplificateur et un oscilloscope. Une tige uniforme ou une barre du matériau soumis à l'essai est fixée à son centre sur le banc d'essai, deux petits entrefers (20 – 50 microns) séparant ses extrémités libres des têtes électrostatiques du banc d'essai. Un champ alternatif appliqué à la tête d'entraînement excite le spécimen afin qu'il produise une résonance, le petit signal à courant continu qui en résulte à la tête de captage est appliqué au moyen de l'amplificateur à l'oscilloscope ou à tout autre détecteur approprié.

Le contrôleur type 1925 peut être utilisé avec un appareillage supplémentaire pour la mesure industrielle, mais sa précision n'est toutefois pas suffisante pour tous les travaux techniques. Le module de Young peut être déterminé avec une précision suffisante sur l'échelle de fréquence de l'oscillateur. Cette précision atteint un degré plus élevé si l'on emploie un fréquencemètre extérieur. Le facteur Q de spécimens à faible Q peut être déterminé sur l'oscillateur, le facteur Q de spécimens à Q moyen pouvant être déterminé sur un fréquencemètre extérieur. Une méthode différente, basée sur le décrément logarithmique, est nécessaire pour les spécimens à Q élevé, et une variante de l'appareil avec ce dispositif supplémentaire peut être obtenue L'oscillateur et l'oscilloscope sont des éléments autonomes et peuvent donc être utilisés indépendamment pour d'autres travaux, ce qui constitue une considération importante dans les laboratoires de collèges où l'espace nécessaire peut parfois faire défaut. Le modèle standard est muni d'un oscilloscope à faisceau unique, sur lequel on obtient les chiffres de Lissajous. Des variantes supplémentaires sont, cependant, livrables avec un oscilloscope à double faisceau ou sans oscilloscope, si l'on entend employer un oscilloscope extérieur ou un voltmètre à lampe comme détecteur.

Le banc d'essai standard peut recevoir des spécimens de 150 à 280 mm de long et de 6 à 15 mm de diamètre, un calibre spécial étant prévu pour faciliter le processus de montage qui prend moins d'une minute. L'oscillateur utilisé est un modèle à grande puissance de 6 W donnant une sortie de 20 Hz à 20 kHz. L'équipement électronique mesure 52 cm × 43,18 cm × 58,42 cm de hauteur.

EE 83 760 *pour plus amples renseignements*

INSTRUMENT DE MESURE DU RAPPORT DE POURCENTAGE

Bendix Electronics Ltd, High Church Street, New Basford, Nottingham

(*Illustration à la page 480*)

Cet instrument de mesure du rapport de pourcentage comprend une série d'instruments transistorisés utilisant les méthodes de comptage numérique pour mesurer et afficher la différence de pourcentage entre deux processus, à savoir la vitesse du rouleau et la vitesse de l'arbre. L'instrument est particulièrement utile pour les processus industriels de calandrage qui nécessitent la mesure du pourcentage d'élargissement et de retrécissement. Des différences de 0,01% peuvent être facilement enregistrées.

L'appareil se compose essentiellement de deux compteurs, l'entrée étant appliquée aux deux compteurs simultanément, dont l'un refoule les deux entrées au terme du comptage de 1000 coups (ou quelle que soit la base de comptage choisie). Les entrées sont reliées de manière à ce que l'indication du second compteur soit égale ou supérieure à 1000 (selon la base de comptage choisie). Le transducteur d'entrée préféré est le 'Rotapulse 191A' ou le générateur d'impulsions 491A. D'autres circuits d'entrée peuvent être montés sur le compteur pour recevoir d'autres types de transducteurs d'impulsions.

Si le rapport mesuré dépasse les limites de 0 à $\pm 9{,}9(9)\%$, un des deux signaux suivants sera déclenché: – pour l'échelle inférieure et + pour l'échelle supérieure. Dans ces conditions l'affichage numérique est refoulé.

EE 83 761 *pour plus amples renseignements*

ÉLÉMENTS À CIRCUIT

Morecombe Electrical Equipment Co. Ltd, Westgate Works, Morecombe, Lancashire

(*Illustration à la page 480*)

Plusieurs nouveaux éléments ont été ajoutés à la gamme "Meecostatic" de modules transistorisés logiques, d'entrée, de sortie et à circuit, conçus pour le contrôle industriel. Ces nouveaux modules comprennent une minuterie RC, une matrice de décodage 'binaires/décades', une porte équivalente à deux circuits, un générateur à rampe, un compteur binaire réversible et des modules réglables de déclenchement sur niveau et hors-niveau.

Ces nouveaux éléments portent le nombre total de modules dans cette gamme à 35. Ils sont tous électriquement et mécaniquement compatibles les uns avec les autres.

Pour assurer l'assemblage satisfaisant de ces modules, on a mis au point un système de châssis de bâti de 48,2 cm dont on voit une section dans notre gravure. Chacun des étages peut contenir jusqu'à cinq châssis d'une largeur de 7,62 cm ou un nombre proportionnellement plus petit d'éléments de 15,24 cm ou d'éléments combinés de 15,24 cm et 7,62 cm. Chacun des châssis peut loger jusqu'à 60 modules de 1,27 cm et les châssis peuvent être branchés sur le cadre. Les modules sont construits principalement en aluminium, d'un fini de haute précision. Les surfaces frontales qui pourraient être visibles de l'extérieur sont anodisées en noir et comportent des poignées chromées permettant de retirer facilement le châssis des étages. D'autres surfaces en aluminium ont un fini de satin. Notre gravure montre comment les châssis peuvent être branchés sur un élément d'interconnexion et peuvent ainsi être rendus aptes aux travaux de contrôle. A l'avant du châssis on peut voir une sonde d'essai. Cette dernière est reliée à un voyant lumineux, de sorte qu'on peut effectuer facilement le contrôle de l'état "d'arrêt" ou de "marche" de n'importe lequel des modules dans cette section de l'appareil.

Il y a lieu de signaler également que cette gamme comporte une série de blocs d'alimentation de ½ A, 1 A et 2 A. Ces blocs donnent une sortie régulée de –24 V, un rapport de stabilisation supérieur à 5000 : 1 et une résistance de sortie à courant continu inférieure 0,002 Ω l'impédance de sortie étant inférieure à 0,1 Ω à 100 kHz. Ces blocs comportent également une alimentation de +24 V ayant une connexion de voltage à zéro commun. Cette alimentation positive est principalement utilisée pour les travaux de polarisation et emploie une régulation Zéner simple.

EE 83 762 *pour plus amples renseignements*

EQUIPEMENT MICROONDES

G.E.C. (Telecommunications) Ltd, Telephone Works, Coventry

(*Illustration à la page 481*)

Une nouvelle gamme d'appareils de radio microondes entièrement à semi-conducteurs est en cours de production par la société G.E.C. (Telecommunications) Ltd. Les appareils microondes à semi-conducteurs ont utilisé jusqu'ici des dispositifs thermoélectroniques dans une partie quelconque de leur circuit. Ces dispositifs étaient habituellement des tubes à ondes progressives, des klystrons ou des triodes microondes dans l'étage de sortie de puissance. La société G.E.C. produit maintenant avec succès des appareils entièrement munis de semi-conducteurs. Les deux principaux avantages de ces nouveaux appareils sont une fiabilité accrue et une réduction importante dans la consommation d'énergie, même pour les appareils n'utilisant les dispositifs thermoélectroniques que dans un étage. Ces deux avantages sont particulièrement importants dans les pays où on a difficilement accès aux stations éloignées et où tant l'alimentation en courant électrique que l'entretien doivent être réduits au minimum.

L'un des deux appareils actuellement produits par la société G.E.C. convient pour les liaisons principales et peut porter 960 circuits de parole, ou pour un canal de télévision et son dispositif sonore dans les bandes de fréquence de 1 700 à 1 900 MHz et de 1 900 à 2 300 MHz. La sortie de puissance de l'émetteur de cet appareil est de 2 W, obtenue uniquement des semi-conducteurs. Le récepteur comprend un étage d'entrée à faible bruit, employant un amplificateur à diode tunnel dont le facteur de bruit est 5½ dB. L'équipement à large bande est relié à un canal radiophonique auxiliaire entièrement à semi-conducteurs et à un appareil à commutation de protection à multi-canaux de haute fréquence.

Le deuxième appareil fonctionne dans la bande de fréquence de 7 425 à 7 725 MHz et en raison de sa capacité de 300 canaux de parole, il est particulièrement indiqué pour fournir des circuits de haute fiabilité aux petites administrations téléphoniques, ainsi que pour les lignes de branchement des lignes principales à haute capacité. Un répétiteur à double voie et à deux directions se trouve dans l'un des châssis. Il est complet avec son appareillage de commutation de fils de transmission et d'alarme télécommandée. Cet appareillage permet de supprimer n'importe quelle portion de la bande de base et de l'introduire dans n'importe quelle station de répétiteur sans démoduler les circuits directs.

EE 83 763 *pour plus amples renseignements*

RELAIS DE VERROUILLAGE MINIATURE

Distributeurs: B & R Relays Ltd, Temple Fields, Harlow, Essex

(Illustration à la page 481)

La société B & R Relays Ltd, chargée de la distribution des relais miniature Siemens et Halske dans le Royaume-Uni, vient d'introduire sur le marché britannique le nouveau relais de verrouillage magnétique Siemens, type P. Ce relais existe en deux formats. Le petit format comprend deux contacts légers de permutation, tandis que le grand format comprend quatre contacts de permutation, six contacts normalement ouverts et six contacts normalement fermés ou deux contacts de puissance à permutation. Les contacts sont soit en argent soit en alliage d'or pour le plupart des versions. Le relais à "berceau" type P (référence V23003) a été conçu pour fonctionner sur impulsions.

Le fonctionnement s'effectue de telle manière que lorsque le relais est muni d'une seule bobine, il peut être excité ou mis en impulsions, et il fonctionnera alors et demeurera dans cet état jusqu'à ce que la bobine soit excitée dans la direction opposée.

En variante, le relais peut être muni de deux bobines et il peut alors être amorcé en excitant une des bobines, et désamorcé en excitant l'autre.

EE 83 764 *pour plus amples renseignements*

COMMUTATEURS À BOUTON-POUSSOIR

Standard Telephone & Cables Ltd, Electro-Mechanical Division, West Road, Harlow, Essex

(Illustration à la page 481)

La gamme des commutateurs à bouton-poussoir réalisée par Mayr, une société associée à la STC, est maintenant vendue par la division d'électromécanique de STC. Ces assemblages offrent à l'ingénieur de contrôle une variété de possibilités de commutation peu habituelle pour ce type d'élément. La souplesse du système fournit de multiples combinaisons de maintien de bouton et de déclenchement de bouton. Ces possibilités comprennent le déclenchement mutuel en groupe ou le déclenchement commun.

En plus des divers boutons de maintien et de déclenchement, dont chacun peut effectuer une variété d'opérations de commutation, les assemblages Mayr peuvent contenir des boutons "d'impulsions" dans un même élément. Un grand nombre de combinaisons électriques et mécaniques sont possibles. Des assemblages de commutation de diverses constructions (à haute ou à faible puissance) peuvent être fournis sur le même chassis à bouton de commutation. Chacun des boutons peut être prévu pour effectuer un nombre déterminé d'opérations de permutation, de fermeture et d'ouverture.

