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Commentary

'RESEARCH and Development in Electronic Capital Goods' is the subject of a fifty-two page article in the November 1965 issue of National Institute Economic Review published by the National Institute of Economic and Social Research. It is an excellent review of the world position in this sector of industry and provides a great deal of food for thought.

The preparation of a review of this type, if it is to be of any real value, is an exceedingly difficult task; no country has satisfactory national statistics and thus they have to be culled from various sources. As a consequence the basis for their calculation is not always the same and they must, perforce, be used with considerable discrimination. Due weight has of course been given to this factor.

Many figures are given in the review and the countries taken into account are the U.S.A., the U.K., Western Germany, France and Japan. So far as world markets are concerned these countries may be said effectively to provide the world's total output. In the production of electronic capital goods the U.K. takes second place to the U.S.A., followed by France, Western Germany and Japan. But, while most engineers are aware of the lead held by the U.S.A., there are probably few who realize just how great is the disparity or just how near the U.S.A. comes to the domination of world markets. Taking the total value of production of the five countries as 100 per cent, then the percentage contribution of the countries is: U.S.A. 81·9, U.K. 6, France 5·1, Western Germany 4·2 and Japan 2·8. In terms of exports the percentages are: U.S.A. 60·5, U.K. 14·5, Western Germany 12·7, France 7·3 and Japan 5. The 60 per cent share of world exports in electronic capital goods which the United States enjoyed in 1963 and 1964 may be compared with one of about 20 per cent in world exports of all manufactured goods and 30 per cent in world exports of machinery. It must also be borne in mind that these estimates of the United States share, both in world production and exports, substantially understate the contribution of American companies to this industry. Subsidiaries of American companies operating in the other principal manufacturing countries account for a significant part of those countries production and exports. In Western Europe the share of American owned companies in the production of electronic capital goods is probably between 15 and 20 per cent: in France it is over a quarter.

One of the fastest growing divisions of the industry is that concerned with electronic data processing equipment and here the dominance of the U.S.A. is even more pronounced with a 74·5 per cent share of world exports. That though, is only half the story for in most other countries American subsidiaries account for almost the whole of production; and all this due largely to the efforts of one firm! The largest American company, International Business Machines (IBM) not only accounts for over two thirds of American installations by value and nearly two-thirds by number but also for about the same proportion outside the United States. IBM is by far the largest manufacturer in France and Japan. Other American computer manufacturers also have large interests in European production. Indeed outside of Britain the only significant non-American producers who still design and develop their own equipment are those making small scientific computers or specialized installations.

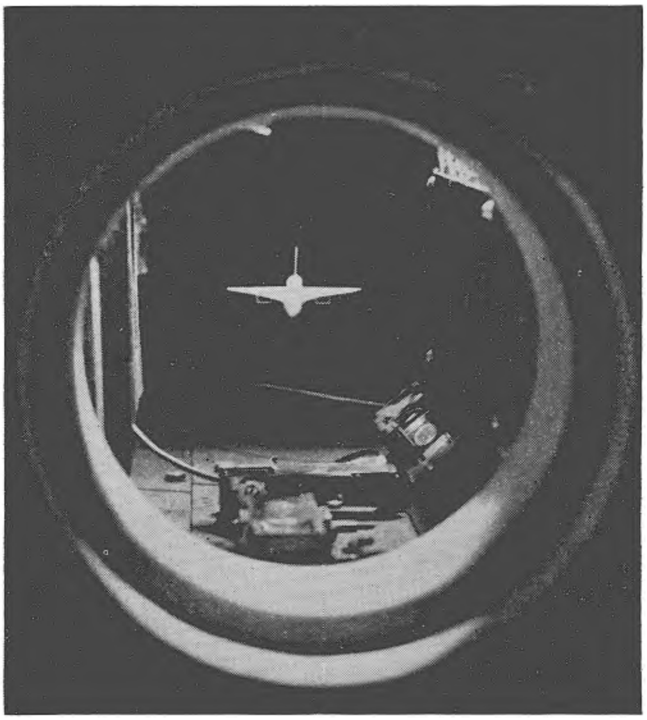
The reasons for this large American lead are by no means simple or straightforward. There are a number of fairly obvious contributory factors but none, taken singly provides a complete answer. With a population some three and a half times that of any single European country America has a large home market and, in general, the American people are more willing to accept innovations than are Europeans; but it must be remembered that it is a world market and America is not dominating it by means of low prices subsidized by the home market. The American Government has sponsored a great deal of research and development in connexion with its space and military programmes, although some of the companies which have been most successful in world markets have received very little government aid; that is also true for some firms in Britain. Usually advanced components are available to American equipment manufacturers earlier than they are to European manufacturers; for example, most types of semiconductor devices have been available in America about two years earlier than in Britain, this is due, to some extent, to the military and space research programmes. Some American companies have vast resources in men and money; for example there are about half-a-dozen American companies any one of which is now spending nearly as much on electronic research and development as the German Federal Republic or France.

This latter point is perhaps the crux of the matter and it is part of a very vicious circle. In a rapidly progressing scientifically based industry such as that of electronic capital goods, competition mainly takes the form of technical innovation and technical service to customers. Each firm which wishes to stay in the business must be capable, if not of making a major innovation itself, at least of imitating those made by its more advanced competitors within a short time. To do this it must have a certain research and development capacity and if it falls below this 'defensive threshold' of R & D it will fail. If the 'threshold' set by the leading American firms for particular products is beyond the resources of individual European firms then—barring mergers with American firms—they can only survive by rationalization of the industry either on a national or European basis.

It is clear that if this, or any European, country wishes to stay in this particular race then a great deal of clear thinking, followed by swift and decisive action, needs to be done. Equally clear is the fact that the Government, and in particular the Ministry of Technology, has a very vital role to play.

Magnetic Suspension for Wind Tunnels

By A. Wilson* and B. F. Luff*



A magnetic suspension system for aerodynamic wind tunnel models is described. The system is designed to hold axi-symmetrical models in a Mach 8 airstream. The models, which may be flown at small degrees of incidence, are controlled in five degrees of freedom, and aerodynamic forces may be measured under conditions more nearly approaching those of free flight.

(Voir page 132 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 139)

WIND tunnels are used to evaluate the performance of aerodynamic shapes. The flow of air over the model may be made visible by various means, the most powerful technique being that of schlieren photography. Aerodynamic forces acting on the model can be measured by using strain gauges in the mechanical support (sting) to which the rear of the model is attached. Unfortunately the presence of this sting modifies the flow of air, so that the visualisation is incorrect and force measurements are wrong.

If the model is made of a ferromagnetic material, then a suitable magnetic field should be able to support it. The French research organization O.N.E.R.A. has built and operated such systems[1]. Work on magnetic suspension for wind tunnel models is also going on in the U.S.A.[2,3] where at least one system using a simple spherical model is in operational use.

The system built for the R.A.E. is designed for operation in the 18cm hypersonic wind tunnel, which operates in the Mach 8 region. This tunnel, which is of the 'blow down' type, has an open jet working section surrounded by a plenum chamber, where the electromagnets are situated.

The general arrangement of the system is shown in Fig. 1. Gravitational force is balanced by the two lift electromagnets situated above the model, while the drag force due to air flow is opposed by the drag solenoid which is wound around the tunnel and exerts an axial force on the model. The lateral electromagnets prevent any sideways displacement or yawing motion of the model.

The position of the model is measured by light beams which are directed across the tunnel working section and fall on to photo-electric cells. There are five light beams to control the power in the appropriate electromagnets. When the model is in the correct position, half the light beam is obscured by the model (Fig. 2); if the model moves from the correct position, the amount of light received by the photocell will change, causing a change of magnetic force on the model, tending to restore it to the original position.

Stability

Analysis of the stability of the system shows that a permanent magnet or a piece of soft iron, cannot be supported by an electromagnet using simple proportional control.

In Fig. 3 is shown the open control loop, where x_i is the displacement of the model detected by the photocell, the output of which is amplified (gain G) and used to drive the electromagnet (time-constant T_L seconds). The force of the electromagnet (F dynes) moves the model (mass m grams) through x_o cm.

Then:

$$x_o/x_i = -\frac{G}{m\omega^2(1 + j\omega T_L)} \quad \ldots\ldots\ldots\ldots\ldots\ldots (1)$$

This asymptotic transfer function is shown in Fig. 4 curve A, where G/m has been chosen as unity, and $T_L = 1/\omega_L = 1/66$sec. The slope of the transfer function is 12dB/octave increasing to 18dB/octave at frequencies greater than $\omega = 1/T_L$. The closed loop is unstable because at no point is the transfer characteristic slope less than 12dB/octave, irrespective of gain.

The above photograph shows a Concorde type model in suspension

* *Royal Aircraft Establishment.*

The phase lag due to the coil can be compensated by adding a simple phase advance network (Figs. 5 and 6) whose transfer function is:

$$E_{out}/E_{in} = T_2/T_1 \frac{(1 + j\omega T_1)}{(1 + j\omega T_2)} \quad \ldots\ldots\ldots\ldots\ldots\ldots (2)$$

where $T_1 = R_1C$ and $T_2 = \frac{R_1R_2}{R_1 + R_2}\ C.$

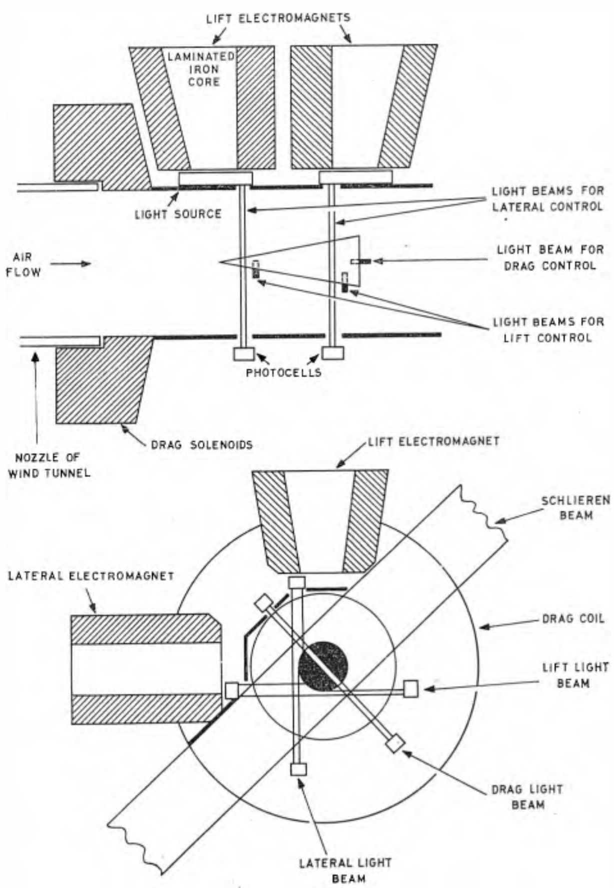
If $T_1 = T_L$ and $T_2 = 1/25 \times T_L$ and the loop gain is increased to compensate for the low frequency attenuation of the network, the slope of the transfer function becomes a constant 12dB/octave at all frequencies below $\omega = 25/T_L$. The closed loop is still unstable however, so another phase advance network is added, having similar characteristics to the first.

This produces a 6dB/octave slope on that part of the transfer function between $\omega = 66$ and $\omega = 160$ approximately. Thus if the loop gain is increased by 78dB or 8 000 times, the closed loop is stable (Fig. 4, curve *B*). The transfer function is now:

$$x_o/x_i\,(j\omega) = -\frac{G}{m\omega^2}\ T_2/T_1 \frac{(1 + j\omega T_1)}{(1 + j\omega T_L)(1 + j\omega T_2)} (T_2'/T_1') \cdot \frac{(1 + j\omega T_1')}{(1 + j\omega T_2')} \quad \ldots\ldots (3)$$

Unfortunately the force available is not sufficient to support a typical model, or to counteract the drag force in the case of the drag coil. There should only be about 0·1cm movement of the model for a force of 300 grams acting upon it, i.e. $G = 3 \times 10^6$ dynes/cm. However, as m is about 75 grams $G = 75 \times 8\,000 = 6 \times 10^5$ dynes/cm.