La plupart des gammes sont livrables avec banc de douze boutons. Ces bancs peuvent être empilés de manière étroite, soit verticalement soit horizontalement. Il sont fournis avec des housses transparentes amovibles pour l'insertion de signaux en couleurs d'identification ou de caractères écrits.

EE 83 765 *pour plus amples renseignements*

CONTRÔLEUR DE FRÉQUENCE

Kent Precision Electronics Ltd, Vale Road, Tonbridge, Kent

(Illustration à la page 481)

La société Kent Precision Electronics Ltd a récemment mis au point un nouveau circuit constitué de corps solides qu'on monte à l'intérieur de l'assemblage mécanique de leurs instruments étalonnés standard constitués de corps solides pour le contrôle, la protection ou l'avertissement dans les travaux de contrôle de fréquence.

Le système se compose essentiellement d'un convertisseur fréquence/courant continu qui alimente un signal de courant continu proportionnel à la fréquence dans une petite gamme d'environ 45 à 55 Hz. La fréquence d'entrée est d'abord convertie en une sortie d'impulsion de la même fréquence, cette dernière étant ensuite injectée à un intégrateur linéaire dont la sortie est transmise à l'entrée d'un contrôleur standard étalonné KPE constitué de corps solides. La plaquette a circuit imprimé supplémentaire se trouve à l'intérieur d'un coffret standard type "Elite" et elle obtient son alimentation du reste de l'instrument. L'alimentation auxiliaire de l'instrument est normalement fournie par la tension alternative de secteur et l'instrument peut être utilisé pour contrôler la fréquence de sa propre alimentation auxiliaire, si nécessaire. Il peut également être utilisé pour contrôler une entrée secondaire.

Des modèles simples ou doubles sont livrables comme appareils standard, l'instrument double ayant des commande concentriques. Dans chacun des cas, un relais intérieur à contact de permutation d'une valeur nominale de 5 A 240 V c.a. peut être actionné au point fixé. La sortie peut aussi consister en un tyrristor proportionnel ou en un élément de contrôle marche/arrêt.

EE 83 766 *pour plus amples renseignements*

LASER AU GAZ DÉMONTABLE À COURANT CONTINU

Scientifica, 148 St. Dunstan's Avenue, Acton, London, W.3

(Illustration à la page 482)

La mise au point du laser à gaz à courant continu Scientifica résulte des efforts mis en oeuvre pour produire un instrument démontable ayant la précision et la stabilité voulue pour les travaux de recherche, tout en étant suffisament robuste pour l'usage général de laboratoire. Sa construction permet de l'utiliser soit comme appareil libre, pouvant être employé dans n'importe quelle orientation, soit comme assemblage de banc optique dont les miroirs résonateurs peuvent rester fixés sur le laser ou être tenus par des supports séparés de banc optique. Cette souplesse de montage permet d'effectuer des expériences avec différents espacements de miroir ainsi qu'une étude détaillée des montages de résonateur stables et instables. Elle permet en outre de placer des composants optiques standard à l'intérieur de la cavité de résonance.

L'instrument comprend d'autre part des miroirs à revêtement solide de 17 couches qui peuvent être nettoyés avec un tissu ordinaire de lentille ou lavés dans l'alcool. Le tube du laser à gaz à courant continu de conception spéciale a une durée de vie exceptionnellement longue. La sortie du tube est une onde progressive continue, polarisée dans la zone infrarouge et rouge du spectre et ayant des sorties typiques de 3 mW à 6328 Å et 10 mW à 11530 Å.

Un seul câble relie le laser à un élément d'entraînement à courant continu doté d'une commande pour le réglage de la sortie de lumière, le niveau d'excitation étant affiché sur un instrument de mesure à cadre mobile de 50 mA. L'appareil peut être utilisé soit à 110 V soit à 230 W c.a. et fournit une sortie en tension continue à faible ondulation qui produit à son tour une modulation inférieure à 1% du faisceau du laser. Tant le laser que l'élément d'entraînement comprennent des composants à caractéristiques nominales des plus élevées et garanties inconditionnellement comme étant exemptes de défauts pendant un an.

Un manuel d'emploi est fourni avec chaque laser et donne des détails des expériences qui peuvent être effectuées en utilisant cette nouvelle source de lumière cohérente. Ces travaux comprennent tant des expériences physiques que géométriques, telles que la diffraction Fraunhofer, la théorie d'Abbé, le déphasage Doppler dû à un miroir mobile, la collimation du faisceau du laser donnant un étalement de quelques centimètres par kilomètre.

Grâce au faisceau monochromatique de grande puissance de l'instrument, la procédure de réglage de toute expérience d'interférence est considérablement simplifiée. En outre les dessins d'interférence obtenus sont d'une intensité nettement supérieure à ceux obtenus par les méthodes antérieures et ils sont donc beaucoup plus frappants. Cet avantage permet d'effectuer des expériences dans des espaces non-obscurcis et de faire voir à de grands auditoires les effets d'interférence.

EE 83 767 *pour plus amples renseignements*

RELAIS MICRORUPTEURS

Keyswitch Relays Ltd, 120-132 Cricklewood Lane, London, N.W.2

(Illustration à la page 482)

Une gamme de relais microrupteurs, basée sur les types postaux standard 300 et 600, spécialement conçus pour le contrôle de température et de circuits, a été créée par la société Keyswitch Relays Ltd. Ces relais se caractérisent par leur valeur de contact élevée, leur faible force de commande et un taux de commutation rapide. Les contacts de fonctionnement sont entièrement enfermés afin de les rendre propres aux atmosphères défavorables. Des courants allant jusqu'à 20 A à 480 V c.c. peuvent être branchés. Un rapport de tranfert de puissance de 1 : 100 000 peut être réalisé.

Les deux modèles de base livrables sont: le type GPO 3 000 (à commutateur simple ou multiple) et le type GPO 600 à commutateur unique. La tension de régime (type 3 000) s'étend d'un minimum de 1 V c.c. à 300 V c.c. et peut atteindre un maximum de 440 V c.a. avec des composants supplémentaires sur le type 600; la tension de régime est de moins de 1 V c.c. à 175 V c.c. et elle peut aller jusqu'à 440 V c.a.

La résistance au courant continu s'étend de moins de 1 Ω à 50 kΩ pour le type 3 000 et de moins de 1 Ω à 10 kΩ pour le type GPO 600.

Les variations de la gamme standard comprennent un double induit sur le type 3 000 pour permettre le contrôle indépendant des microrupteurs.

Une polarisation de fonctionnement réglable peut être fixée sur le type 600 et le type 3 000 à commutateur unique. Le verrouillage à réenclenchement manuel peut être fixé sur le type 3 000. Les bobines peuvent être tropicalisées sur demande et des connexions de commutation à vis ou à soudure peuvent être fournies.

EE 83 768 *pour plus amples renseignements*

FICHES ET DOUILLES MULTIPLES

Oxley Developments Ltd, Priory Park, Ulverston, Lancashire

(Illustration à la page 482)

La société Oxley Developments Co. Ltd vient d'introduire un nouveau type de fiches et douilles multiples pour montage sur châssis. Les connecteurs sont munis de 24 contacts dorés indépendants, prévus pour les tiroirs d'équipement de panneaux de bâti et de consoles. Le connecteurs sont munis d'un dispositif de verrouillage mécanique incorporé (tenu par vis) qui applique simultanément une pression sur les contacts et préserve ainsi la résistance de contact réduite pendant de longues périodes dans des conditions climatiques ardues.

EE 83 769 *pour plus amples renseignements*

FORETS POUR CIRCUITS IMPRIMÉS

Drill Service (Horley) Ltd, 89 Albert Road, Horley, Surrey

La plupart des plaquettes à circuit imprimé sont faites en un matériau abrasif provoquant l'usure rapide des forets normaux HSS. Non seulement les pointes s'émoussent très rapidement mais même les perforateurs cylindriques s'usent assez vite et produisent des trous de dimensions inférieures.

Les forets de plaquettes à circuit imprimé au carbure de tungstène solide ont été spécifiquement conçus pour la perforation des plaquettes à circuit imprimé. L'emploi du carbure de tungstène élimine entièrement l'usure et assure aux pointes une durée de vie supérieure à la normale. 6 000 trous peuvent être normalement pratiqués entre les meulages.

Les plaquettes de circuit exigent également des trous sans bavures, ce qui nécessite un angle élevé de ratissage et une spirale prononcée. En outre, le dégagement des copeaux aux grandes vitesses d'alimentation utilisées, nécessite une flasque très mince.

Les forets en question ont une spirale très élevée et une flasque mince avec des cannelures meulées à un degré élevé de microfinissage.

Les tolérance de diamètre sont de 0,0003″. Cette dimension est nécessaire pour produire en quantité des trous de précision d'une tolérance de 0,0005″. La longueur hors-tout est la même pour tous les formats, soit 3,81 cm, la longueur de la cannelure étant de 1,27 cm.

EE 83 770 *pour plus amples renseignements*

APPAREIL DE MESURE DE VIBRATIONS

Associated Engineering Ltd, Cawston, Rugby, Warwickshire

(Illustration à la page 483)

La société Associated Engineering Ltd a ajouté trois nouveaux éléments de matériel électronique à sa gamme d'instruments de mesure de déplacement et de vibration. Ils ont été conçus pour pouvoir être employés avec des transducteurs subminiature de déplacement inductif. L'équipement se compose d'un oscillateur stable à cinq gammes de commutation de fréquence entre 50 kHz à 150 kHz et d'un amplificateur qui, lorsqu'il est entraîné par l'oscillateur, fournit six sorties à très faible impédance de 20 V de crête à crête. Ces deux éléments peuvent exciter six canaux d'équipement de mesure à l'aide de transducteurs inductifs, chacun des canaux fonctionnant en liaison avec un filtre détecteur qui fournit une charge équilibrée pour le transducteur et assure une détection appropriée. Les canaux sont, en outre munis d'un filtre spécial pouvant séparer le signal modulé de la porteuse.

Grâce à cette combinaison, une sensibilité de 10 mV par mouvement d'un millième de pouce peut être obtenue et des mouvements de fréquence atteignant 25 kHz peuvent être indiqués sans distorsion perceptible.

Notre gravure montre chacun des trois éléments mentionnés ainsi que quatre transducteurs inductifs des deux formats actuellement produits.

EE 83 771 *pour plus amples renseignements*

GÉNÉRATEUR DE SIGNES NUMÉRIQUES

Digital Measurements Ltd, Salisbury Grove, Mychett, Aldershot, Hampshire

(Illustration à la page 483)

Le générateur de signes Datapulse modèle 208, assure le contrôle souple du format et de la teneur de données pour la simulation de données numériques multivoies à vitesse élevée.

L'appareil se caractérise par un taux de minutage de 5 Hz à 5 MHz, de caractères parallèles à 8 chiffres binaires 1 à 16 caractères en série par bloc de données et à répétition de caractère ou de bloc, une sortie de 6 V dans 50 Ω, et des formats RZ ou NRZ.

Dans le mode parallèle, les caractères parallèles à longueur de bloc variable et à contrôle de bloc par image permettent la simulation de données sur papier ou sur bande magnétique. La répétition contrôlée numériquement peut être effectuée pour produire des dessins utiles d'une plus longue durée.

Dans le mode en série, huit signes différents en série de seize chiffres binaires peuvent être commutés sur une seule ligne avec contrôle numérique des mots par image pour la simulation de données de caractères parallèles ou de systèmes multiplex. Des lampes de lecture assurent l'indication continue de l'état du programme. Des tampons de sortie de 50 Ω entraînent directement les câbles d'interconnexion sans qu'il y ait lieu d'employer des tampons supplémentaires. Les images peuvent être répétées de façon continue et remises en cycle à la main ou répétées par commande extérieure de signaux. Un bouton-poussoir sur le panneau frontal permet de réenclencher le dispositif de mise en marche. Le programme peut être arrêté par commande extérieure de signaux.

La sortie synchrone est positive et à impulsions de 6 V, d'une durée de 50 nsec avec une impédance de source de 100 Ω et on peut la faire coïncider avec n'importe quel caractère ou combinaison de caractères. La synchronisa ion de la mise en marche coïncide avec e premier chiffre binaire ou caractère dans chaque image. La synchronisation et l'image coïncident avec le dernier chiffre binaire ou caractère.

EE 83 772 *pour plus amples renseignements*

ENREGISTREUR D'UNITÉS D'INTENSITÉ DU CHAMP MAGNÉTIQUE

Preformations Ltd, Cheney Manor, Swindon, Wiltshire

(*Illustration à la page 483*)

La société Preformations Ltd, une filiale de la société Plessey, a mis au point un nouvel enregistreur portatif d'unités Gauss, assurant la mesure statique linéaire du champ magnétique dans la gamme étendue de 5 000 à 105 000 Gauss. La société Preformations Ltd, dont les produits sont vendus sous la marque déposée "Magloy", est un des plus important fabricants britanniques d'aimants permanents.

Cet instrument serait le premier à faire usage de l'effet magnétostrictif, relativement peu connu. Il utilise une sonde se composant d'une petite plaque d'antimoniure d'indium ou d'arseniure d'indium. Cette plaque change de résistance par rapport à sa surface lorsqu'elle est placée dans un champ magnétique normal, ce changement étant contrôlé par un ohmmètre transistorisé qui alimente un amplificateur à courant continu.

Ce changement de résistance, lorsqu'il dépasse 5 kG, devient une fonction sensiblement linéaire de la densité de flux. Ce phénomène a été vérifié dans des champs magnétiques atteignant 105 kG.

L'étalonnage intérieur a été prévu de manière à pouvoir se passer de champs d'étalonnage extérieurs. L'instrument peut aussi être utilisé en liaison avec un oscilloscope, un enregistreur à plume ou un dispositif télémétrique.

La minceur de la sonde pouvant se réduire à 0.15 mm, l'instrument peut être utilisé même dans de très petits entrefers et l'indication est toujours positive, quelle que soit la polarité de champ. L'un des avantages de ce type de sonde est qu'elle peut être utilisée pour effectuer des relevés détaillés de champ magnétique pour mesurer la linéarité et l'homogénéité.

Cet instrument, conçu à l'origine pour les recherches de résonance ferromagnétique aux fréquences microondes, peut être utilisé également dans les solénoïdes super-conductrices et dans les appareils de recherche de résonance de protons, de spectrométrie de masse et de physique des plasmas. En reliant l'instrument à un oscilloscope, on peut mesurer des courants à impulsions très élevées comme par exemple ceux utilisés dans les blocs d'alimentation de lasers et dans les magnétiseurs d'impulsions, et afficher les formes d'ondes produites.

EE 83 773 *pour plus amples renseignements*

PROJECTEUR À MIRE DE POINTS

Marconi Instruments Ltd, St. Albans, Hertfordshire

(*Illustration à la page 483*)

Le projecteur à mire de points type TM8098 emploie un miroir semi-réflecteur pour permettre la vision normale d'indications oscilloscopiques ainsi que la projection de différents types de données de référence sur le même plan que l'image elle-même, sans le moindre parallaxe. Les applications caractéristiques de cet accessoire d'une grande souplesse d'emploi sont la projection de mires de points pour travaux de haute précision, les mires de points spéciales pour le contrôle d'impulsions, les formes d'ondes ordinaires et les lignes de limite pour le contrôle de production. Les données devant être projetées peuvent être produites par une réduction photographique des dessins principaux, par la photographie directe de formes d'ondes de référence ou par des esquisses sur des feuilles "Scribecoat". L'assemblage compact se fixe simplement et rapidement sur le couvercle rectangulaire dont sont munis actuellement tous les oscilloscopes Marconi. Le courant nécessaire à l'éclairage du projecteur s'obtient par la douille fixée à tous les instruments courants. Cette douille transmet l'alimentation du courant fourni pour l'éclairage normal des mires de points. Cet accessoire existe également en une autre version, le modèle TM8 097, pouvant être utilisée pour les instruments dont le diamètre du couvercle est de 14,28 cm. Ce modèle obtient son courant de 6, V au moyen d'un connecteur MODAC qui s'adapte à la sortie du palpeur cathodique des appareils TF2 200. Pour d'autres oscilloscopes une alimentation séparée de 6.3 V est nécessaire.

EE 83 774 *pour plus amples renseignements*

Résumés des Principaux Articles

Le Velocimètre "Sing-Around" et son emploi pour mesurer la grandeur des remous turbulents dans la mer par D. J. Dunn

Résumé de l'article aux pages 432 à 436

Différentes méthodes pour mesurer les variations de vitesse du flux d'eau turbulente sont analysées. Un brève description est donnée du vélocimètre "Sing-Around" dans laquelle il est démontré que cet appareil peut être utilisé également pour mesurer les fluctuations de vitesse du flux.

Un Intensimètre à transistors à résolution de 2 microsecondes par B. Shepherd

Résumé de l'article aux pages 437 à 441

Cet article décrit un intensimètre à impulsions aléatoires comportant sept gammes de taux de 10 impulsions par seconde et six constantes de temps moyennes de 0,25sec à 50sec. Chaque impulsion d'entrée produit une impulsion carrée à largeur stable qui, à son tour, force une charge de quantité fixe à pénétrer dans un condensateur d'emmagasinage. La tension qui en résulte est amplifiée dans un amplificateur à relais modulateur enfermé dans un cadre de Miller pour donner le temps d'intégration voulu. L'indication de sortie de l'instrument et des sorties d'enregistrement de 1mA et 100mV d'une précision de 2% peuvent être obtenus.

Un potentiomètre cartésien à courant alternatif et à auto-contrôle pouvant être utilisé dans la gamme de 100 à 10 000Hz par E. R. Wigan

Résumé de l'article aux pages 442 à 449

Un nouveau type de potentiomètre à courant alternatif cartésien, basé sur un jeu de réseaux à rayons cathodiques est décrit dans cet article. La théorie, la conception et les applications de cet appareil sont examinées par rapport à un instrument modèle.

L'auto-contrôle de l'appareil permet d'étalonner en termes absolus ces paramètres de performance. La précision de l'étalonnage peut être vérifiée en utilisant l'appareil pour déterminer les rapports de tension vectorielle dans quelques circuits à performance critique.

La gamme de fréquence s'étend par plots de 100Hz, de 100Hz à 10 000Hz et les interpolation entre ces plots peuvent être effectuées à l'aide d'un complément artificiel. Un pont de fréquenc fait partie du réseau cartésien et permet l'étalonnage de l'oscillateur d'entraînement connexe.

Bien que l'appareil ne soit pas de précision, il est capable d'effectuer des mesures d'une précisio de 0,5%, pouvant être améliorées jusqu'à 0,1% en utilisant une méthode appropriée de correctior

Aucune tentative n'a été faite jusqu'ici pour mesurer la tension en termes absolus, la fonctio primordiale de l'instrument étant de mesurer les rapports entre tensions alternatives. On peu cependant, réaliser des mesures d'une très grande précision.

Calculatrice analogique à transistors à grand signal par G. G. Bloodworth et D. N. Axford

Résumé de l'article aux pages 450 à 455

On a construit une calculatrice analogique qui déclenche le comportement du modèle de transisto normalement utilisé dans la réalisation des circuits, sauf que la fréquence de coupure effective est . échelle réduite. Le comportement électrique de la région de base est représenté par un réseau passı linéaire, tandis que les transformations nécessaires de tension non linéaire aux jonctions sont produite par les caractéristiques des diodes. Cette calculatrice peut être utilisée pour indiquer l'importanc des divers paramètres d'un transistor et pour prédire sa performance dans les applications de circuits

Intensimètre/Intégrateur linéaire Harwell série 2000 par J. W. Wilkins et D. J. Marsh

Résumé de l'article aux pages 456 à 459

En raison du degré supérieur de précision quelles assurent, les méthodes numériques on remplac les méthodes analogiques au cours des dernières années pour la plupart des mesures de comptag et de taux. Cet article décrit des perfectionnements de circuit par rapport aux systèmes analogique originaux, qui ont permit de réaliser des instruments de précision plus simples et à double fonctio utilisant une alimentation à basse tension.

Un appareil de mesure de la hauteur d'impulsion par C. I. Sach

Résumé de l'article aux pages 460 à 464

L'appareil décrit dans cet article mesure la valeur de pointe d'une impulsion à polarité négativ unique. Cet appareil avait été conçu pour la mesure d'impulsions à bord frontal négatif prononcé d'une ampleur variant entre un et cinq volts, suivie d'une diminution exponentielle jusqu'au retour zéro avec une constante de temps de deux secondes. La valeur du signal de pointe est indiqué avec une erreur inférieure à 5% sur un voltmètre à courant continu, à condition que la lecture so effectuée en moins de 10 secondes après l'événement. Les caractéristiques de l'appareil le render propre à la lecture de la valeur de pointe d'impulsions rectangulaires uniques d'une durée minima a 3 millisecondes.

Un générateur de faux bruits aléatoires à basse fréquence par C. Kramer

Résumé de l'article aux pages 465 à 467

L'auteur décrit un circuit qui produit des bruits dans la gamme subacoustique au moyen d'un en registreur de décalage à réaction. Les bruits sont quasi aléatoires et leur répartition est à amplitud bidirectionnelle. La spectre est plat dans la zone d'intérêt. Les caractéristiques statistiques sont pré visibles et d'une haute stabilité.

Un compteur de décades réversible avec codage 1242 symétrique par I. Scollar

Résumé de l'article aux pages 468 à 476

Un compteur de décades binaires utilisant un code chargé symétrique est décrit dans cet article Ce dispositif surmonte certaines des difficultés inhérentes au compteur réversible précédemment décri par l'auteur. Le compteur peut être aisément assemblé à l'aide de modules logiques et il peu enregistrer des vitesses s'élevant à peu près à la moitié de celles des binaires employées.