Fig. 1. General arrangement of the system



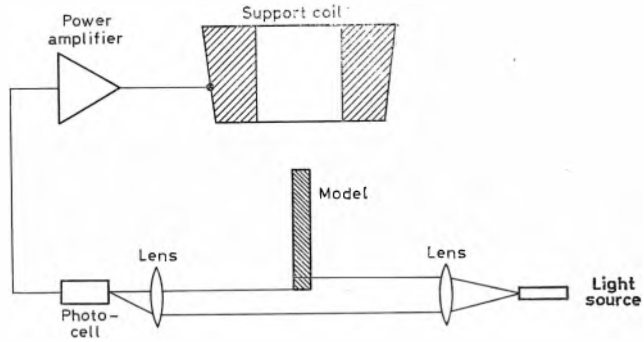


Fig. 2. Light and photocell arrangement

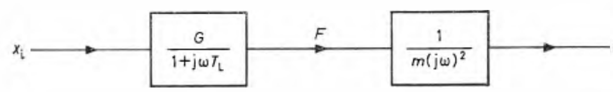


Fig. 3. Basic arrangement of control network

This is not sufficient force so more loop gain is required.

The gain at low frequencies can be increased by using an integrating circuit which only gives an extra 6dB/octave slope down to $\omega = 1/T_L$. This can be achieved by connecting an integrating amplifier in parallel with the second phase advance network, and arranging that their relative gains are equal at about $\omega = 1/T_L$.

The individual transfer functions are added, thus making the overall transfer function:

$$x_o/x_i\,(j\omega) = -\frac{G}{m\omega^2}\ T_2/T_1 \frac{(1 + j\omega T_1)}{(1 + j\omega T_L)(1 + j\omega T_2)} \left[(T_2'/T_1') \cdot \frac{(1 + j\omega T_1')}{(1 + j\omega T_2')} + \frac{\alpha}{1 + j\omega T_R} \right] \quad \ldots\ldots\ldots\ldots (4)$$

where T_R is the time-constant of the integrator, and α is the ratio of the gain of the integrator to the gain of the phase advance network.

The increase in gain obtained by this modification of the transfer function is decided by the zero frequency gain of the integrating amplifier, which in this case is adjusted to be 20 times. Thus the overall gain becomes 12×10^6, which is more than sufficient. The final transfer function curve is shown at *C* in Fig. 4.

The complete block diagram of the control loop is shown in Fig. 7 and the theoretical response of the correcting networks in Fig. 8.

A schematic diagram of one of the five control systems is illustrated in Fig. 9.

At first a horse-shoe electromagnet was used with a coil wound on each limb. However, the mutual coupling between the two coils made stable control difficult, so the connecting yoke was removed, leaving two separate iron-cored solenoids. This raised the required support currents by 10 per cent, but improved the suspension stability. When the length of the model is similar to the distance between itself and the supporting coils, some undesirable cross-coupling between the front and the rear of the model occurs. The front solenoid not only attracts the front of the model but it also rejects the rear. Similarly the rear solenoid repels the front of the model (Fig. 10). The effect may be reduced by cross-coupling the detecting photocells.

Model Characteristics

As the model is part of the magnetic circuit, its magnetic properties play an important part in the system performance. A permanent magnet may be inserted in a non-magnetic aerodynamic shape, or the whole model may be made of soft iron.

The advantage of the soft iron model is that the induced saturation flux density is higher than the remanent flux density of permanent magnet material, and therefore greater forces on the model may be realized. Also, there is more room in the model for telemetry equipment which will be necessary if physical parameters inside the model are to be measured.

The disadvantage of soft iron is that the cross-coupling effects are accentuated due to the pole strength depending on induced magnetism. Cross-coupling therefore makes calibration of a model for force measurements a lengthy business.

The permanent magnet in a model does not enable so much force to be obtained, however cross-coupling is reduced considerably, although not completely. Materials having a coercive force of 750 oersted show variations of about 1 per cent in support currents. A more stable material having a coercive force of 4 800 oersted is expected to reduce this variation.

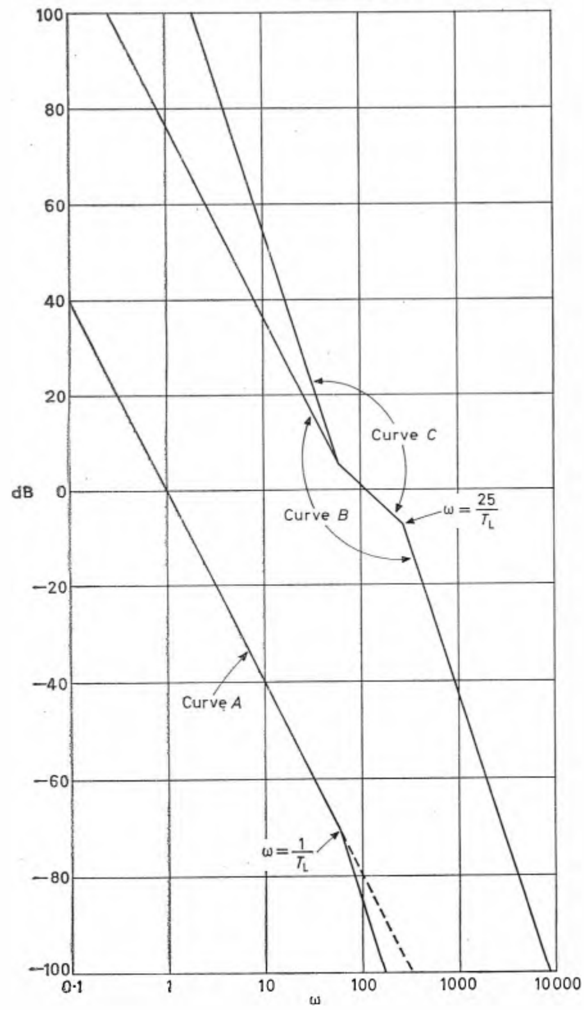
Optical System

The optical system must give an accurate and stable measurement of the position of the model in the noisy and unsteady environment of the wind tunnel. A simple collimated beam of light from a rugged straight filament incandescent lamp, reflected via a 90° prism to save space, gives a simple and effective source.

The photocell must be unaffected by magnetic fields, have a reasonably fast response time, be insensitive to temperature, and preferably have a linear input/output characteristic. Silicon solar cells have been found satisfactory, provided the load is matched to the internal impedance of the cell. A cylindrical Perspex lens in front of the cell reduces the effect of beam movement.

Absolute rigidity in the optical components is essential, but provision must be made to move them to accommodate different model shapes and sizes.

Fig. 4. Theoretical loop gain of control loop



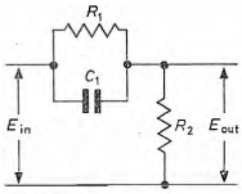


Fig. 5. Phase advance network

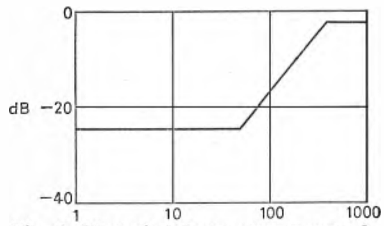


Fig. 6. Theoretical frequency response of phase advance network

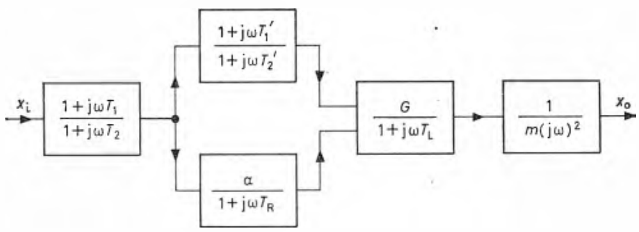


Fig. 7. Arrangement of complete control network

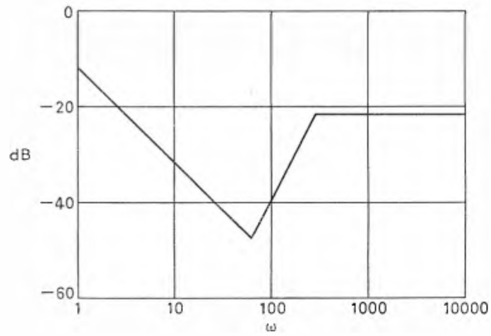


Fig. 8. Theoretical frequency response of complete correcting network

Electromagnets

The design of the electromagnets is a compromise between the magnetic force required and their size. Scale model tests showed that the suspension coils should provide 25 000 ampere turns per leg. The resistance of the coil is fixed by the method of providing the power. At the time of early development, the maximum peak inverse voltage rating of the thyristors used to control the power was 400V, giving a safe maximum output of about 130V. In order to provide the necessary magnetic field in the space available, a coil having a resistance of 2Ω, with 650 turns, and capable of passing 40A was suitable. As the power supplies are capable of delivering a maximum of 240A, the coil windings are arranged in two halves connected in parallel giving a resistance of 0·5Ω, the time-constant being 50msec.

The temperature rise due to the large amount of heat (3kW) dissipated in the coil is reduced by winding it with copper tubing through which cooling water flows. The

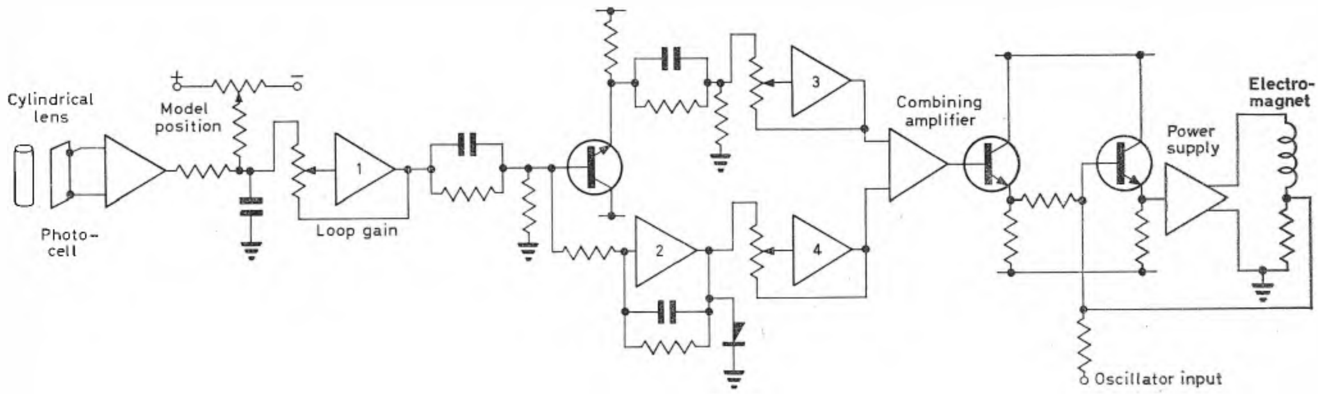


Fig. 9. Circuit of control system

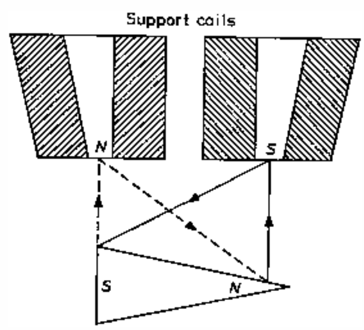


Fig. 10. Magnetic cross coupling

tube has an outside diameter of 2·40mm and an inside diameter of 0·94mm. The water has to be carefully filtered to prevent obstruction of the bores. Each layer of the coil is brought out to a manifold which connects the layers in series electrically but in parallel for water flow; this arrangement is necessary to achieve adequate water flow from reasonable water pressures, normally about 4·3kg/cm^2 (60 lb/in^2).

The iron cores were of Sankey's Lohys 25, but because the hysteresis of this material made force calibration difficult, it is being changed to Permalloy C, even though less magnetic force will be available due to the lower saturation density; it is possible that air-cored coils will have to be used to give accurate force measurements.

The drag coil has to work at higher peak currents than the other coils. The drag force can be larger than the gravitational force, and iron cannot be used to raise the permeability. This coil is wound of copper tubing having dimensions of 3·20mm outer diameter and 1·50mm inner diameter. The coil can produce forces of 0·9kg (2 lb) on soft iron models working at a distance of 15cm from the coil centre with 65 000 ampere turns.

The lateral coils do not have to provide such large forces as the other coils, but must be capable of reversing their flux polarity, as there is no steady side force to oppose. A bias winding is therefore used, the control winding sharing the same iron circuit as the bias and working in flux opposition to it. The bias current must be fed through a resistance high in value compared to the coil resistance to prevent a 'shorted turn' effect. This would cause a phase lag in the control flux and thus a lag in the control feedback path, which would adversely affect the lateral stability. The coils are wound with copper tubing having the same dimensions as that used for the suspension coils, and provide 10 000 ampere turns with reversible polarity.

All the coils have thermocouples built into the windings, so that internal temperatures may be continuously measured. Water pressure switches turn off the power supplies to the coils in the event of the pressure falling below the safe value for efficient cooling. Initially the water was taken from the public supply mains, but variations in pressure, purity and conductivity have led to the use of a closed circuit system using distilled water, which is pumped through a heat exchanger.

Power Supplies

The power supplies have to provide up to 10kW for the drag and suspension circuits.

The power must be controllable linearly from approximately 5 to 100V, and the frequency response extend to 100c/s. Three-phase full-wave rectification is used in all supplies, which gives a ripple frequency of 300c/s, high enough to prevent oscillation of the model. Thyristors are used to control the power, as they have excellent efficiency and do not require any warm-up time. Control is obtained by varying the phase of the firing pulse applied to the gate electrode of the thyristor with respect to the anode voltage, thus varying the conduction time over each half cycle. The phase is variable between 5° and 130°. The three gate drive units driving the six thyristors have identical sensitivities to prevent the production of ripple frequencies lower than 300c/s.

When a thyristor conducts through an inductive load a large voltage pulse is produced, the amplitude and ringing frequency of which depends on the load inductance, stray capacitance and transformer leakage inductance. This pulse is radiated by the load coil, supply leads and transformer, and can cause premature firing of other thyristors. Filters and careful attention to wiring layout are necessary to eliminate effects from these pulses.

A thyristor will stay conducting when its anode is negative with respect to its cathode, if a small 'holding' current is present. Inductive energy from the load coil can provide this current, which if allowed to, will hold the thyristor 'on' and reduce efficiency. Two solutions are possible—(a) make the load look resistive by resistance-capacitance compensation, or (b) provide an alternative path for the inductive current by means of a diode. The latter solution is adopted.

To improve the linearity of the supply, current negative feedback is used. This is derived from a small value resistor in series with the load coil, the resulting voltage across this resistor being fed back to the output stages of the control circuit. This feedback also raises the corner frequency of the coil, and improves the impulse response of the system. The output circuits are not earthed directly, but via a 100Ω resistor, which reduces loop earth currents,

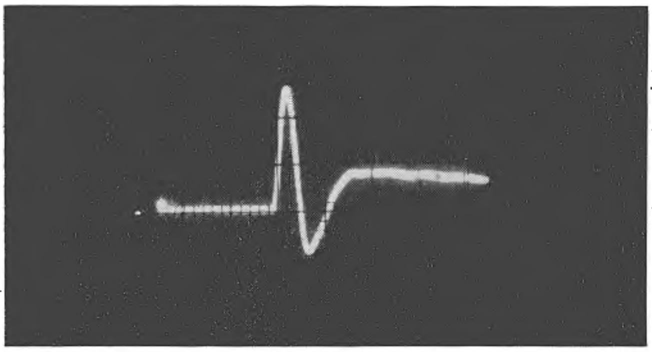


Fig. 11. Model movement after step disturbance

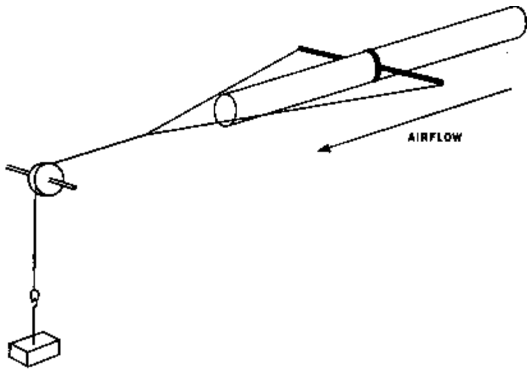


Fig. 12. Method of applying drag forces

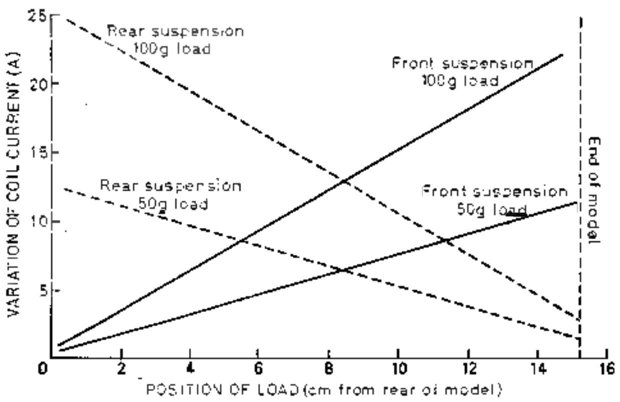


Fig. 13. Variation of suspension currents with pitching moments

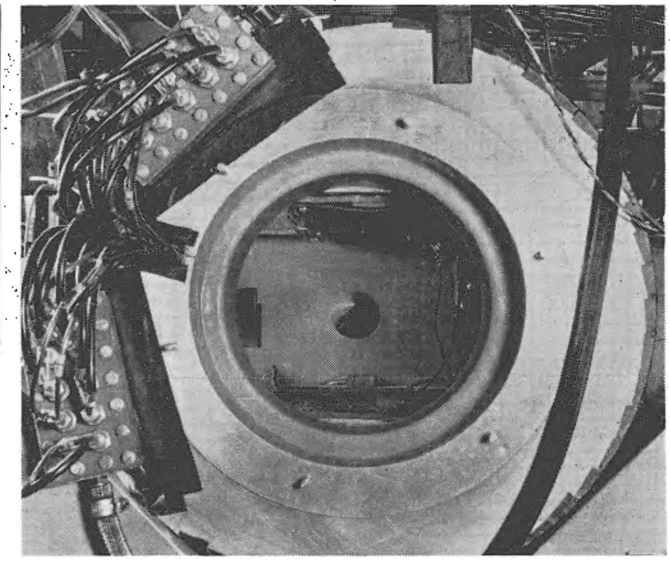
but ties the circuit to earth enabling negative feedback to operate.

The bias supplies also utilize three-phase full-wave rectification, six silicon rectifiers being used in each supply. The current can be varied by means of motor-driven variable transformers.

The maximum capacities of the power supplies are drag and suspension 120V at 240A, lateral control 120V at 60A and lateral bias 100V at 60A.

Testing and Calibrating

Although the stability margin is large enough to accommodate models with different weights and magnetic properties, the best performance for any given model will be obtained when the loop gain is optimized. A convenient method for finding the right gain setting is to disturb the model and observe the time and amplitude characteristics of its recovery. The recovery can be observed on an oscilloscope connected to the appropriate position sensing



Conical model in suspension

photocell; movements as small as 0·02mm can be observed. The model is disturbed by dropping weights into a pan attached to the model by a harness. The gain of the control loop is then adjusted to produce the least displacement and desired shape of overshoot. Fig. 11 shows a typical oscillogram of model movement following a step disturbance.

If the coil currents are to be used to obtain aerodynamic force measurements, it is necessary to calibrate a given model before carrying out the tunnel test. This can be done by applying known forces and moments to the model while it is being magnetically supported, and recording the various currents. The currents are measured by a digital voltmeter which is connected to a high grade shunt in series with the appropriate coil.

As the force on the model changes with position, it is essential that the model be maintained in the position occupied during calibration. If movements greater than 0·05mm occur, then the force measuring accuracy cannot be better than ± 1 per cent.

Drag calibration forces are applied to the model via a harness attached to the model. The harness passes over a pulley and terminates in a receptacle which receives the calibration weights (Fig. 12). Lift and pitching moments are obtained by hanging a range of weights on various positions along the length of the model. A typical pitching moment calibration is shown in Fig. 13 for a permanent magnet model. It can be seen that the pitching moment produces a linear change in the suspension currents.

Although more work is necessary to complete the magnetic suspension system for aerodynamic models, it is already evident that it is a powerful technique, enabling previously difficult or impossible measurements to be made in conditions more closely approaching those of free flight.

Acknowledgment

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The Design of Differential D.C. Amplifiers with High Common Mode Rejection

By G. Meyer-Brötz* and A. Kley*

The amplification of small d.c. voltages, as a signal in a data-processing system for instance, requires differential amplifiers that amplify the signal but reject interference voltages on earth lines. Starting off with two basic amplifiers (a three-terminal amplifier and four-terminal amplifier) this article goes on to discuss the common-mode rejection attained in four differential amplifiers.

(Voir page 132 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 139)

THE problem of directly fitting analogue or digital computer systems into a control process is becoming increasingly important in automation. The function of the process computer is to continuously collect and handle the unknown quantities of the process (pressures, temperatures, etc.). The d.c. voltages from the signal sources

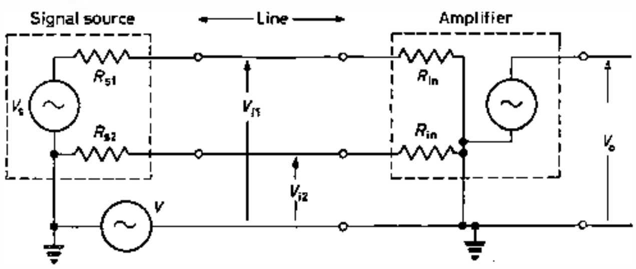

Fig. 1. Signal amplifier designed as a differential amplifier with high common-mode rejection

(resistance bridges, thermocouples, etc.) range from 10mV to 1V and are usually too low to be used directly as inputs. For adaptation to the usual level of 10V they must be amplified by a factor of 10 to 1 000. The d.c. voltage amplifiers have to meet the most exacting requirements as to accuracy and stability of amplification, linearity, zero-point error and drift, interference, and input resistance[1].

Since the load impedance of the amplifier is always grounded, one of its output terminals must be connected to chassis (Fig. 1). Grounding of the signal source is required in many cases, and is a fact even where not required, as a result of low-resistance leakages. An interference voltage V (Fig. 1), especially such as a hum at the line frequency, appears between the two ground points, which may be spaced a considerable distance apart; the amplitude of this voltage may be as high as several volts. The important problem of rejecting this interference voltage will be discussed in this article.

Assume that the signal amplifier shown in Fig. 1 is a differential amplifier that responds to the difference of the input voltages V_{i1} and V_{i2} so that, ideally, its output voltage is:

$$V_o = A\,(V_{i1} - V_{i2}) \quad \ldots\ldots\ldots\ldots\ (1)$$

With $V_{i1} = V_s + V$ and $V_{i2} = V$, the desired function of the amplifier for infinitely high input resistances R_{in} is to yield an output voltage:

$$V_o = AV_s \quad \ldots\ldots\ldots\ldots\ldots\ (2)$$

Thus, the signal voltage V_s is amplified and the interference voltage V is completely rejected.

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The output voltage of a real amplifier deviates from that given by equation (1):

$$V_o = A\,.\,[V_{i1} - (1 + \delta)\,V_{i2}] \quad \ldots\ldots\ldots\ (3)$$

i.e. it does not respond exactly to the difference of the two input signals, but an unbalance $\delta \ll 1$ is present.

Furthermore, the finite input resistance R_{in} of the

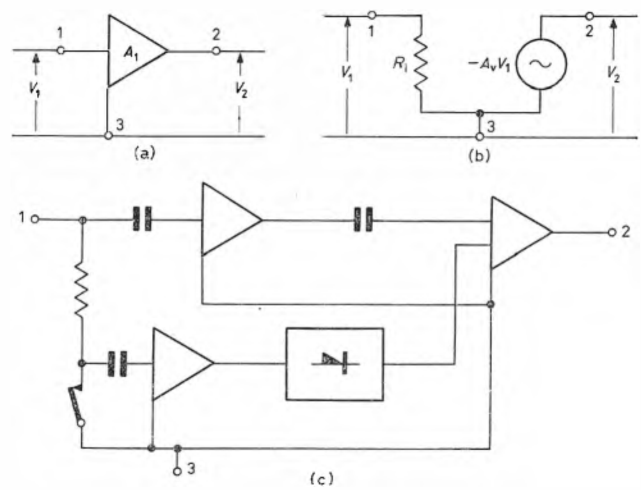


Fig. 2. Three-terminal amplifier

(a) symbolic representation
(b) equivalent circuit
(c) typical design with a chopper-stabilized amplifier

amplifier causes the interference voltage V to be unevenly distributed among the two inputs of the amplifier through the unequal resistances R_{s1} and R_{s2} of the signal source and leads, so that here, unlike equation (2), the output voltage is:

$$V_o = A\,(V_s + (1/D)\,V) \quad \ldots\ldots\ldots\ (4)$$

where the common-mode rejection factor D for R_{s1}, $R_{s2} \ll R_{in}$ is given by:

$$1/D = \frac{R_{s2} - R_{s1}}{R_{in}} + \delta\,; \quad \ldots\ldots\ldots\ (5)$$

which increases with increasing input resistance R_{in} and with decreasing amplifier unbalance[2].