Linéarisation harmonique de l'oscillateur à pont de Wien par M. A. Kaaz

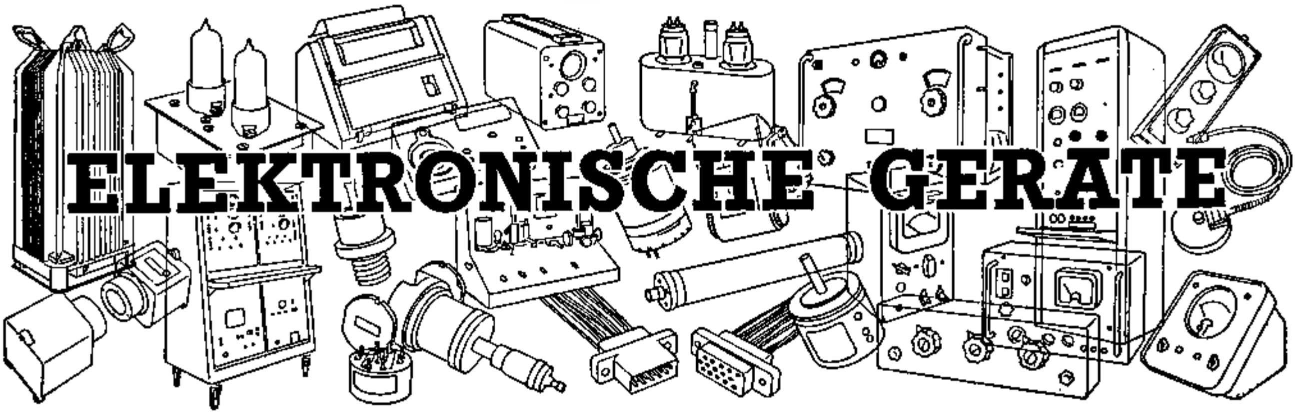
Résumé de l'article aux pages 470 à 471

Afin d'assurer la stabilité du fonctionnement d'un oscillateur, il faut qu'il y ait une certaine non linéarité dans le système. Cette non-linéarité doit être suffisante pour permettre un contrôle auto matique approprié de l'amplitude, mais il faut également qu'elle ne tolère qu'une très faible dis torsion harmonique. L'auteur analyse cet aspect de l'oscillateur à pont de Wien utilisant l concept de la fonction descriptive.

Une méthode perfectionnée pour la mesure photoélectrique par Matthews, Faulkner et Green

Résumé de l'article aux pages 472 à 474

La mesure des seuils photoélectriques exige parfois la mesure de courants photoélectriques allan jusqu'à 10^{-16} ampères. Cette mesure s'effectue le mieux en interrompant la radiation d'excitatior incidente et en utilisant un système à courant alternatif où l'on obtient une largeur de bande trè. étroite en utilisant un redresseur sensible aux phases. Cet article décrit en détail la manière d monter un tel système à courant alternatif avec des composants électroniques vendus dans le commerc et il décrit aussi les avantages de cette méthode par rapport aux dispositifs classiques à électromètr à courant continu.



ELEKTRONISCHE GERÄTE

Beschreibung neuer Bauelemente, Zubehörteile und Prüfgeräte auf Grund der von Herstellern gemachten Angaben.

Übersetzung der Seiten 478 bis 483

Schrittschaltermodul

Flight Refuelling Ltd, Wimborne, Dorset

(*Abbildung Seite 478*)

Flight Refuelling Ltd, Wimborne, Dorset

Die Industrial Electronics Division der 'light Refuelling Ltd kann jetzt den chrittschalter RSC-41 liefern, der im irunde ein Schnelldrehwähler ist, dessen chaltkontakte aus Zungenschaltern be- ;ehen. Er wird in Modultechnik her- estellt, und jeder Modul ist etwa 41 mm ing, 22,9 mm tief und 27,8 mm hoch; em Konstrukteur wird damit die grösste ,npassungsfähigkeit mit den geringsten 'erdrahtungsschwierigkeiten geboten.

Jeder Modul ist eigentlich eine Stufe ines Schrittschalters, und da die Anzahl er zusammengeschalteten Moduln prak- sch unbegrenzt ist, kann der Kon- trukteur einen Schrittschalter mit enauso vielen Schritten bekommen wie r braucht. Andererseits können zwei Moduln zusammen einen wahren bi- tabilen Modul bilden.

Die Stiftpositionen im Sockel jedes Moduls entsprechen dem 2,54 mm- Normraster für gedruckte Schaltungen nd sind so angeordnet, dass alle Ver- indungen zwischen Moduln durch irekte Leiter erstellt werden können.

Obwohl der Modul RSC-41 in erster ,inie zur Deckung des Bedarfs an chrittschaltern gedacht war, bestehen uch für andere Schaltzwecke Anwen- ungsmöglichkeiten.

EE 83 751 *für weitere Einzelheiten*

Sende-Empfangsweiche ohne Vorionisation

Ferranti Ltd, Kings Cross Road, Dundee, Schottland

Ferranti Ltd hat eine neue Sende- :mpfangsweiche JF.20 für das S-Band ntwickelt. Die als Überlastungsschutz ür Wanderfeldröhren und parametrische Verstärker in Impulsradarsystemen kon- struierte Weiche hat eine radioaktive Füllung, und die externe Speisespannung für die Vorionisation ist daher über- flüssig. Das Stehwellenverhältnis liegt innerhalb der Grenzfrequenzen 2 700 und 3 100 MHz unter 1,25 und eine typische Einfügungsdämpfung bei 0,3 dB.

Bei Einsatz von Detektorköpfen gibt die JF.20 in Verbindung mit einer weiteren vor kurzem bekanntgegebenen Entwicklung — dem Varactordioden- begrenzer 104 SSS—ausreichenden Schutz (Überschwingspitze $\simeq$0,07 Erg/Impuls) für eine Einfügungsdämpfung von unter 1,0 dB und ohne irgendwelche externe Stromversorgungen.

Selbst wenn das Radar nicht in Betrieb ist, wird der Detektorkopf durch die Kombination JF.20/104 SSS ge- schützt, und es ist keine mechanische Hohlleiterblende erforderlich.

Parametrische Verstärker sollen zwischen die Weiche JF.20 und den Begrenzer 104 SSS geschaltet werden, so dass die Weiche den Verstärker mit der niedrigsten Einfügungsdämpfungsbela- stung schützt, während der Begrenzer den durch den Verstärker auf den Kristall- diodenmischer rückwirkenden Durch- schlag auf einen zulässigen Pegel reduziert.

Die JF.20 ist für eine Spitzenleistung von 100 kW und eine mittlere Leistung von 100 W bemessen; wo Betrieb mit höheren Leistungen erwartet wird, muss ein Ferritzirkulator oder Gasduplexgerät, z.B. der Ferranti-Baureihe 10D, vorge- schaltet werden.

EE 83 752 *für weitere Einzelheiten*

Funktionsverstärker

Fenlow Electronics Ltd, Springfield Lane, Weybridge, Surrey

(*Abbildung Seite 478*)

Der A7 ist ein mittels Fotowiderstand stabilisierter Transistorverstärker, dessen Ausgangsbereich dem von Röhrengeräten vergleichbar ist. Gleichzeitig werden jedoch Drift- und Rauscheigenschaften von Verstärkern erreicht, die mit viel niedrigeren Pegeln arbeiten, was eine wesentliche Erhöhung der Rechen- genauigkeit gestattet.

Der Verstärker wurde für Anwendung in allen Standard-Analog- und Hybrid-Rechnern und in der Instrumentierung entworfen. Seine hohe Ausgangsleistung gestattet direkten Einbau in ziemlich grosse Servosysteme.

Der völlig kurzschlussfeste Verstärker ist durchgehend mit Siliziumtransistoren und -dioden bestückt.

Bis zu $\pm$100 V bei 50 mA Ausgang arbeitet der Verstärker linear und gibt bis zu 5 kHz volle Ausgangsleistung. Die Gleichstromverstärkung ist grösser als 10×10^6 und das Produkt Ver- stärkung $\times$ Bandbreite 10 MHz. Der Eingangswiderstand ist 100 kΩ, die Rauschspannung niedriger als 50 μV_{ss} und die Drift niedriger als 1 μV/°C.

EE 83 753 *für weitere Einzelheiten*

Kabelkanal

Insuloid Manufacturing Co. Ltd, Leestone Road, Wythenshawe, Manchester

Ein neues Kabelkanalprogramm mit viel zeit- und geldsparenden Konstruk- tionsmerkmalen wurde von Insuloid Manufacturing Co. Ltd bekanntgegeben. Der Kanal ist durchgehend längsweise geschlitzt, und da jeder Schlitz eine enge Öffnung hat, lassen sich an jedem be- liebigen Punkt Kabel herausführen. Diese kleinen Öffnungen sind so kon- struiert, dass sie die Kabel während des Verlegens am Losbrechen verhindern. Im Verlauf können Kabel durch Schnappbrücken voneinander getrennt werden, und der Deckel wird mit einer selbstverankernden Schnappbefestigung aufgesetzt. Insuloid Kabelkanäle haben Rillen in der Unterfläche, um schnelles und genaues Bohren zu erleichtern; das Profil ist glatt und gestattet Parallelstapeln. Der Kanal ist in vier Grössen und mit drei verschiedenen Schlitzpositionen lieferbar; er wird aus leichtem Hart-PVC gefertigt, ist daher korrosionsfest und erfordert nach Installation keinerlei Wartung.

EE 83 754 *für weitere Einzelheiten*

Kontrollwerkzeuge für Relais

P. W. Allen & Co., 253 Liverpool Road, London, N.1

(Abbildung Seite 479)

Ein neuer und einzigartiger Werkzeugsatz für die Untersuchung von Kontakten und Schleifern von Fernsprecher-Schrittschaltern wurde von P. W. Allen & Co. angekündigt.

Der mit Allen-Typ TS bezeichnete Satz besteht aus einem Handgriff mit Batterie, drei Lichtsonden und kleinen Spiegelansätzen.

Mit diesem Satz kann man—gegebenenfalls mit angesetztem Spiegel—Kontakte direkt beleuchten und, selbst wenn sie in Mehrfachreihen hinten liegen, die Kontaktfläche genau beobachten.

Der Satz ist für die Kontrolle von Kontakten und Schleifern sowie dem Mechanismus von Drehwählern, Hebdrehwählern und Relais geeignet. Fernsprechingenieure haben diesen Satz aus dem breiten Sortiment von Allen-Lichtsonden zusammengestellt, da es der wirtschaftlichste Satz ist, der allen normalen Kontrollansprüchen in Fabrik und Fernsprechamt genügt.

Die kleinen Sonden haben einen Durchmesser von etwas über 3 mm, und die Spiegel haben dieselben Abmessungen. An die Sonden sind ×2- oder ×3-Vergrösserungsgläser angebaut und geben die für Nahbesichtigung geeignete niedrige Vergrösserung.

EE 83 755 *für weitere Einzelheiten*

Datenerfassungs- und Registrierausrüstung

Hawker Siddeley Dynamics Ltd, Whitley Works, Coventry, Warwickshire

(Abbildung Seite 479)

Die vollintegrierte Konstruktion dieses neuen Datenerfassungs- und Registriersystems für Schiffahrt, Industrie und Wissenschaft ermöglicht Einsatz für einfaches Protokollieren oder Ausbau in ein hochentwickeltes Datenerfassungs- und Regelsystem mit prozessgekoppelter Datenverarbeitung.