The practical requirements may be illustrated by means of a numerical example. If a signal voltage of $V_s = 10$mV is to be amplified with an accuracy of 0·1 per cent, the rejection D of the interference voltage $V = 10$V must exceed 10^6 (120dB) according to equation (4). If the unbalance of the source resistance $R_{s2} - R_{s1} = 100\Omega$, the input impedance R_{in} of the amplifier must, according to equation (5), exceed 100MΩ and the amplifier unbalance δ must be lower than 10^{-6}.

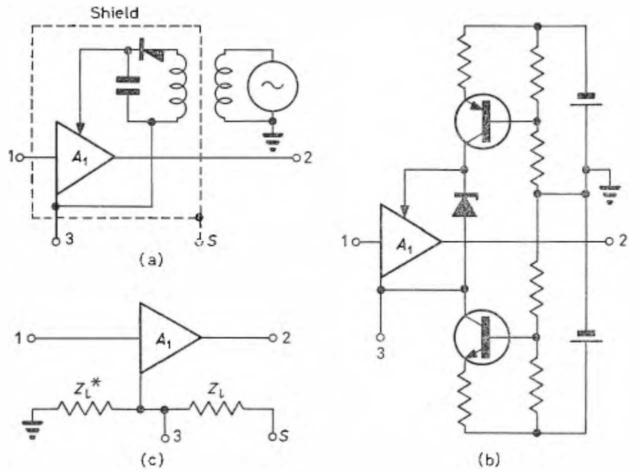


Fig. 3. Power supply of the three-terminal amplifier

(a) with floating power supply unit
(b) with Zener diode biased by current sources
(c) symbolic representation of the leakage impedances to the shield and chassis

The required stability of amplification and the high input resistance of the d.c. amplifiers employed can be realized only by means of strong negative feedback. The signal amplifier shown in Fig. 1 will then consist of a network of high-gain d.c. amplifiers, capable of a large amount of negative feedback, as are known in principle from analogue computer technology[3,4], where the feedback resistances alone determine the gain of the negative feedback amplifier.

The following is a discussion of some of these circuits and of the attainable common mode rejection—a vital problem in the technology of the amplifiers used.

The Three-Terminal Amplifier

Fig. 2 shows the circuit symbol and the equivalent circuit of the three-terminal amplifier, whose input and output have one terminal in common. Its input resistance is R_i, the voltage gain A_v. Fig. 2(c) illustrates its realization in a practical circuit, as used in the operational amplifiers of an analogue computer[3,4]. Transistorized amplifiers may provide input resistances of 100kΩ and a voltage gain $A_v > 10^6$, i.e. transfer resistances of $R_i A_v > 10^{11}\Omega$. The problem of realizing such an amplifier, especially that of stabilizing it against self-excitation under strong negative feedback, is not dealt with in this article.

As will be seen later in this discussion, all terminals of the three-terminal amplifier must be completely available for the design of the differential amplifier. This seems impossible at first sight because the power supply, introduced at terminal 3 of the amplifier, is usually grounded. This reference to the ground potential can be avoided by using a floating power supply (Fig. 3(a)). The amplifier and the secondary side of the power supply are surrounded by a shield, so that in principle all three terminals of the amplifier are isolated from ground. Actually, however, there is an essentially capacitive impedance Z_l between terminal 3 and the shield as well as a considerably higher feed-through impedance Z_l^* with respect to chassis. The amplifier thus obtains a fourth terminal S by way of the static shield, which in some circuits can be connected to a point having no potential difference with respect to terminal 3 of the amplifier, so that the impedance Z_l is of no consequence.

The coupling of terminal 3 of the three-terminal amplifier to the chassis and to the shield, unavoidable even with a floating power supply, is represented symbolically in Fig. 3(c).

An alternative floating supply is shown in Fig. 3(b). The supply voltage of the amplifier is furnished by a Zener diode biased by two current sources, in this example by transistors in grounded-base connexion. Coupling to ground takes place only through the high output resistance of the supply transistors with $|Z_l^*| > 10\text{M}\Omega$. There is neither a shield nor the corresponding terminal S.

The Four-Terminal Amplifier

The use of a floating power supply, which is necessary for the three-terminal amplifier, complicates the design to a certain extent.

Attention will now be turned to a second basic amplifier, a four-terminal amplifier, whose input and output circuits are completely isolated in the idealized case (Fig. 4). If this isolation was complete, the output voltage would be:

$$V_2 = -A_v V_1$$

depending only on the input voltage between the terminals 1 and 2. Actually, there is a feed-through factor α for a common-mode voltage between the terminals 2 and 3 (Fig. 4(a)), so that the output voltage is:

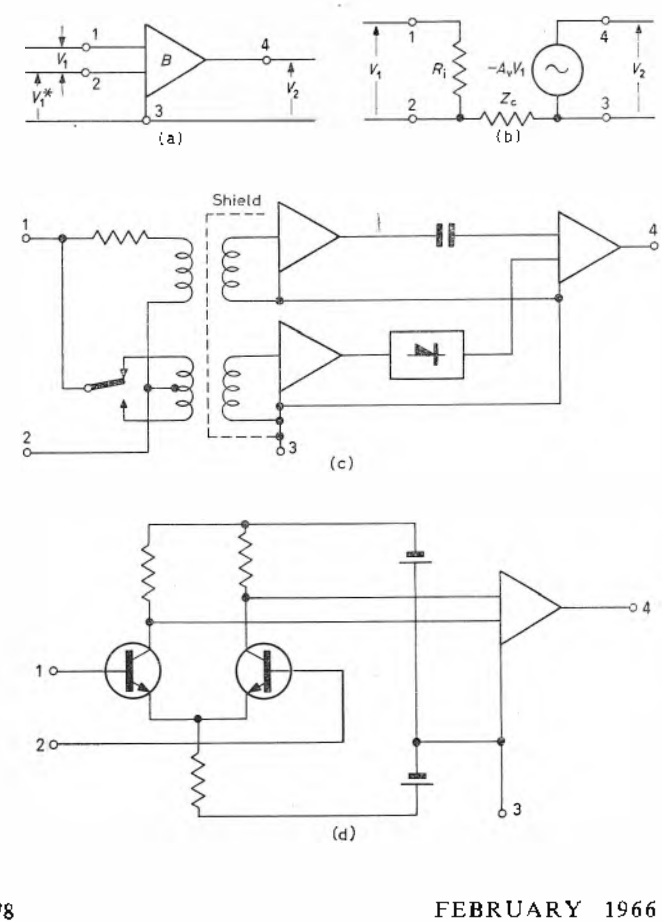
$$V_2 = -A_v (V_1 - \alpha V_1^*),$$

α being very small with respect to unity. In addition, a coupling impedance Z_c exists between the input and output terminals (Fig. 4(b)).

The input and output can be isolated by means of a

Fig. 4. Four-terminal amplifier

(a) symbolic representation
(b) equivalent circuit
(c) typical design with transformer-coupling
(d) typical design using differential input stage



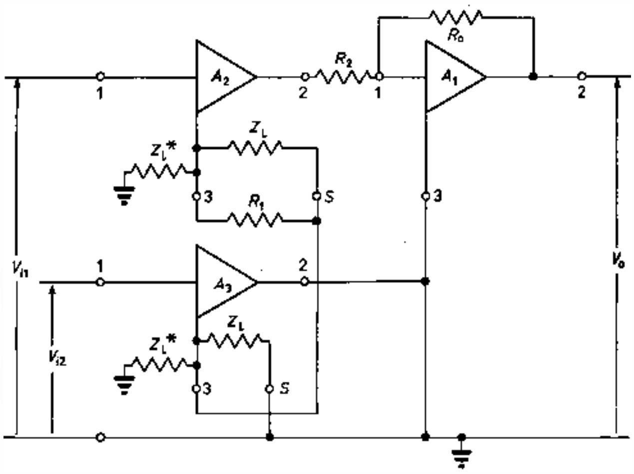


Fig. 5. D.C. amplifier, designed from three three-terminal amplifiers

transformer coupling (Fig. 4(c)), provided that the bandwidth of the upper transformer is extremely large. Although it is not difficult to make the factor α very low and the coupling impedance Z_c very high by using carefully shielded transformers, this circuit is unsuitable for wideband amplifiers.

An alternative wideband design of the four-terminal amplifier can be achieved by using a differential input stage (Fig. 4(d)), in which α can be minimized at will through suitable tricks in circuit layout[5,6].

By isolating the input and output circuits of the four-terminal amplifier a way of grounding the output terminal 3 and thus the power supply is provided in any case. The capacitances to ground in the amplifier circuit are thus no longer of any significance and, in particular, have no effect on the bandwidth.

Differential Amplifier Design

After having introduced the basic amplifiers (three-terminal and four-terminal amplifiers) and having discussed their chief characteristics, the design of differential amplifiers as shown in Fig. 1 will now be dealt with using the basic amplifiers mentioned. In this article only the effect of the various circuit parameters on the common-mode rejection factor D will be considered, which according to equation (5) is in turn determined by the unbalance δ and the input resistance R_{in} of the amplifier.

Fig. 5 shows a differential amplifier circuit composed of three three-terminal amplifiers. The two front-end amplifiers require a floating power supply, for which the coupling to terminal 3, as in Fig. 3(c), is indicated. The shield S of the amplifier A_2 is connected to terminal 3 of the amplifier A_3, which is at the same potential as terminal 3 of the amplifier A_2 with respect to the common-mode voltage V. The leakage impedance Z_l therefore does not affect the unbalance:

$$\delta = 1/A_v + (R_1/Z_l^*)$$

which is only caused by the considerably higher feed-through impedance Z_l^* and the finite voltage gain A_v. With the input impedance

$$R_{in} \simeq R_i (A_v/A)$$

the common-mode rejection factor is given by:

$$1/D = \frac{R_{s2} - R_{s1}}{R_i} (A/A_v)+(1/A_v)+(R_1/Z_l^*) \quad \ldots\ldots\ldots (6)$$

according to equation (5).

The gain:

$$A = (R_0/R_1)(1 + (1/D)) \frac{A_v}{A + A_v} \simeq (R_0/R_1)$$

is determined by the passive negative feedback resistances with a high degree of accuracy. The influence of the common-mode rejection on the error in gain can be completely neglected.

Shielding the entire amplifier A_2 is important because the capacitances to ground, which determine the feed-through impedance Z_l^*, must remain low. Due to this shield there appear, however, small leakage impedances Z between terminals 3 and S of amplifier A_2, which are in parallel with the feedback resistance R_1 and influence the frequency response of the amplification.

A further design alternative for the circuit of a differential amplifier using three-terminal amplifiers is shown in Fig. 6. The symmetry of the amplifier is here established by means of a bridge-like arrangement, composed of the resistances R_0, R_1, R_1^*, R_0^*, in which R_0^*/R_1^* should equal R_0/R_1. The shields S of the floating amplifiers A_2 and A_3 are connected to the output of the amplifier A_4; the magnitude of the leakage impedances Z_l is thereby effectively increased by a factor A. If the error in bridge balance defined by:

$$R_0^*/R_1^* = R_0/R_1 (1 + \epsilon)$$

is taken into account, the unbalance of the amplifier is:

$$\delta = (1/A_v)+(R_1/Z_l^*)+(1/A)\left[(R_1/Z_l) + \epsilon\right]$$

as $A_v \gg A$ the additional second term decreases with increasing gain:

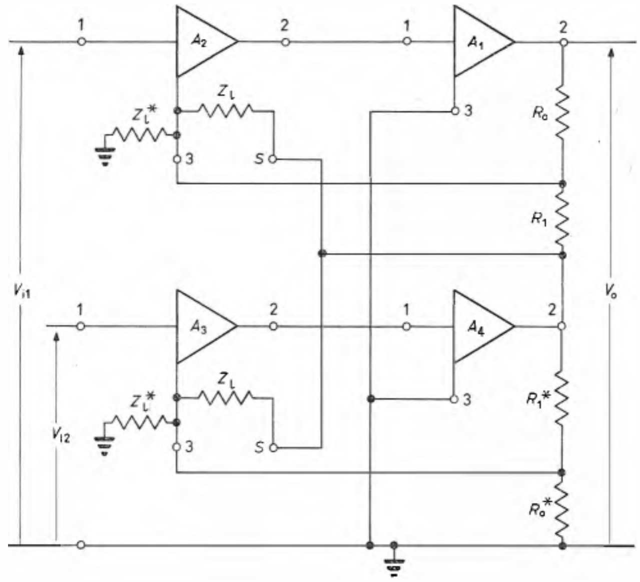
$$A \simeq \frac{R_0 + R_1}{R_1}$$

The influence of Z_l and ϵ thus shows the same relative decrease as the signal voltage V_s. In other words, the interference at the output of the amplifier introduced by the bridge error ϵ and the leakage impedance Z_l remains constant irrespective of the gain.

With the input impedance:

$$R_{in} \simeq R_i (A_v/A)$$

Fig. 6. D.C. amplifier, designed from four three-terminal amplifiers



he common-mode rejection is:

$$1/D = \frac{R_{s2} - R_{s1}}{R_i}(A/A_v)+(1/A_v)+(R_1/Z_1^*)+(1/A)\,[(R_1/Z_1) + \epsilon] \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (7)$$

ıccording to equation (5).

The floating amplifiers must again be shielded, so that he feed-through impedance Z_1^* directly coupled to ground ɔecomes high.

The combined use of three-terminal and four-terminal ımplifiers in a differential amplifier is shown in Fig. 7. With:

$$R_o^* = R_o(1 + \epsilon)$$

ınd

$$R^* = R(1 + \rho)$$

he unbalance of the differential amplifier becomes:

$$\delta = (1/A_v)+(R_1/2Z_1)+(1/A)\,[(R/Z_c)+\alpha+\epsilon+\rho]$$

with the gain:

$$A \simeq \frac{R_o + R_1}{R_1}$$

The fundamental difference between this and the two previous circuits is that now the leakage impedance Z_1 appears explicitly in the unbalance, so that considerably higher values are required for Z_1 than before. The shields of the floating three-terminal amplifiers are directly connected to chassis, because no point in the circuit could serve as a low impedance source at the same potential as the interference. The characteristics of the four-terminal amplifier (Fig. 4(b)) determined by the feed-through factor α and the coupling impedance Z_c appear in the unbalance only as divided by the gain A of the overall amplifier.

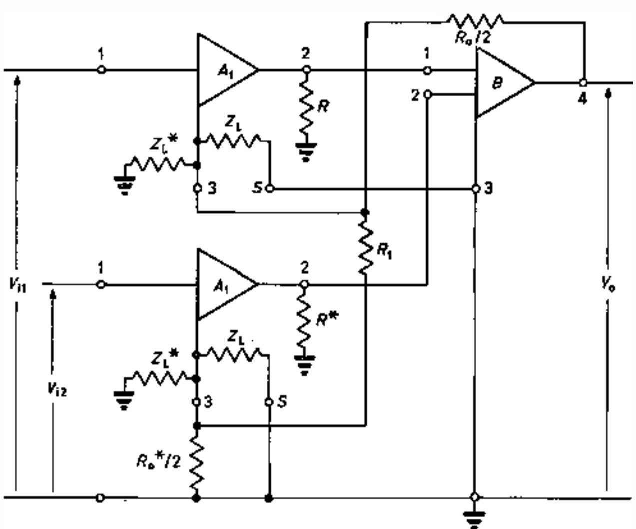


Fig. 7. D.C. amplifier, designed from two three-terminal and one four-terminal amplifier

With the input impedance:

$$R_{in} \simeq R_i(A_v/A)$$

the common-mode rejection of the differential amplifier in Fig. 7 becomes:

$$1/D = \frac{R_{s2} - R_{s1}}{R_i}(A/A_v)+(1/A_v)+(R_1/2Z_1)+(1/A)\,[(R/Z_c) + \alpha + \epsilon + \rho] \quad \ldots\ldots\ldots\ldots \quad (8)$$

according to equation (5).

Finally, Fig. 8 shows a differential amplifier composed of two four-terminal amplifiers; its design has a certain similarity to that of the circuit shown in Fig. 6.

With the error in bridge balance defined by:

$$R_o^*/R_1^* = R_o/R_1(1 + \epsilon)$$

the unbalance of the differential amplifier becomes:

$$\delta = (R_1/Z_c) + \alpha + (\epsilon/A)$$

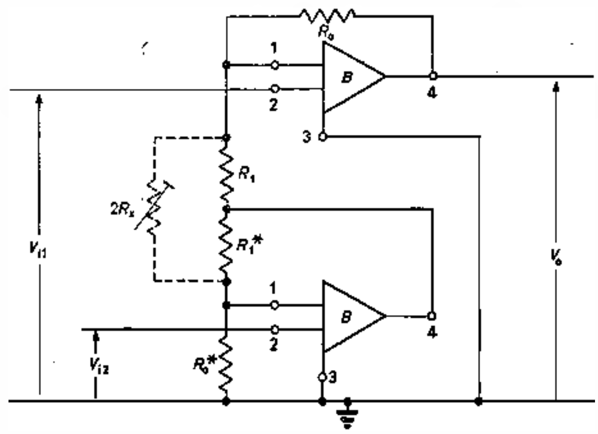


Fig. 8. D.C. amplifier, designed from two four-terminal amplifiers

The factor α and the coupling impedance Z_c of the four-terminal amplifiers appear in their entirety in the unbalance, whereas the bridge error once more appears as reduced by the gain. The required low α and R_1/Z_c values are relatively simple to achieve by means of the four-terminal amplifier designs proposed in Fig. 4. The gain is again:

$$A \simeq \frac{R_o + R_1}{R_1}$$

With the input impedance:

$$R_{in} \simeq R_i(A_v/A)$$

the common-mode rejection factor is:

$$1/D = \frac{R_{s2} - R_{s1}}{R_i}\,.\,(A/A_v)+(R_1/Z_c)+\alpha+(\epsilon/A) \quad \ldots\ldots \quad (9)$$

according to equation (5).

If the gain A of an amplifier shown in the circuit of Fig. 8 is to be variable, then in principle two resistances R_o and R_o^* or R_1 and R_1^* must always be adjusted in order to maintain the relation $R_o^*/R_1^* = R_o/R_1$. The connexion shown by means of a dashed line in Fig. 8, where only one resistance R_x is varied, is a simpler solution. With $R_o^* = R_o$ and $R_1^* = R_1$, the gain:

$$A = \frac{R_o + R_1}{R_1} + R_o/R_x$$

is variable between a minimum $(R_o + R_1)/R_1$ (at $R_x = \infty$) and any desired maximum. The common-mode rejection is again given by equation (9) if $R_1R_x/(R_1 + R_x)$ is substituted for R_1.

Conclusions

Examination of the expressions for common mode rejection, given by equations (6) to (9), for the differential-amplifier circuits shown in Figs. 5 to 8, leads to the following conclusions:

In all four circuits the common-mode rejection contains a term which is determined by the transfer resistance R_1A_v and which increases with increasing gain A. For $R_1A_v >$

$10^{11}\Omega$, $A = 1\,000$ and $R_{s2} - R_{s1} < 100\Omega$, the contribution to $1/D$ from this term is less than 10^{-6}.

A second contribution, arising from the leakage impedances of the floating power supply and from the error in bridge balance, designated as δ and ϵ respectively, decreases with increasing gain; for $A = 1\,000$ its influence on $1/D$ is less than 10^{-6}, when ϵ, $\delta < 10^{-3}$, which can be readily achieved with the aid of precision resistors.

The most important term is that which is independent of A. Since the factor $1/A_v$ appears in the common mode rejection for all circuits with three-terminal amplifiers, a voltage gain $A_v > 10^6$ is required for $1/D < 10^{-6}$. For reasons of stability, the frequency response of the voltage gain, which enters into the loop gain of a negative feedback amplifier, must be such that the loop gain becomes less than unity before the phase rotates through 180° (Nyquist criterium). This also narrows down the bandwidth of the common mode rejection.

In the circuit shown in Fig. 7 the leakage impedance Z_l (rather than AZ_l) directly determines the common mode rejection, so that in this case an extremely good isolation from chassis is required. Moreover, the fact that this amplifier contains no internal, low-impedance point at the same potential as the interference voltage V will be a serious drawback in applications where the amplifier is connected to multiplexers.

Of course, it is not necessary that each individual term in the expression for $1/D$ be less than 10^{-6}; they can be mutually compensated for by means of trimmers. However, the lower the individual contributions initially, the simpler and more stable in time is such a compensation.

Consideration of the maximum output capabilities of the component amplifiers in the various circuits[2] leads to the result that all the differential amplifiers discussed can only be used for gain of $A > 8$. But this is a restriction of no practical importance.

The complexity of the differential amplifiers shown cannot be directly judged from the number of basic amplifiers involved, since these are not necessarily the same for different circuits or even within one circuit.

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A New Colour Television Camera

An entirely new, fully transistorized, colour television camera, the Mark VII, has been announced by The Marconi Co. Ltd. This camera, which uses four plumbicon pick-up tubes, represents a significant advance in stability of operation.

It is in fact, sufficiently stable for 'hands-off' operation from a simple control panel mounted in the studio suite. The use of plumbicon tubes has made it possible to design a comparatively small and light camera, which is easy to handle and simple to use.

Although the main market at the present time is in the U.S.A. and Canada, the Mark VII is aimed at markets throughout the world. The camera can be switched to either the 525 line or 625 line standards, and will provide signals suitable for coding to any of the systems that have been proposed—NTSC, PAL, or SECAM.

A major advantage of a four tube camera is that the colour pictures are not so dependent on accurate registration of the three separate colour tubes. In long term operation this is certain to produce better colour pictures on receivers in the home, as well as better compatible monochrome pictures.

Another major feature of the camera is a built-in 'relay' optical system, which enables the full range of standard image orthicon camera lenses to be used. This will considerably increase flexibility of operation in studios and outside broadcasts using both colour and monochrome cameras.

The camera is fully transistorized, with the exception of a single Nuvistor valve used in each of the four vision head amplifiers, where existing transistors could not be found to give a sufficiently high noise immunity.

The use of a separate camera tube for the 'luminance' channel, greatly reduces the dependence of the colour picture on extremely accurate registration of the pictures from the three colour tubes. The definition of the final colour picture is dominated by the single luminance channel.

The camera has been designed primarily to work with the same integrated zoom lens packages as the Mark V monochrome camera. It is suitable for both studio and outside broadcast work, and can be adapted for telecine use without difficulty.

The new camera follows the practice of previous Marconi black-and-white cameras, the Mark IV and the Mark V, in providing 'hands-off' operation.

The stability of the channel is such that after the main 'setting-up' adjustments for the complete channel have been carried out at the main camera control unit panel, operational control can be switched to a single panel, similar to the Mark V control panel, and containing only iris, master black level and master gain control of the complete channel, in addition to the on/off and capping controls.

In this type of operation, all four video chains have to be matched very closely, with very stable amplitude characteristics. Special circuit techniques, including the extensive use of thin film circuits, have been employed to ensure that all the video amplifiers, and in particular the gamma correction in each video chain, are matched with a high degree of stability to ensure that the colour balance is maintained without adjustment.

'Hands-off' operation enables the equipment to be set up outside programme hours when skilled personnel and proper equipment can be made available. Operation from a simple panel reduces the risk of maladjustment of setting-up controls during transmission time, when corrections would be extremely difficult.

The small size of the operational control panel, only $3\frac{1}{4} \times 8$in permits several to be mounted in a single control console. One operator can then control a number of cameras from the actual production suite, while the camera control units themselves can be removed to a central technical equipment area.

The Mark VII camera uses a relay optical system which enables the complete range of standard camera lenses used with image orthicon cameras to be applied to this colour camera. The relay system provides an optical path length within the camera in which the light splitting systems can be situated. It also permits the camera tubes to be mounted with their axes nearly parallel.