Bis zu 1000 hoch- oder niederpeglige Analogsignale können mit einer Geschwindigkeit von 20 Eingaben pro Sekunde abgefragt werden. Durch unabhängigen Schutz jedes Eingangs und den nach neusten Kenntnissen entwickelten Analog-Digitalumsetzer (Patent angemeldet) wird der Einfluss von Gleichtaktstörsignalen innerhalb niedriger Grenzen gehalten.

Die Ausrüstung ist für Dauerbetrieb von mindestens sechs Monaten ohne Nacheichung ausgelegt. Selbsteichung erfolgt nach jedem Zyklus, und für jedes Eingangssignal kann man unterschiedliche und getrennte Toleranzen eingeben.

Zyklen können entweder laufend wiederholt oder manuell bzw. in von einem Taktgeber gesteuerten Intervallen einmal abgerufen werden.

Eingangssignalwerte und -kennung, Alarmzustände und Zeitangabe lassen sich in verschiedenen Formaten ausdrucken und durch Sichtanzeige darstellen, wobei gleichzeitig ein redigierter Lochstreifen erstellt werden kann.

EE 83 756 *für weitere Einzelheiten*

Flächenprofilmonitor

G. V. Planer Ltd, Windmill Road, Sunbury-on-Thames, Middlesex

(Abbildung Seite 479)

Der Flächenprofilmonitor wurde für die Darstellung der Beschaffenheit von Oberflächen entwickelt und ist für die Dickenmessung von Dünnfilmen im Bereich 0...50 000 Å geeignet.

In der Ausrüstung gibt ein Messwertwandler die Bewegung eines Abtastfühlers in bezug auf einen Schlitten, die beide die auszuwertende Fläche durchlaufen; es sind dafür zwei wahlweise einstellbare Geschwindigkeiten vorgesehen. Das Messresultat wird entweder durch den Zeiger des Registriergerätes angezeigt oder nach Wahl auf einem Streifen aufgezeichnet, von dem die Dicke des Films mit einer Unsicherheit von unter 100 Å bestimmt werden kann.

Die Ausrüstung ist besonders für die Dickenmessung von im Vakuum aufgedampften Schichten sowie für die Überwachung der Oberflächenbeschaffenheit von Materialien geeignet.

Die Ausrüstung besteht aus zwei Geräten, und zwar dem Detektorgerät und dem Kontrollgerät, die beide in Stahlgehäusen untergebracht sind.

EE 83 757 *für weitere Einzelheiten*

Doppelimpulsgeber

Venner Electronics Ltd, Kingston By-Pass, New Malden, Surrey

(Abbildung Seite 479)

Venner Electronics Ltd hat einen durchweg mit Silizium-Halbleitern bestückten Festkörper-Doppelimpulsgeber TSA 628 eingeführt, der bei Einsatz als Einzelimpulsgeber einen Frequenzumfang von 2,5 Hz...2,5 MHz und als Doppelimpulsgeber von 2,5 Hz ... 2 MHz hat. Drei Betriebsarten sind möglich: frei laufend, extern durch einen Impuls von 2...50 V oder Kontakte getriggert, oder drucktastengesteuert.

Mittels eines Stufenabschwächers können Ausgänge von 20, 10, 5, 2, 1 und 0,5 V eingestellt werden, und ein Feinregler gestattet Einstellung auf ein Minimum von 0,2 V. Bei 20 V ist die Ausgangsimpedanz 100 Ω, sonst 50 Ω.

In bezug auf Masse wahrhaft positiv und negative Impulse können eingestell werden. Die Anstiegs- und Abfallzeite sind 10 ns für die Vorderflanke und 1 ns für die Hinterflanke, und die Impuls dauer- und Verzögerungsschaltunge sind zwischen 100 ns und 10 ms kon tinuierlich regelbar. Der Vorimpuls ha bei Leerlauf eine Mindestausgangsspar nung von ±5 V wechselstromgekoppe bei 100...150 ns Dauer.

EE 83 758 *für weitere Einzelheite*

Mikrowechselspannungsmesser

Levell Electronics Ltd, Park Road, High Barne Hertfordshire

(Abbildung Seite 479)

Dieses tragbare Instrument kann ma als Wechselspannungsmesser oder -ve stärker einsetzen. Der Frequenzbereic ist 1 Hz...3 MHz. Es sind 16 Bereiche i 10-dB-Stufen mit Skalenendwerten vo 15 μV bis zu 500 V vorhanden, di durch einen Einzeldrehschalter, der auc Positionen für "Aus" und "Speisespar nungsanzeige" hat, eingestellt werden.

Das Gerät hat zwei getrennte Gegen kopplungsverstärker. Die Verstärkun des ersten wird in 10-dB-Stufen bis z 80 dB variiert, wobei die Verstärkun durch Gegenkopplung stabilisiert wir Der Ausgang liegt an Buchsen auf de Frontplatte und am Eingang des zweite Verstärkers, der einen Vollweggleic richter und ein Drehspulenmessger treibt. Die Gegenkopplung dieses Ve stärkers ist so angeordnet, dass der durc das Messgerät fliessende Strom zur Ei gangsspannung in linearer Beziehun steht. Vollausschlag wird durch eine Au gangsspannung des ersten Verstärke von 150 mV hervorgerufen. Der Ei gangswiderstand des zweiten Verstärke ist so hoch, dass er für den Ausgang de ersten Verstärkers eine vernachlässigba niedrige Last darstellt und das Signa an den Ausgangsbuchsen von auf di Gleichrichterbelastung zurückzuführe den Verzerrungen frei ist. Ein weitere Vorteil dieser Anordnung ist, dass di Bandbreite des Gerätes durch Anlege von Blindwiderständen an die Ausgang klemmen verringert werden kann. D Herabsetzung der Bandbreite reduzie auch den Eingangsrauschpegel, was b sonders beim Messen kleiner Signal über einen beschränkten Frequenzbereic sehr nützlich ist.

Das Gerät ist als Vorverstärker f einen Oszillografen ideal, da es eine sehr niedrigen Rauschpegel, gering Klingfähigkeit und kein Brummen au weist und von der Netzversorgung u abhängig ist.

EE 83 759 *für weitere Einzelheite*

Werkstofftester

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

(Abbildung Seite 480)

Der dynamische Werkstofftester 182 ist in erster Linie für Vorführungszweck in Lehranstalten bestimmt. Er mis

lastizitätsmodul, Q und Dämpfungs-
ktor zur Resonanz erregter Prüfstäbe.

Die Ausrüstung besteht aus Oszillator, ufwärtstransformator und Gleichstrom-ırmagnetisierungsquelle, zusammen mit :äzisionsprüfbank, Verstärker und szillografen. Ein gleichförmiger Stab ier eine Stange aus dem zu prüfenden 'erkstoff wird in der Mitte in die ·üfbank geklemmt, so dass zwei kleine ıftspalte (20...50 μm) die freien Enden ın den elektrostatischen Köpfen der ·üfbank trennen. Der mit einem Wech-lfeld beaufschlagte Treiberkopf erregt :n Prüfling zur Resonanz, und das eine dann am Abtastkopf auftretende 'echselstromsignal wird über den Ver-ärker dem Oszillografen oder einem ıderen geeigneten Nachweisgerät zuge-hrt.

Mit einem zusätzlichen Gerät Typ ı25 kann die Ausrüstung auch für dustrielle Messungen eingesetzt werden, doch wird nicht behauptet, dass sie r alle technischen Zwecke genau :nug ist. Mit Hilfe der Frequenzskala :s Oszillators kann man den Elastizi-tsmodul mit annehmbarer Genauigkeit, it einem externen Frequenzmesser mit össerer Genauigkeit bestimmen. Bei :üflingen mit niedrigem Q kann man esen Wert mit dem Oszillator, für lche mit mittlerem Q-Wert mit einem .ternen Frequenzmesser ermitteln. Für ·üflinge mit hohem Q muss man ein ıderes Verfahren mit logarithmischen ekrementen anwenden, und für solche ·üflinge wird eine Sonderausführung :s Gerätes angeboten. Der Oszillator ıd der Oszillograf sind in sich ge-hlossene Moduln und können daher ıabhängig für andere Aufgaben ein-setzt werden, was für Laboratorien in chnischen Lehranstalten mit knappem .atz oder geringem Einkommen von osser Bedeutung ist. Die Standardaus-ıstung wird mit einem Einstrahloszillo-afen geliefert, auf dessen Schirm ssajousche Figuren dargestellt werden. 'eitere Varianten mit Zweistrahloszillo-afen oder ohne Oszillografen für nsatz mit externen Oszillografen oder öhrenvoltmetern für Anzeige sind doch auch lieferbar.

Die Standardprüfbank kann Prüflinge ın 150...280 mm Länge und 6...15 mm urchmesser aufnehmen; für die Auf-ıstung, die weniger als eine Minute ıuert, gibt es eine Spezialvorrichtung. ls Oszillator dient ein 6-W-Hochlei-ıngsmodell mit einem Frequenzbereich ın 20 Hz...20 kHz. Die Abmessungen :r elektronischen Ausrüstung sind etwa !5 × 432 × 584 mm hoch.

EE 83 760 *für weitere Einzelheiten*

Prozentsatzquotientenmesser

Bendix Electronics Ltd. High Church Street, New Basford, Nottingham

(Abbildung Seite 480)

Der Prozentsatzquotientenmesser be-:ht aus einer Reihe von Geräten, die e Prozentsatzdifferenz zwischen zwei Arbeitsvorgängen, z.B. Walzengeschwindigkeiten, Wellengeschwindigkeiten usw., mittels Digitalzählverfahren bestimmen und anzeigen. Das Gerät ist besonders beim industriellen Kalandrieren nützlich, wo prozentuales Recken und Schrumpfen gemessen werden muss. Differenzen von 0,01 Prozent sind ohne Schwierigkeiten nachweisbar.

Das Gerät besteht im Grunde aus zwei Zählern, an die gleichzeitig Eingangssignale gelegt werden; einer der Zähler unterbindet beide Eingänge, sowie 1000 Impulse (oder eine andere vorbestimmte Zählbasis) gezählt sind. Die Zähler sind so angeschlossen, dass die Zählung im zweiten 1000 oder mehr ist (natürlich je nach der gewählten Zählbasis). Vorgezogen wird als Eingangsmesswandler der Rotapulse 191A oder der Impulsgeber 491A. Der Zähler ist mit geeigneten Eingangsschaltungen für andere Impulsgeber lieferbar.

Wenn der gemessene Quotient ausserhalb der Grenzwerte 0... ±9,9(9) Prozent liegt, wird eins der beiden Vorzeichen gezeigt und die Ziffernanzeige unterbunden. Für Ergebnisse unter dem unteren Grenzwert erscheint "−", für solche über dem oberen Grenzwert "+".

EE 83 761 *für weitere Einzelheiten*

Schaltungsmoduln

Morecombe Electrical Equipment Co. Ltd. Westgate Works, Morecombe, Lancashire

(Abbildung Seite 480)

Das Fertigungsprogramm "Meecostatic" für transistorisierte Logik-, Eingangs-, Ausgangs- und Schaltungsmoduln für industrielle Steuerung wurde durch folgende neue Moduln erweitert: einen RC-Zeitgeber, eine Decodiermatrix für Binär-Dekadenentschlüsselung, eine Torschaltung für zwei Eingänge, einen Sägezahngenerator, umkehrbaren Zähler und einstellbare Triggerpegelmoduln.