Two filter wheels are mounted between the external objective lens and the field lens of the camera relay system. Behind this field lens a light splitting prism surface is mounted. This surface reflects a suitable amount of light into the luminance tube of the camera and transmits the remainder to the colour tubes. An important feature of this surface is that it reflects only that amount of light which is necessary to provide the correct signal in the luminance tube. The maximum available light is thus transmitted to the colour tubes at all times.

This transmitted light is split into its red, green and blue components by prism type dichroic surfaces. This type of dichroic is completely free from ghost images and astigmatism. The three colour components pass from the dichroic system through colour trimming filters and a relay lens into the three chrominance camera tubes.

For black-and-white operation a fully reflecting surface is moved into position in place of the luminance/chrominance light splitting surface. This device passes all the incoming light to the luminance tube, and is controlled by a simple mechanical switch.

The Design of Cage-Driven M.F. Aerials

By P. Knight*, B.A., A.M.I.E.E.

If a medium-frequency mast-radiator is surrounded by a cage of wires connected to it at a point approximately a quarter of a wavelength above ground level, the base of the mast may be earthed and the bottom of the cage driven instead. The need for load-bearing base insulators is thereby avoided and the mast can readily be used to support v.h.f. or u.h.f. aerials, since isolating transformers are not required. The performance of the mast as an m.f. aerial is, however, slightly modified if the cage of wires does not form a perfect screen.

(Voir page 132 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 139)

THE use of mechanically-stressed insulators at the base of a medium-frequency mast or tower radiator can be avoided by driving the aerial with a concentric cage of wires. This arrangement, applied to a mast, is shown in Fig. 1; a similar arrangement can be used with a self-supporting tower. The base of the mast is earthed and the bottom of the cage is driven. The top of the cage, which is about 0·25λ long, is bonded to the mast. Ideally, the cage should form a perfect screen, but in practice this is not possible. The purpose of this article is to discuss the effect of the imperfect screening, and to present formulae to enable the performance of the aerial to be calculated. Throughout, the term 'mast' should be understood as referring either to masts of uniform cross-section or to tapered self-supporting towers, provided the cage surrounding the latter tapers in such a way that the ratio of the cross-sections of cage and tower remains unchanged with height.

Screening Factor

Before describing the behaviour of an imperfectly-screened mast it is convenient to investigate the behaviour of a system of parallel conductors and to define a quantity called the screening factor.

Fig. 2 shows a cross-section of a mast surrounded by a cage of wires. Since all the wires are effectively connected in parallel this may be regarded as a two-conductor system, the mast forming an inner conductor and the cage an outer conductor.

If both conductors are connected in parallel and driven by a generator† the current flowing from the generator will divide between the two conductors, the proportion of the current flowing on the outer conductor being greatest when the latter most effectively screens the inner conductor. The current ratio may therefore be regarded as a measure of the effectiveness of the outer conductor as a screen, a factor k being defined in terms of the current ratio as follows:

$$k = \frac{\text{current on inner conductor}}{\text{current on outer conductor}} \quad \ldots\ldots\ldots\ldots (1)$$

It may be shown that k also defines the ratio of the voltages to ground which would appear on the conductors if they were used as a transmission line and carried equal and opposite currents. There is, therefore, an alternative definition:

$$k = -\frac{\text{voltage on outer conductor}}{\text{voltage on inner conductor}} \quad \ldots\ldots\ldots\ldots (2)$$

The minus sign appears in the definition because the voltages are in antiphase.

* *British Broadcasting Corporation.*

† *This could be achieved by insulating the base of the mast, connecting the bottom of the cage to the mast and driving the parallel combination.*

It may also be shown that, if the conductors are of finite length and well insulated, k is the ratio in which an electrostatic charge divides between them if they are charged to the same potential, and the ratio between their potentials if they carry equal and opposite charges. This

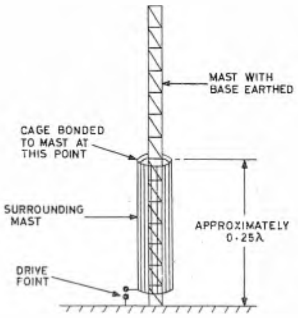


Fig. 1. A cage-driven mast-radiator

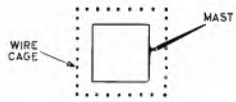


Fig. 2. Cross-section of lower part of aerial

property enables the factor k for a system of conductors to be calculated from electrostatic principles. It may also be measured with models, using a capacitance bridge in the manner described in Appendix (2).

When considering the performance of a mast surrounded by a cage of wires it is more convenient to describe the screening properties of the cage in terms of a screening factor s, defined as follows:

$$s = \frac{\text{current on outer conductor}}{\text{total current}} = \frac{1}{1+k} \quad \ldots\ldots (3)$$

Since k may have any value between zero and infinity, the value of s must lie between 0 and 1, the value $s = 1$ corresponding to a perfect screen.

The formula for the screening factor of a mast surrounded by a cylindrical wire cage is given in Appendix (1).

The Performance of a Cage-Driven Aerial

Fig. 3(a) is the equivalent circuit of a cage-driven aerial, the left-hand conductor representing the cage and the right-hand conductor the base-earthed mast. The performance of the aerial may be analysed by a method similar to that used for the folded dipole. The generator in Fig. 3(a) is replaced by two generators, and two opposing generators are placed in the earthed conductor, as shown in Fig. 3(b); arrows indicate the polarity of the generators. This arrangement is then separated into radiating and non-radiating modes, as shown in Figs. 3(c) and 3(d); this is made possible by the particular choice of generator voltages adopted*.

* *In the analysis of the folded dipole the two conductors are usually assumed to be identical, in which case $k = 1$ and the four generators of Fig. 3(b) are equal. However, the conductors need not be identical and k may have any value between 0 and ∞.*

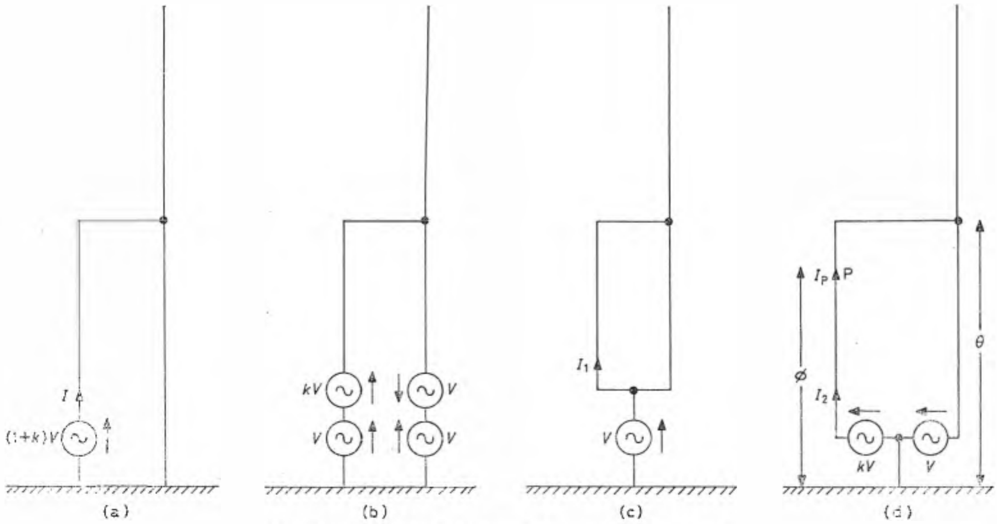


Fig. 3. The equivalent circuit of a cage-driven aerial

current is the same as the current which would flow if the mast were base driven; it divides between the cage and the mast, the fraction flowing in the cage being equal to the screening factor. Thus when s is close to unity the greater proportion of the radiating current flows on the screen.

The non-radiating current flowing at a point P (Fig. 3(d)) in the cage is:

$$I_P = -j(1+k)(V/Z_0)\frac{\cos(\theta-\phi)}{\sin\theta} = -j(V/sZ_0)\frac{\cos(\theta-\phi)}{\sin\theta} \quad \text{............ (7)}$$

where θ is the electrical length of the cage, ϕ is the electrical height of the point P above the ground and Z_0 is the characteristic impedance of the transmission line formed by the mast and the cage*. The current is greatest at the top of the cage, where its magnitude is equal to $V/(sZ_0\sin\theta)$.

Input Impedance

The input current and hence the input impedance may now be determined. In Fig. 3(c) the total current flowing from the generator is V/Z where Z is the impedance which the mast would have if it were base driven with cage attached. This current divides between the two conductors according to the ratio given by the factor k. Thus the current in the left-hand conductor is:

$$I_1 = \frac{1}{1+k}V/Z \quad \text{............... (4)}$$

In Fig. 3(d) the current flowing in the left-hand conductor is:

$$I_2 = \frac{(1+k)V}{jX} \quad \text{.................. (5)}$$

where jX is the reactance of the short-circuited length of transmission line formed by the inner and outer conductors. If the length of the cage is 0·25λ, X is infinite. However, the cage need not be of this length and in some circumstances it may be advantageous to use a length slightly different from 0·25λ.

From equations (4) and (5) the total current flowing in the left-hand conductor can be found; this is equal to the input current I. Therefore:

$$I = I_1 + I_2 = \frac{1}{1+k}(V/Z) + (1+k)V/jX$$

$$= (1+k)V\left[\frac{1}{(1+k)^2 Z} + (1/jX)\right] = \frac{V}{s}\left[s^2/Z + 1/jX\right] \quad \text{........... (6)}$$

Since the generator voltage of Fig. 3(a) is $(1+k)V = V/s$ it follows that the input impedance of the aerial is equivalent to an impedance Z/s^2 in parallel with a reactance jX. Since the aerial would have an input impedance equal to Z if it were base-driven in the usual way, it will be seen that the effect of imperfect screening is to raise the input impedance (and therefore the drive voltage for a given power) by a factor $1/s^2$. Thus if high powers are contemplated, sufficient wires should be used to make the screening factor as close as possible to unity.

Current Distribution

The equivalent circuit of Fig. 3 also enables one to calculate the currents flowing in the cage and in the mast. From Fig. 3(c) it will be seen that the total radiating

It is of interest to compare the relative magnitudes of the radiating and non-radiating currents in a typical case. Suppose the mast is 0·5λ high; it then has a loop radiation resistance of 100Ω, and the maximum radiating current for a power input of 100kW is 32A. If the mast is perfectly screened ($s = 1$) the input voltage will be about 8kV. Thus if the cage is 0·25λ high and has a characteristic impedance of 50Ω the current at the point where it is connected to the mast will be 160A; at this point the non-radiating current is five times greater than the radiating current. This ratio will be even greater if the mast is imperfectly screened, since the increased input voltage will cause a greater non-radiating current to flow; this increase, however, would be partly offset by a small increase in Z_0 caused by the imperfect screening.

Although the non-radiating current can be reduced by increasing the characteristic impedance of the transmission line formed by the mast and the cage, a Z_0 greater than 50Ω is undesirable because the relatively large size of the cage would degrade the performance of the aerial as a radiator. The aerial would then behave as a cylinder of greater diameter than the mast itself; this would reduce its power gain† and degrade its anti-fading properties.

Power Dissipated by Non-Radiating Currents

Since the non-radiating currents are large compared with the radiating currents, special care should be taken to minimize copper losses, both in the cage and in the section of the mast inside the cage. The cage should contain sufficient wires for its resistance to be small, and it should be well bonded to the mast at the point of connexion. It is also advisable to run copper strips down the faces of the portion of the mast inside the cage. In order to assess the number of wires and strips required, the following formula for the power loss due to the non-

* *The formula for the Z_0 of a mast surrounded by a cylindrical cage is given in Appendix 1.*

† *With a Z_0 of 50Ω the presence of the cage reduces the gain by about 0·2 to 0·3dB.*

radiating currents may be used:

$$P = (V/sZ_0)^2\,(\lambda r/4\pi)\ \frac{\theta + \sin\theta\cos\theta}{\sin^2\theta} \quad \ldots\ldots\ (8)$$

where V is the voltage with which the mast would be driven if it were base-insulated

s is the screening factor

Z_0 is the characteristic impedance of the line formed by the mast and the cage

λ is the wavelength

r is the h.f. resistance* per unit length of the series combination of cage and mast

θ is the electrical length of the cage in radians.