Mit diesen zusätzlichen Typen gibt es jetzt in diesem Programm 35 Moduln, die alle elektrisch und mechanisch miteinander kompatibel sind.

Die Abbildung zeigt einen Teil des für diese Moduln als Einschub für 19"-Gestelle entwickelten Systems. Jeder Abschnitt kann bis zu fünf 3" (76 mm) breite Chassis, oder eine entsprechend kleinere Anzahl von 6" breiten Chassis, oder eine Kombination von sechzig 1/2"-(12,7 mm)-Moduln aufnehmen und ist als Einschub in den Rahmen ausgelegt. Die Konstruktion besteht hauptsächlich aus hochpoliertem Aluminium, dessen sichtbare Frontflächen schwarz eloxiert und mit verchromten Handgriffen zum Herausziehen versehen sind. Andere Aluminiumflächen sind mattiert. Die Abbildung zeigt, wie man Chassis in Verbindungseinheiten einschieben kann, so dass sie für Tests stromführend sind und aus dem Hauptgerät herausstehen.

Vorn am Chassis ist ein Tastkopf zu sehen, der mit einer kleinen Anzeigelampe verbunden ist, so dass jeder der Moduln in einem Abschnitt daraufhin geprüft werden kann, ob er ein- oder ausgeschaltet ist.

Ausserdem wurde das Programm durch eine Serie von Stromversorgungen für Ströme von 0,5 A, 1 A und 2 A ergänzt. Diese Bausteine geben eine Konstantspannung von −24 V ab. Das Regelverhältnis ist grösser als 5000:1, der Gleichstromausgangswiderstand unter 0,002 Ω und die Ausgangsimpedanz bei 100 kHz niedriger als 0,1 Ω. In diesen Stromversorgungen gibt es auch eine +24 V Speisespannung, und beide haben einen gemeinsamen Nullpunkt. Diese positive Speisespannung hat eine einfache Zenerregelung und wird hauptsächlich für die Vorspannung verwendet.

EE 83 762 *für weitere Einzelheiten*

Mikrowellengeräte

G.E.C. (Telecommunications) Ltd, Telephone Works, Coventry

(Abbildung Seite 481)

Die Fertigung eines neuen Programmes durchweg mit Halbleitern bestückter Mikrowellen-Funkgeräte ist jetzt bei G.E.C. (Telecommunications) Ltd angelaufen. Bisher haben mit Halbleitern bestückte Mikrowellengeräte in Teilen ihrer Schaltung Glühelektronenröhren enthalten, üblicherweise Wanderfeldröhren, Klystrons oder Mikrowellentrioden in der Leistungsendstufe. G.E.C. stellt jetzt erfolgreich ausschliesslich mit Halbleitern bestückte Geräte her. Die beiden Hauptvorteile sind erhöhte Zuverlässigkeit und wesentlich geringerer Aufwand für die Stromversorgungen selbst im Vergleich mit anderen Geräten, die nur in einer Stufe Röhren enthalten. Beide Vorteile haben besondere Bedeutung in Ländern mit entfernt liegenden, schwer zugänglichen Stationen, wo Stromversorgung und Wartung auf ein Mindestmass beschränkt werden müssen.

Zwei G.E.C.-Ausrüstungen sind in laufender Produktion, von denen eine für Hauptfernverkehr geeignet und für 960 Sprechwege oder einen Fernseh- und zugehörigen Tonfrequenzkanal in den Frequenzbändern 1700...1900 MHz oder 1900...2300 MHz ausgelegt ist. Die 2-W-Ausgangsleistung des Senders dieser Ausrüstung wird ausschliesslich mit Halbleitern erzielt; der Empfänger hat eine rauscharme Eingangsstufe mit Tunneldiodenverstärker und einen Rauschfaktor von 5½ dB. Zur Breitbandausrüstung gehört ein durchweg mit Halbleitern bestückter Hilfsfunkkanal und ein Mehrfach-HF-Wählersystem mit Kanalschutz.

Die zweite Ausrüstung arbeitet im Frequenzband 7425...7725 MHz mit einer Kapazität von 300 Sprechwegen und ist hauptsächlich für kleine Fernsprechnetze oder als Abzweigverbindung für Breitbandstecker grosser Kapazität geeignet. Ein Doppelweg-Wechselverkehr-Verstärker ist komplett mit Schalteinrichtung und Fernalarmeinrichtung in einem Gestell untergebracht. Ein nützliches Merkmal dieser Ausrüstung ist die Möglichkeit, einen beliebigen Teil des Grundbandes fallen zu lassen und ihn in einer beliebigen Verstärkerstation ohne Demodulation der Durchgangswege wieder einzuschalten.

EE 83 763 *für weitere Einzelheiten*

Miniatur-Stromstossrelais

Vertrieb: B & R Relays Ltd, Temple Fields, Harlow, Essex

(*Abbildung Seite 481*)

B & R Relays Ltd vertreibt Siemens & Halske-Miniaturrelais in Grossbritannien und hat nunmehr das neue magnetische Siemens-Stromstossrelais Typ P auf dem britischen Markt eingeführt, das in zwei Grössen kommt. Die kleinere Ausführung hat zwei Schwachstrom-Umschaltkontakte; die grössere hat vier Umschalter, sechs Arbeitskontakte und sechs Ruhekontakte oder zwei Umschalter mit Hochleistungskontakten. Für die meisten Ausführungen sind Silber- oder Goldlegierungskontakte lieferbar. Das Relais Typ P (Katalog-Nr. V23003) wird durch einen Impuls betätigt.

Das mit einer Spule ausgerüstete Relais kann erregt oder impulsbetätigt werden—es wird dann betätigt und bleibt in diesem Zustand, bis die Spule in entgegengesetzter Richtung erregt wird.

Auf Wunsch ist das Relais auch mit zwei Spulen lieferbar; nach Erregung der einen Spule spricht es an, nach Erregung der anderen fällt es ab.

EE 83 764 *für weitere Einzelheiten*

Tastenschalter

Standard Telephones & Cables Ltd, Electro-Mechanical Division, West Road, Harlow, Essex

(*Abbildung Seite 481*)

Der elektromechanische Fachbereich der STC kann nunmehr die von Mayr, einer STC-Beteiligungsgesellschaft, hergestellten Tastenschalter liefern. Diese Bausteine bieten dem Verfahrensingenieur eine für diese Bausteintype ungewöhnliche Vielseitigkeit der Schaltmöglichkeiten. Die Anpassungsfähigkeit des Systems ermöglicht Kombinationen der Tastenzustände "Halten" und "Löschen". Es gibt Einzelauslösung und gegenseitige Auslösung, auf Wunsch gegenseitige Auslösung in Gruppen oder mit gemeinsamer Löschung.

Ausser den verschiedenen Mehrfachzusammenstellungen von Halte- und Löschtasten, die ihrerseits jeweils verschiedene Schaltfunktionen ausführen können, ist Einbau von "Impulstasten" in denselben Mayr-Baustein möglich. Dadurch lässt sich eine grosse Anzahl elektrischer und mechanischer Kombinationen erreichen. So können z.B. in dasselbe Tastenschalterchassis Schalter unterschiedlicher Konstruktion für hohe und niedrige Leistungen eingebaut werden.

Die meisten Typen sind in Streifen von 12 Tasten lieferbar. Die Konstruktion gestattet sehr enges vertikales und horizontales Stapeln der Tastenstreifen. Die Knöpfe haben abnehmbare transparente Kappen für das Einlegen von Farbsignalen oder geschriebenen Zeichen zur Kennzeichnung.

EE 83 765 *für weitere Einzelheiten*

Frequenzwächter

Kent Precision Electronics Ltd, Vale Road, Tonbridge, Kent

(*Abbildung Seite 481*)

Kent Precision Electronics Ltd hat vor kurzem eine Festkörperschaltung entwickelt, die in den mechanischen Zusammenbau der Standardausführung ihres geeichten Festkörperkontrollgerätes für Regelung, Schutz oder Warnung in Frequenzüberwachungsanlagen eingebaut wird.

Das System besteht im Prinzip aus einem Frequenz-Gleichstromwandler, der ein der Frequenz über einen kleinen Bereich—z.B. 45...55 Hz—genau proportionales Gleichstromsignal abgibt. Die Eingangsfrequenz wird anfänglich in einen Impulsausgang derselben Frequenz umgeformt und in einen linearen Integrator gespeist, dessen Ausgang im Eingang eines geeichten KPE-Standard-Festkörperkontrollgerätes liegt. Der zusätzliche Frequenz-Gleichstromwandler ist als getrennte Druckschaltungskarte in einem Messgerätgehäuse Typ "Elite" ausgeführt und wird vom Netzgerät des übrigen Gerätes gespeist. Das Gerät wird normalerweise an das Wechselstromnetz angeschlossen und kann je nach Wunsch dessen Frequenz oder die eines zweiten Eingangs überwachen.

Einzel- oder Doppelgeräte sind als Standardausführungen lieferbar, wobei das Doppelgerät mit konzentrischen Bedienelementen ausgelegt ist. Jedes Gehäuse enthält ein internes Relais, dessen Kontakte für 5 A, 240 V~ bemessen sind und das auf Betätigung beim vorgewählten Punkt eingestellt ist. Andererseits kann der Ausgang aus proportionalen oder Ein-Aus-Thyristorelementen bestehen.

EE 83 766 *für weitere Einzelheiten*

Zerlegbarer Gleichstromlaser

Scientifica, 148 St. Dunstan's Avenue, Acton, London, W.3

(*Abbildung Seite 482*)

Die Entwicklung des Scientifica-Gleichstrom-Gaslasers ist das Ergebnis der Bemühungen, ein zerlegbares Instrument herzustellen, das die für Forschungsarbeiten erforderliche Präzision und Stabilität besitzt, jedoch robust genug für allgemeinen Laboreinsatz ist. Die Konstruktion gestattet Verwendung als freistehendes Instrument in beliebiger Orientierung oder in Form einer optischen Bank mit Resonatorspiegeln, die entweder im Laser montiert verbleiben oder von getrennten Stützen der optischen Bank getragen werden. Die[s] vielseitigen Anordnungsmöglichkeite[n] gestatten, Versuche mit verschiedene[n] Spiegelabständen auszuführen, was ein[e] eingehende Untersuchung der stabile[n] und unstabilen Resonatoranordnunge[n] sowie das Anbringen von optische[n] Standardelementen im Hohlraumreson[a]tor ermöglicht.

Der Spiegel hat einen 17schichtige[n] harten Überzug, den man mit normalem Linsenzelltuch reinigen oder in Alkoh[ol] waschen kann, und die speziell en[t]wickelte Gleichstrom-Gaslaserröhre h[at] eine ungewöhnlich lange Lebensdaue[r]. Die Laserröhre gibt kontinuierlich i[m] roten und infraroten Gebiet des Spek[t]rums linear polarisierte ungedämpft[e] Wellen ab; typische Ausgangsleistunge[n] sind 2 mW bei 6328 Å und 10 mW b[ei] 11 530 Å.

Nur ein Kabel verbindet den Laser m[it] einem Treibergerät, an dem Regler fü[r] die Einstellung der Lichtabgabe vorhan[]den sind; der Erregerpegel wird au[f] einem 50-mA-Messgerät angezeigt. Di[e] Ausrüstung wird mit 110 V~ oder 23[0] V~ betrieben und hat eine niedrige Res[t]welligkeit des Gleichstromausgangs, di[e] weniger als 1 Prozent Modulation de[s] Laserstrahls hervorruft. Die Bauelement[e] im Laser und Treibergerät sind reichlic[h] bemessen, von hoher Qualität und ohn[e] Vorbehalt für ein Jahr garantiert.

Mit jedem Laser wird ein Handbuc[h] mitgeliefert, in dem die mit dieser neue[n] Quelle kohärenten Lichtes ausführbare[n] Versuche eingehend beschrieben werden. Es handelt sich sowohl um physikalisch[e] als auch geometrische Versuche wie z.B. die Fraunhofersche Beugung, Abbe Theorie, auf einen verstellbaren Spieg[el] zurückzuführende Doppler-Verschiebung, Kollimation des Laserstrahls, durch di[e] der Strahl sich nur um wenige Zent[i]meter pro Kilometer erweitert, usw.

Mit dem monochromatischen Hoch[]leistungsstrahl des Gerätes wird da[s] Rüsten für einen Interferenzversuc[h] wesentlich vereinfacht. Ausserdem ist di[e] Intensität der erzielten Interferenzbilde[r] mehrere Grössenordnungen höher, al[s] die der mit älteren Methoden erzielbaren und daher viel eindrucksvoller. Solch[e] Versuche lassen sich also in nicht ver[]dunkelten Räumen durchführen, un[d] ausserdem kann eine grosse Hörerschaf[t] die Interferenzwirkungen sehen.

EE 83 767 *für weitere Einzelheite[n]*

Mikroschalterrelais

Keyswitch Relays Ltd, 120-132 Cricklewood Lane, London, N.W.2

(*Abbildung Seite 482*)

Ein auf den Standardrelaistypen 300[0] und 600 der britischen Postbehörde be[]ruhendes Programm von Mikroschalter[]relais wurde von Keyswitch Relays Lt[d] speziell für Temperaturregler un[d] -fühler eingeführt. Besondere Eigen[]schaften sind u.a. hohe Kontaktbelast[]barkeit, niedrige Betätigungskräfte un[d] hohe Schaltgeschwindigkeit (Schnapp[]kontakte). Die Arbeitskontakte sind fü[r]

Betrieb in gefährdender Atmosphäre völlig eingekapselt, und es können bis zu 20 A bei 480 V~ geschaltet werden. Ein Energieübertragungsverhältnis von 1:100 000 ist erreichbar.

Drei Grundtypen sind lieferbar: GPO 3000 mit Mehrfach- und Einzelschalter, sowie GPO 600 mit Einzelschalter. Betriebsspannung der Typen 3000 ist von unter 1 V—... 300 V— und mit zusätzlichen Teilen bis zu 440 V~, die der Typen 600 von unter 1 V—...175 V— und bis zu 440 V~.

Der Gleichstromwiderstand ist von unter 1 Ω...50 kΩ für die Typen 3000 und von unter 1 Ω...10 kΩ für die GPO-Typen 600.

Unter den lieferbaren Abwandlungen der Standardtypen 3000 ist die Ausführung mit Doppelanker für getrennte Betätigung des Mikroschalters zu erwähnen.

Typ 3000 mit Einzelschalter und Typ 600 können mit einstellbarer Vormagnetisierung ausgerüstet werden, beide Typen 3000 mit manueller Rückstellverriegelung. Spulen sind auf Wunsch tropenfest lieferbar, die Anschlüsse für Schraub- oder Lötverbindung.

EE 83 768 *für weitere Einzelheiten*

Mehrfachsteckverbindungen

Oxley Developments Ltd, Priory Park, Ulverston, Lancashire

(*Abbildung Seite 482*)

Oxley Developments Ltd hat einen neuartigen Typ von Steckverbindungen für Chassismontage eingeführt. Sie haben 24 vergoldete Einzelkontakte und eignen sich wegen der Konstruktion ihrer Führung vor allem für Einschübe in Gestelle und Pulte. Die Steckverbindungen haben eine durch Schrauben gehaltene mechanische Verriegelung, die gleichzeitig Druck auf die Kontakte ausübt, so dass selbst für lange Zeit unter rauhen Klimabedingungen ein niedriger Kontaktwiderstand aufrechterhalten wird.

EE 83 769 *für weitere Einzelheiten*

Bohrer für Druckschaltungen

Drill Service (Horley) Ltd, 89 Albert Road, Horley, Surrey

Nahezu alle Druckschaltungsplatten werden aus Werkstoffen hergestellt, die normale Schnellarbeitsstahlbohrer in kurzer Zeit abnutzen. Dabei wird einmal die Spitze sehr schnell abgestumpft, und zum anderen werden die zylindrischen Schneidenrücken abgerieben, so dass Untermasslöcher entstehen.

Starbide-Vollhartmetallbohrer für Leiterplatten wurden speziell für das Bohren von Druckschaltungsplatten entwickelt. Die Verwendung von Hartmetall beseitigt den Abrieb des Schneidenrückens völlig und gibt auch der Bohrerspitze eine wesentlich längere Lebensdauer. Mit diesen Bohrern werden 60 000 Löcher zwischen Nachschliffen als ganz normal betrachtet.

Löcher in Schaltungsplatten müssen auch gratfrei sein; dafür muss man einen steilen Spanwinkel und eine steile Spirale haben. Entfernung der Späne bei den benutzten sehr hohen Vorschubgeschwindigkeiten erfordert einen sehr schwachen Steg.

Starbide-Bohrer für Druckschaltungsplatten haben eine sehr steile Spirale und einen schwachen Steg; die Nuten sind in Feinstbearbeitung geschliffen. Die Toleranz für den Durchmesser beträgt 0,0076 mm und ist erforderlich, wo Präzisionslöcher mit Toleranzen von 0,013 mm in Quantitäten hergestellt werden. Bei einer Nutenlänge von 12,7 mm ist die Gesamtlänge für alle Grössen 38,1 mm.

EE 83 770 *für weitere Einzelheiten*

Schwingungsmessgeräte

Associated Engineering Ltd, Cawston, Rugby, Warwickshire

(*Abbildung Seite 483*)

Associated Engineering Ltd hat ihr Programm für Weg- und Schwingungsmessgeräte ergänzt, und zwar durch drei elektronische Geräte für Einsatz mit den induktiven Subminiaturweggebern der Firma. Die Ausrüstung besteht aus einem konstanten Oszillator mit fünf umschaltbaren Teilbereichen zwischen 50 kHz und 150 kHz sowie einem Verstärker, der—mit dem Oszillator als Treiber—sechs sehr niederohmige Ausgänge bis zu 20 V_{SS} gibt. Diese beiden Geräte können sechs Messkanäle mit induktiven Messwandlern erregen; jeder der Kanäle arbeitet mit einem Nachweis- und Filtergerät zusammen, das für den Messwandler eine angepasste Last darstellt, geeigneten Demodulationseinrichtungen und einem speziell für die Trennung des modulierten Signals vom Träger geeigneten Filter.

Mit dieser Kombination wird eine Empfindlichkeit von 10 mV je Tausendstel Zoll (0,025 mm) Verschiebung erzielt; Bewegungen mit Frequenzen von 25 kHz können ohne messbare Verzerrung angezeigt werden.

Die Abbildung zeigt Beispiele für jedes der erwähnten drei Geräte und ausserdem vier induktive Messwandler in den zur Zeit gefertigten zwei Grössen.

EE 83 771 *für weitere Einzelheiten*

Digitalwortgeber

Digital Measurements Ltd, Salisbury Grove, Mychett, Aldershot, Hampshire

(*Abbildung Seite 483*)

Der Datapulse-Wortgeber 208 gibt eine anpassungsfähige Kontrolle über Datenformat und -inhalt für die Simulation schneller, mehrkanaliger Digitaldaten.

Das Gerät hat einen Taktfrequenzbereich von 5 Hz...5 MHz, 8-Bit-Parallelzeichen, 1 bis 16 Serienzeichen pro Datenblock plus Zeichen- oder Blockwiederholung, 6 V Ausgangsspannung an 50 Ω, und Rückkehr-zu-Null-Formate oder Ohne-Rückkehr-zu-Null-Formate.

In der Parallelarbeitsweise gestatten Parallelzeichen mit veränderlicher Blocklänge und Kontrolle der Blöcke pro Rahmen Simulation von Lochstreifen- oder Magnetbanddaten. Digitalgesteuerte Wiederholung steht für Erzeugung längerer Simulationen zur Verfügung.

In der Serienarbeitsweise können für das Simulieren von PCM-Daten oder Multiplex-Systemen acht verschiedene 16-Bit-Serienworte in eine Einzelzeile mit Digitalkontrolle der Worte pro Rahmen kommutiert werden. Abtastlampen zeigen kontinuierlich den Programmzustand an. 50-Ω-Ausgangspuffer treiben Verbindungskabel direkt, so dass keine zusätzlichen Pufferschaltungen erforderlich sind.

Rahmen können laufend wiederholt, manuell regeneriert oder durch externes Befehlssignal wiederholt werden. Eine Drucktaste im Bedienfeld gestattet Rückführung in den Rahmen-Start-Zustand. Eine Einrichtung zum Stoppen des Programmes durch ein externes Steuersignal ist vorhanden.

Zur Synchronisierung wird ein 50 ns breiter, positiver 6-V-Impuls mit 100 Ω Quellenwiderstand abgegeben und kann so vorgewählt werden, dass er gleichzeitig mit einem beliebigen Zeichen oder einer Zeichenkombination auftritt. Die Startsynchronisierung des Rahmens erfolgt gleichzeitig mit dem ersten Bit oder Zeichen in jedem Rahmen. Synchronisierung des Rahmenendes erfolgt gleichzeitig mit dem letzten Bit oder Zeichen.

EE 83 772 *für weitere Einzelheiten*

Tragbares Gaussmeter

Preformations Ltd, Cheney Manor, Swindon, Wiltshire

(*Abbildung Seite 483*)

Ein neues tragbares Gaussmeter für lineares statisches Messen magnetischer Feldstärken mit dem grossen Messumfang von 5000 bis zu 105 000 Gauss wurde von der Plessey-Tochter Preformations Ltd entwickelt. Die Firma, deren Erzeugnisse unter der Schutzmarke "Magloy" vertrieben werden, ist einer der grössten britischen Hersteller von Dauermagneten.

Man glaubt, dass es sich hier um das erste Messgerät handelt, das auf einem wenig bekannten magnetostriktiven Effekt beruht und einen aus einer kleinen Indiumantimonid- oder Indiumarsenidplatte bestehenden Tastkopf verwendet. Diese Platte ändert ihren Widerstand in einem senkrecht zu ihrer Oberfläche wirkenden Magnetfeld; die Änderung wird mittels eines transistorisierten Widerstandsmessers überwacht, der einen Gleichstromverstärker speist.