It may be shown that the power dissipated is a minimum when the cage is $\lambda/4$ long ($\theta = \pi/2$); it is then given by:

$$P_{min} = (V/sZ_0)^2\,\lambda r/8 \quad \ldots\ldots\ (9)$$

Although the optimum cage length for minimum power dissipation is $0{\cdot}25\lambda$, with masts higher than $0{\cdot}5\lambda$ there may be some advantage to be gained from the use of a shorter cage. This will have an inductive reactance which will partly cancel the capacitive reactance of the input impedance and may facilitate the matching of the aerial. The actual length used would depend on a compromise between reduced matching network loss and increased mast and cage copper loss.

Conclusions

There appears to be no fundamental difficulty involved in driving a base-earthed mast or tower as a medium-frequency aerial by means of a cage of wires; examples of such arrangements are to be found in several countries. The use of mechanically-stressed base insulators is avoided and an additional advantage is that the mast can be used as a support for v.h.f. or u.h.f. aerials. The cage should contain sufficient wires to effectively screen the mast or tower; if this is not done the input impedance and base voltage will be higher than it would be if the mast were base-insulated and driven in the usual way. The number of wires should also be sufficient to minimize copper losses; for the same reason, copper strips should be fitted to the portion of the mast or tower inside the cage. The width of the cage should not be appreciably greater than that of the structure it surrounds; if the cage is unduly wide the gain of the aerial will be reduced and its anti-fading properties degraded.

Acknowledgment

The author is indebted to the Director of Engineering of the BBC for permission to publish this article.

Appendix

(1) Formulae for the Screening Factor and Characteristic Impedance of a Conductor Surrounded by a Cylindrical Cage of Wires

Fig. 4 shows the arrangement considered. The formulae given below may be regarded as accurate provided the wire radius is small compared with the spacing between adjacent wires, and between the wires and the inner conductor. If the inner conductor is square or triangular in section the formulae may still be applied, provided the inner conductor is replaced by an equivalent cylinder whose radius is $0{\cdot}59w$ for a square or $0{\cdot}42w$ for a triangle, where w is the width of the faces of the inner conductor. Some error will occur, however, if the spacing between the corners of the square or triangle and the cage of wires is small.

Screening factor

$$s = \frac{\log(R/r_1)}{\log(R/r_1) + (1/n)\log(R/nr_2)} \quad \ldots\ldots\ (10)$$

Characteristic impedance of transmission line formed by inner conductor and cage

$$Z_0 = 60\log(R/r_1) + (60/n)\log(R/nR_2) \quad \ldots\ldots\ (11)$$

where r_1 = radius of inner conductor

r_2 = wire radius

R = radius of cage

n = number of wires

Equations (10) and (11) are valid provided both $r_1 n \ll Rn$ and $nr_2 \ll R$. It should be noted that the first term in the expression for Z_0 is the characteristic impedance which would result if the outer cage were a perfect screen; the second term may therefore be regarded as a correction term required by the imperfect screening.

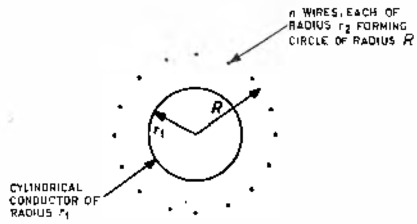


Fig. 4. A conductor surrounded by a cylindrical cage of wires

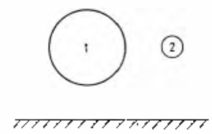


Fig. 5. Two parallel conductors and an earth plane

Example

Consider a square mast of 6ft side width surrounded by a cylindrical cage of radius 6ft. The cage consists of 32 wires each of ¼in diameter.

The radius of the cylinder equivalent to the square mast is $0{\cdot}59w$, where w = 6ft.

Therefore:

$$r_1 = 0{\cdot}59w \qquad R = w$$
$$r_2 = 0{\cdot}0017w \qquad n = 32$$

The conditions $r_1 n \ll Rn$ and $nr_2 \ll R$ are both satisfied.

Therefore:

$$s = \frac{\log 1{\cdot}70}{\log 1{\cdot}70 + (1/32)\log 18{\cdot}4} = 0{\cdot}85$$

$$Z_0 = 60\log 1{\cdot}70 + 1{\cdot}88\log 18{\cdot}4 = 31{\cdot}8 + 5{\cdot}5 = 37{\cdot}3\Omega$$

(2) The Measurement of the Screening Factor and Characteristic Impedance of a System of Parallel Conductors

The derivation of the formulae for the screening factor and characteristic impedance of arrangements of parallel conductors presents considerable difficulty in many cases and it is therefore often more convenient to measure these quantities, using a model of the proposed arrangement. Since the definitions of k (and therefore of s) given in equations (1) and (2) are independent of frequency, and apply also to static potentials and charges, measurements may be made at any convenient frequency. In the method described in this appendix, a model of a length of the conductor system is placed inside a large screening cage and measurements made with a capacitance bridge between various terminals.

Before describing this method it is necessary to consider the properties of parallel conductors in greater detail.

Fig. 5 represents two conductors (or groups of con-

* *Tables and formulae which enable the h.f. resistance of conductors of various shapes to be calculated are to be found in a number of references. See, for example, "Radio Engineering Handbook" by Terman, Section 2.*

ductors) insulated from each other and from earth. Following the notation used by Jeans*, the relation between the static charges and potentials of the conductors is given by:

$$\begin{aligned} Q_1 &= q_{11}V_1 + q_{12}V_2 \\ Q_2 &= q_{21}V_1 + q_{22}V_2 \end{aligned} \quad \text{.............. (12)}$$

where Q_1 and Q_2 are the charges on the conductors

V_1 and V_2 are the potentials of the conductors

q_{11} and q_{22} are the coefficients of self-capacitance

$q_{12} = q_{21}$ is the coefficient of mutual-capacitance

If $V_1 = V_2$, the ratio between the charges is equal to k and, if conductor 2 represents the inner conductor, then:

$$k = Q_2/Q_1 = \frac{q_{22} + q_{12}}{q_{11} + q_{12}} \quad \text{................. (13)}$$

The expression for the screening factor is therefore:

$$s = \frac{1}{1 + k} = \frac{q_{11} + q_{12}}{q_{11} + 2q_{12} + q_{22}} \quad \text{......... (14)}$$

To determine s, a suitable length of the two-conductor system is placed in a large earthed cage and the following capacitance measurements made:

(1) Between conductor 1 and earth (i.e. between conductor 1 and the large earthed cage) with conductor 2 earthed. Measured capacitance, C_1.

(2) Between conductor 2 and earth, with conductor 1 earthed. Measured capacitance, C_2.

(3) Between conductors 1 and 2 (connected together) and earth. Measured capacitance, C_3.

For measurement (1), $V_2 = 0$ and from equations (12) $q_{11} = Q_1/V_1 = C_1$. Similarly, from measurement (2), $q_{22} = C_2$. For measurement (3), $V_1 = V_2 = V$ and from equations (12):

$$C_3 = \frac{Q_1 + Q_2}{V} = q_{11} + 2q_{21} + q_{22} \quad \text{... (15)}$$

Algebraic manipulation involving equation (14) and the expressions for C_1, C_2 and C_3 then leads to the following expression for the screening factor:

$$s = \frac{C_1 - C_2 + C_3}{2C_3} \quad \text{............... (16)}$$

The characteristic impedance of the transmission line formed by the two conductors may be found by measuring the capacitance between the two conductors. If the two conductors are similar a balanced bridge should be used and the large earthed cage removed, but if one conductor substantially screens the other the outer conductor should be earthed; the presence of the outer cage then has an insignificant effect on the measurement. The characteristic impedance may be calculated from the expression:

$$Z_o = 10^4/3C$$

where Z_o is in ohms and C is the measured capacitance in picofarads per metre.

* *See "Magnetism and Electricity" by Jeans, Chapter IV. In the book, E is used for electric charge instead of Q. Jeans also uses the terms "coefficients of capacitance, coefficients of induction" for the coefficients of self- and mutual-capacitance.*

A Computer-Based Communications System

A new computer-based communications system has been introduced by the Data Systems Division of Standard Telephones and Cables Ltd. Known as the 6300 ADX, it is an on-line stored-program system designed specifically to operate on a real-time basis in telecommunication systems having up to 60 duplex lines.

A message received at a typical station in, for example, a telegraph network, may need to be re-transmitted to several other stations. These in turn may undertake re-routing and processing and may themselves be connected in a complex pattern. With increasing traffic it is becoming evident that conventional methods of handling messages and data at these switching centres can waste time and money. For instance, in 'torn-tape' centres incoming message tapes are torn from a machine by an operator who duplicates on other machines and takes the copies to a number of transmitters for re-transmission to further destinations. STC's ADX (automatic data exchange) controls each relay centre according to a stored-program and automatically routes messages, data or enquiries to their destinations without human intervention, whether they arise from manual keyboards, automatic machines or from other computers.

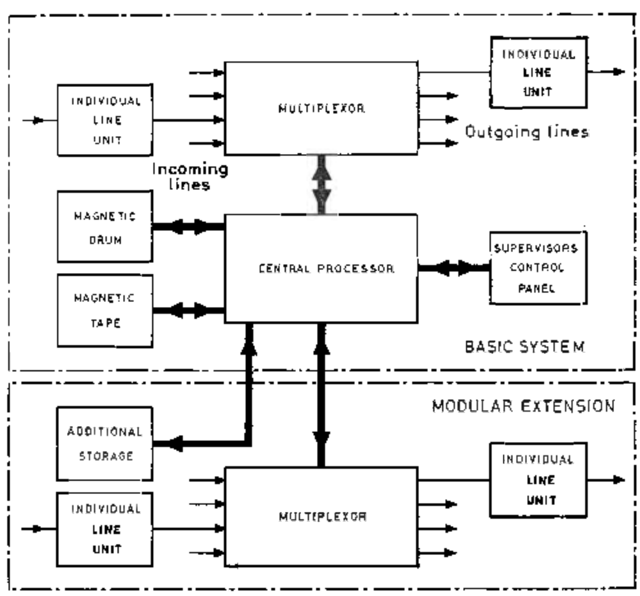
The 6300 ADX performs these tasks automatically and electronically. In the case of telegraph system relay centres, there are no telegraph machines to send and receive messages and no tapes. Instead, the incoming lines are sensed electronically and the information stored in magnetic storage systems. The stored signals then drive the outgoing lines electronically. Routing and queuing decisions are performed by computer program at electronic speed. There is no need for each message to be fully received before it is processed and placed on the outgoing queue.

The 6300 ADX will minimize delays to information passing through the network. It will also allow maximum communication circuit utilization leading in turn to reduced operating costs. Flexibility of operation is provided by the stored-program technique and the system is fully compatible with established network features and procedures. The 6300 ADX has a high standard of operational efficiency and system reliability.

Typical applications of the system are: telegraph message and data switching; message collection and distribution; man-computer information networks; integration of multiple computer networks; message accounting and traffic statistics; stock control systems; reservation systems.

STC is able to provide a tailor-made system to meet any specific requirements for hardware, software and operating procedures.

Basic arrangement of the system



Design of a Current Source and Isolation Transformer for Electrophysiological Experiments

By A. van Oosterom*

This article describes an output stage generating current impulses in a range from 10μA to 25mA. Some design considerations are given for a transformer, which has been developed for this source, accepting square impulses up to 12mA and 12msec duration into loads up to 10kΩ.

(Voir page 133 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 140)

An important field of electrophysiology is the investigation of reactions of nerves and muscles to electrical stimuli. These studies are—among others—performed by passing a current through, or by applying a voltage across, the tissue. In many cases a current source is used in order to make the strength of stimulation independent of the resistance between the stimulating electrodes and the tissue concerned. This method has also been chosen for the experiments at the cardiological department of the "Wilhelmina Gasthuis" (University of Amsterdam) for which the circuit to be described has been developed.

The current source is electrically isolated from the heart by means of a screened 'current transformer'. In this way the so-called 'stimulus artefact', a signal directly generated by the stimulus, is reduced.

Problems arise, however, from the fact that if square current impulses are fed into a transformer the core may be saturated. A technique is discussed and results are given for reducing this effect by leading a compensating current through an extra winding on the transformer.