Über 5 kG wird diese Widerstandsänderung eine merklich lineare Funktion der Kraftliniendichte bis zu Feldstärken von 15 kG.

Eine interne Eicheinrichtung macht externe Eichfelder überflüssig. Das Messgerät kann auch zusammen mit einem Oszillografen, Schreiber oder einer Telemetrie-Einrichtung eingesetzt werden.

Man kann Tastköpfe herstellen, die nur 0,15 mm dick sind, so dass sie selbst in sehr kleinen Luftspalten verwendbar sind; die Anzeige ist immer positiv—unabhängig von der Feldpolarität. Ein Vorteil der Tastköpfe dieses Types ist, dass man mit ihnen eingehende Untersuchungen der Linearität und Gleichmässigkeit magnetischer Felder anstellen kann.

Das ursprünglich für Untersuchungen der ferromagnetischen Resonanz bei Mikrowellenfrequenzen entwickelte Gerät kann auch in supraleitenden Magnetspulen sowie in Apparaten für die Protonresonanzforschung, Massenspektroskopie und Plasmaphysik Anwendung finden. Bei Verbindung des Instrumentes mit einem Oszillografen wird das Messen sehr hoher Impulsströme—wie sie z.B. in der Laserpumpenspeisung und in Impuls-Magnetisiergeräten auftreten—und die Darstellung ihrer Wellenform möglich.

EE 83 773 *für weitere Einzelheiten*

Rasterprojektor

Marconi Instruments Ltd, St. Albans, Hertfordshire

(Abbildung Seite 483)

Der Rasterprojektor TM 8098 hat einen Halbreflexspiegel, um normale Beobachtung der Oszillografdarstellung zu gestatten, während Bezugsdaten verschiedenster Art völlig parallaxfrei in dieselbe Bildebene projiziert werden. Typische Anwendungsmöglichkeiten für dieses vielseitige Zubehör sind: Projektion von Rasternormalen für hochgenaue Untersuchungen, Spezialraster für Impuls- und Balkentest, Standardwellenformen und Grenzwertlinien für die Fertigungskontrolle. Die zu projizierenden Daten werden durch fotografisch Reduktion von Stammzeichnungen durch direktes Fotografieren der Bezugs wellenformen oder Skizzieren au "Scribecoat"-Blättern erstellt. De kompakte Zusammenbau wird einfac und schnell an den nunmehr zur Stan dardausrüstung aller Marconi-Oszillo grafen gehörenden Haltering geklemmt Die für die Projektorlampe erforderlich Energie wird über eine Klinke entnom men, mit der alle Instrumente aus de laufenden Fertigung ausgerüstet sind un die die Speisespannung für die normal Rasterbeleuchtung umlegt. Der Zubehö ist auch in einer anderen Ausführun TM 8097 lieferbar, die für Haltering mit 143 mm Durchmesser geeignet ist Die 6,3-V-Spannung für diese Ausfüh rung wird über einen MODAC-Stecke und die in älteren TF 2 000-Modelle vorhandene Buchse für die Speisung de Kathodenfolgermesskopfes entnommen Für andere Oszillografen ist eine ge trennte 6,3-V-Stromversorgung erforder lich.

EE 83 774 *für weitere Einzelheite*

Zusammenfassung der wichtigsten Beiträge

Der pfeifende Geschwindigkeitsmesser und seine Anwendung für die Bestimmung der Grösse turbulenter Wirbel in der Se von D. J. Dunn

Zusammenfassung des Beitrages auf Seite 432-436

Verschiedene Verfahren zum Messen der Änderungen in der Strömungsgeschwindigkeit turbulente Wassers werden besprochen. Eine Kurzbeschreibung des pfeifenden Geschwindigkeitsmessers wir gegeben, und es wird gezeigt, dass auch er sich zum Messen von Strömungsgeschwindigkeitsschwan kungen eignet.

Ein Transistor-Ratemeter mit 2-Mikrosekunden-Auflösungsvermögen von B. Shepherd

Zusammenfassung des Beitrages auf Seite 437-441

Der Beitrag beschreibt ein Ratemeter für regellose Impulse mit sieben Teilbereichen von 10 Imp/Sek und sechs mittelwertbildenden Zeitkonstanten von 0,25 s bis zu 50 s. Jeder Eingangsimpuls erzeug Leinen Rechteckimpuls konstanter Breite, der seinerseits den Speicherkondensator mit einer fest adungsquantität beaufschlagt. Die sich ergebende Spannung wird in einem in eine Millerschleife einbegriffenen Chopperverstärker verstärkt und gibt die gewünschte integrierte Zeit. Der Ausgan wird auf einem Messgerät angezeigt, und ausserdem stehen für Schreiber an Klemmen 1 mA un 100 mV zur Verfügung. Die Messunsicherheit ist 2 Prozent.

Ein selbstkontrollierender kartesischer Wechselspannungsteiler für den Bereich 100 . . . 10 000 Hz von E. R. Wigan

Zusammenfassung des Beitrages auf Seite 442-449

Eine auf C-R-Netzwerken beruhende neue Form eines kartesischen Wechselspannungsteilers wir beschrieben und seine Theorie, Konstruktion und Anwendungsmöglichkeiten in Ausdrücken eine Modellgerätes besprochen.

Mittels eines selbstkontrollierenden Programmes werden die Leistungsparameter des Modells i Absolutwerten geeicht. Nachfolgend wird die Präzision der Eichung vorgeführt, indem das Mode zur Bestimmung der Vektorspannungsverhältnisse in einigen Schaltungen mit kritischen Betriebseigen schaften eingesetzt wird.

Der Frequenzbereich liegt zwischen 100 und 10 000 Hz in 100-Hz-Stufen, und durch einen Kunst griff kann man zwischen diesen Stufen interpolieren. Eine Frequenz-Messbrücke bildet einen Te des kartesischen Netzwerkes und dient zur Eichung des zugehörigen Steueroszillators.

Trotzdem das Modell nicht mit Präzision gebaut wurde, beträgt die Messunsicherheit nur 0,5 Prozent; durch Anwendung eines Spezialprogrammes kann sie auf 0,1 Prozent verbessert werden. Es wurde nicht versucht, Spnnungen in Absolutwerten zu messen, da die Hauptfunktion des Instrumentes das Messen von Verhältnissen zwischen Wechselspannungen ohne Stromentnahme ist. Spannungsmessungen können auf Wunsch mit der Genauigkeit von Zeigermessgeräten ohne Stromentnahme durchgeführt werden.

Analogie eines Grossignaltransistors von G. G. Bloodworth und D. N. Axford

Zusammenfassung des Beitrages auf Seite 450-455

Für die Einleitung des Verhaltens des Transistormodells, das man normalerweise für den Schaltungsentwurf annimmt, wurde eine Analogie erstellt, bei der jedoch die wirksamen Grenzfrequenzen massstäblich verkleinert sind. Das elektrische Verhalten des Basisgebietes wird durch ein lineares, passives Netzwerk dargestellt, und die erforderlichen nichtlinearen Spannungstransformationen an der Übergangszone werden durch die Charakteristiken von Dioden erzeugt. Die Analogie kann zur Erklärung der Bedeutung der verschiedenen Transistorparameter sowie zur Voraussage ihrer Leistung bei Anwendung in Schaltungen herangezogen werden.

Der lineare Ratemeter-Integrator der Harwell-Baureihe 2000 von J. W. Wilkins und D. J. Marsh

Zusammenfassung des Beitrages auf Seite 456-459

Auf Grund ihrer höheren Genauigkeit hat die Digitaltechnik im Verlauf der letzten Jahre die Analogtechnik bei der Geschwindigkeitsbestimmung und Zählung abgelöst. Der Beitrag beschreibt Schaltungsverbesserungen der ursprünglichen Analogsysteme, die ein sehr einfaches Zweizweck-Präzisionsinstrument mit niederspanniger Stromversorgung ergeben.

Ein Impulshöhenmessgerät von C. I. Sach

Zusammenfassung des Beitrages auf Seite 460-464

Ein Gerät, das den Spitzenwert eines Einzelimpulses negativer Polarität misst, wird beschrieben. Es wurde ursprünglich für den Fall eines Impulses mit steiler negativer Vorderflanke in der Grössenordnung von 1 . . . 5 V mit nachfolgendem exponentiellen Abfall auf Null mit einer Zeitkonstante von zwei Sekunden entwickelt. Der Wert des Spitzensignals wird mit weniger als 5 Prozent Unsicherheit auf einem Gleichspannungsmesser angezeigt, vorausgesetzt dass die Ablesung innerhalb 10 Sekunden nach dem Vorgang erfolgt. Durch seine Eigenschaften ist das Gerät für Anzeige des Spitzenwertes von Einzelrechteckimpulsen mit Impulsdauer bis zu drei Millisekunden herunter geeignet.

Ein niederfrequenter pseudoregelloser Rauschgenerator von C. Kramer

Zusammenfassung des Beitrages auf Seite 465-467

Eine beschriebene Schaltung erzeugt Rauschen im unterhörfrequenten Bereich mittels eines Rückkopplungsschieberegisters. Das Rauschen ist nahezu regellos mit binomischer Amplitudenverteilung. Die Kurve ist im Interessengebiet linear. Die statistischen Kenndaten sind hochkonstant und voraussagbar.

Ein umkehrbarer Dekadenzähler mit symmetrischer 1-2-4-2-Kodierung von I. Scollar

Zusammenfassung des Beitrages auf Seite 468-470

Ein binärer Dekadenzähler mit symmetrisch bewertetem Code (1-2-4-2) wird beschrieben, der einige der Schwierigkeiten überkommt, die einem früher vom Autor beschriebenen umkehrbaren Zähler eigen sind. Die Dekade kann ohne Schwierigkeiten aus greifbaren logischen Moduln aufgebaut werden und verarbeitet Zählgeschwindigkeiten bis zu etwa der Hälfte derjenigen der benutzten Binärwerte.

Harmonische Linearisierung des Oszillators der Wien-Brücke von M. A. Kaaz

Zusammenfassung des Beitrages auf Seite 470-471

Zur Erreichung des frequenzkonstanten Betriebes eines Oszillators muss im System eine gewisse Nichtlinearität vorhanden sein; sie muss für ausreichende automatische Amplitudenregelung genügen, sollte aber nur die geringstmögliche harmonische Verzerrung mit sich bringen. Für diese Seite des Wien-Brückenoszillators wird unter Benutzung des Konzeptes der durchlaufenden Funktion eine Analyse gegeben.

Verbessertes fotoelektrisches Messverfahren von Matthews, Faulkner und Green

Zusammenfassung des Beitrages auf Seite 472-474

Bei der Bestimmung fotoelektrischer Schwellenwerte muss man auch Lichtstromstärken bis zu 10^{-15}A herunter messen. Das geschieht am besten, indem man die auffallende Erregerstrahlung zerhackt und ein Wechselstromsystem anwendet, in dem mit Hilfe eines phasenempfindlichen Gleichrichters eine sehr schmale Bandbreite erreicht wird. In diesem Beitrag wird eingehend beschrieben, wie man ein solches Wechselstromsystem aus kommerziell greifbaren Bauelementen aufbauen kann. Die Vorteile dieses Verfahrens über die herkömmlichere Methode mit einem Gleichspannungselektrometer werden besprochen.