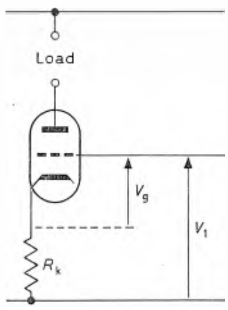


Fig. 1. Basic circuit for obtaining high impedances; V_g the cathode to grid potential

The Current Source

General

Basically a current source has an infinite output impedance. Practically this means an output impedance which is high compared with the impedance of the external load. A way to achieve a high output impedance in a valve circuit is to put a high resistance (R_k) in the cathode lead and to raise the grid to a potential V_1 (Fig. 1).

The well-known formula for the output impedance is:

$$Z_{out} = R_i + (\mu_1 + 1)\,R_k$$

with, respectively, R_i and μ_1 the internal resistance and the amplification factor of the valve.

If $(\mu_1 + 1)\,R_k \gg R_i$ and $\mu_1 \gg 1$ then $Z_{out} \simeq \mu_1\,R_k$. (1)

The current equals:

$$I_{out} = \frac{V_1 - V_g}{R_k}$$

V_g being the grid to cathode potential.

If

$$V_1 \gg V_g \text{ then } I_{out} = V_1/R_k \quad \dots\dots\dots\dots \quad (2)$$

From equations (1) and (2) it can be seen that if a high output impedance is required at high output currents, the voltage V_1 must be high too:

$$V_1 = \frac{I_{out}\,Z_{out}}{\mu_1} \quad \dots\dots\dots\dots\dots\dots \quad (3)$$

The problem of having a high V_1 and consequently of having a high supply voltage can in certain cases be avoided by using a cascode stage (Fig. 2).

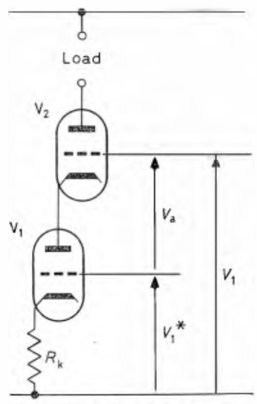


Fig. 2. Cascode circuit; V_a the anode voltage of V_1

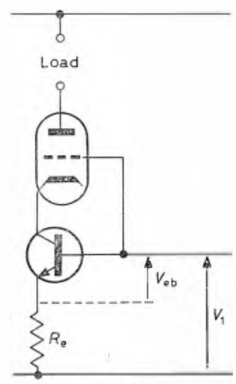


Fig. 3. Hybrid circuit; V_{eb} the emitter to base voltage

The output current is the same as in the circuit of Fig. 1 (equation (2)) whereas the output impedance equals:

$$Z_{out} = R_{i2} + (\mu_2 + 1)\ \{R_{i1} + (\mu_1 + 1)R_k\}\,.$$

Assuming:

$(\mu_1 + 1)\,R_k \gg R_{i1}$; $\mu_1 \gg 1$; $\mu_2 \gg 1$ and $\mu_1\mu_2 R_k \gg R_{i2}$,

leads to:

$$Z_{out} = \mu_1\mu_2 R_k \quad \dots\dots\dots\dots\dots\dots \quad (4)$$

Combining equations (2) and (4) gives:

$$V_1' = \frac{I_{out}\,Z_{out}}{\mu_1\mu_2} \quad \dots\dots\dots\dots\dots\dots \quad (5)$$

The voltage V_1 at the grid of the upper valve (Fig. 2) is:

$$V_1 = V_1' + V_a = \frac{I_{out}Z_{out}}{\mu_1\mu_2} + V_a \quad \dots\dots \quad (6)$$

with V_a the anode to cathode voltage of V_1. For equal values of V_1 one can compare the output impedances of both circuits:

$$Z_{1(out)} = \frac{\mu_1 V_1}{I_{out}} \qquad \text{(see equation (3))},$$

$$Z_{2(out)} = \frac{\mu_1\mu_2(V_1 - V_a)}{I_{out}} \qquad \text{(see equation (6))}.$$

The advantage of the cascode stage can be expressed by the ratio:

$$\frac{Z_{2(out)}}{Z_{1(out)}} = \mu_2\,(1 - V_a/V_1).$$

* *Medical Physics Laboratory, University of Amsterdam.*

The development of silicon npn transistors has made it possible to devise a circuit with a high output impedance requiring a much lower supply voltage (Fig. 3). In this circuit a transistor acts as a cathode resistor. The equivalent resistance is the collector impedance of the transistor with grounded base.

This impedance can be calculated using the equivalent circuit of Fig. 4.

$$I_1(R_e + r_e + r_b) + I_2 r_b = 0, \quad (7)$$

and

$$I_1(r_b + \alpha r_c) + I_2(r_b + r_c) V \quad (8)$$

Combining equations (7) and (8) gives:

$$Z = V/I_2 = \frac{(R_e + r_e) r_b/r_c + R_e + r_e + (1-\alpha) r_b}{R_e + r_e + r_b} r_c$$

In a transistor the following applies $r_e \ll r_b$ and $r_c \gg r_b$. If furthermore $R_e \gg r_b$: $Z = (r_c + r_b) \simeq r_c$.

Now this impedance represents the cathode impedance of the valve in the circuit depicted in Fig. 3. Substitution

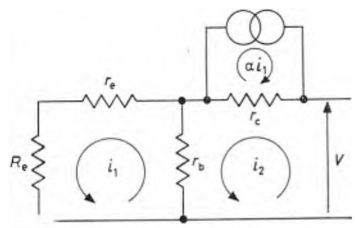


Fig. 4. Equivalent circuit of the transistor in Fig. 3

of this value in equation (1) for R_k gives the total output impedance of the circuit:

$$Z_{out} = \mu_1 Z_k = \mu_1 r_c.$$

The collector of the transistor (Fig. 3) is:

$$I_{out} = \frac{V_1 - V_{eb}}{R_e}\alpha.$$

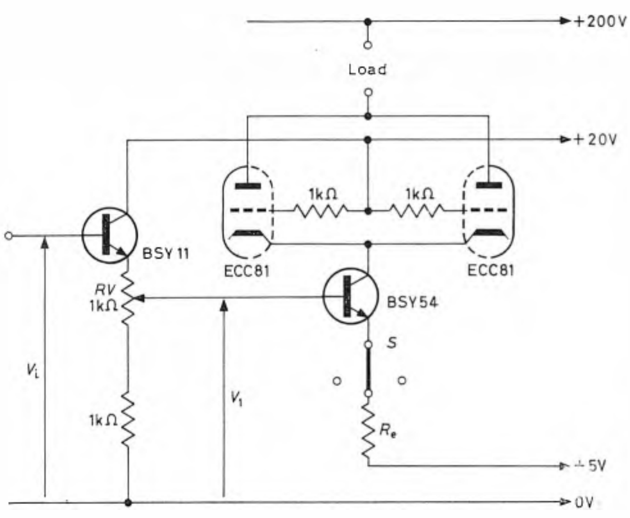
If $\alpha \simeq 1$ and $V_1 \gg V_{eb}$ one has $I_{out} = V_1/R_e$ (compare equation (2)).

Practical Circuit

An output stage based on the principle just described is shown in Fig. 5.

The output current (neglecting V_{eb}) is $V_1 - 5/R_e$. For input voltages $V_i < 5V$, the transistor VT_1 is cut off. During this state the output current is equal to the collec-

Fig. 5. Complete hybrid output stage with emitter-follower input



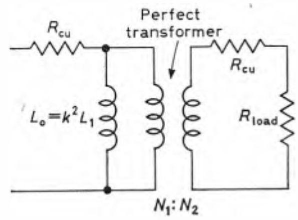


Fig. 6. Equivalent circuit of transformer

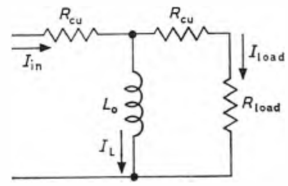


Fig. 7. Simplified circuit of Fig. 6

tor to base leakage current I_{cbo}. Silicon transistors with values of I_{cbo} of less than 20nA are easily obtainable and this value is, for most applications, small enough to be neglected.

With switch S different values of R_e can be selected giving a wide range of I_{out}. Potentiometer RV provides the vernier control. The maximum output current is restricted by the power limitations of the valve (for ECC81 P_a max $= 2 \times 2{\cdot}5 = 5$W) whereas the lower limit is set by the restriction:

$$I_{cbo} < 1\%\, I_{out} \qquad \text{(for type BSY54: } I_{cbo} < 20\text{nA).}$$

In the circuit given in Fig. 5, having a range from 25mA (for a supply voltage $V_b = 200$V: $P_a = 25{\cdot}10^{-3} \times 200 = 5$W) to 10$\mu$A, the current is within these limits.

The output impedance has been measured at different values of I_{out} and results are given in Table 1.

Under practical circumstances the load resistance is usually less than 10kΩ so when the circuit is considered as a current source, the error made is never more than 0·5 per cent.

The triodes can deliver the maximum output current at an anode voltage of 150V. For a supply voltage of 200V, the maximum allowable voltage swing across the load at maximum current is therefore 50V. At lower output currents the triodes need less anode voltage and the swing can be higher with an absolute maximum of 150V at the minimum output current. If the circuit is to be used for

TABLE 1

Values of Z_{out} at Different Output Currents

I (mA)	Z_{out}(MΩ)
20	11
7·5	55
2	208
0·75	590

pulse output only, power will be dissipated during the working state and so the supply voltages can be raised to $V_{sup} = 200/d$ where d is the maximum duty cycle of impulses. Of course limiting values of the anode voltage must be taken into account.

The Transformer

General

A transformer had to be designed, capable of handling square current impulses of either positive or negative sign with values up to 10mA and durations ranging from 20μsec up to 10msec. This problem has been solved by using the theory of equivalent circuits[1]. According to this theory the behaviour of a transformer can be calculated by considering it to be 'perfect', i.e., having an impedance transfer ratio equal to the squared ratio of the

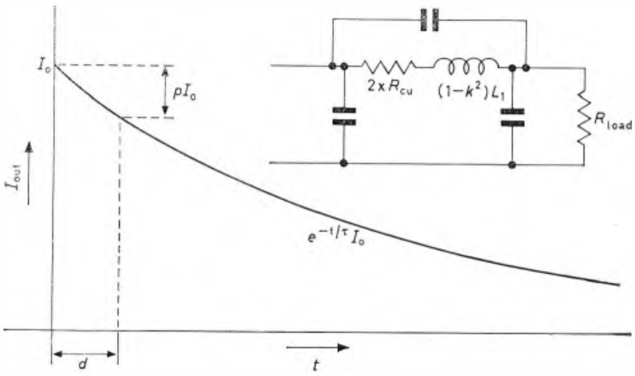


Fig. 8. Exponential decay of I_{out}

Fig. 9. (Inset) An equivalent circuit for a transformer at high frequencies

number of turns, and subsequently adding the imperfections caused by the physical limitations. In the equivalent circuit of Fig. 6 the influences of the copper resistance R_{cu}, the number of turns N and the coupling factor k have been included. L_1 represents the inductance of the primary winding.

As the current source had to be used with and without transformer, the turns ratio equals one.

To study under this condition the influence of the imperfections on the output waveform the ideal transformer T can be omitted from the equivalent circuit, which simplies thus to that of Fig. 7.

In order to find the required primary inductance the output current at the application of a step input current I_o is considered:

$$I_{out}(t) = I_o\, e^{-t/\tau}; \qquad \text{with } \tau = \frac{L_o}{R_{cu} + R_{load}}$$

If the current impulses have a duration d and allowing a given percentage of say p (see Fig. 8), the inductance L_o for given values of R_{load} can be calculated as follows:

$$I_{out}\,(t = d) = I_o\, e^{-d/\tau} = (1 - (p/100))\, I_o;$$

for $d \ll \tau$, $I_{out}(t)$ can be approximated by

$$I_{out}\,(t = d) = I_o\,(1 - d/\tau) = (1 - (p/100))\, I_o,$$

$$\text{hence } \tau = \frac{100\, d}{p}$$

$$\text{and } L_o = \frac{100\, d\,(R_{cu} + R_{load})}{p}$$

The requirements for this particular design are: $d_{max} =$ 10msec, $p = 10$ per cent and $R_{load(max)} = 10\text{k}\Omega$. The minimum value of L_o is therefore:

$$L_o = \frac{10^3 \cdot 10^4 \cdot 10^{-2}}{10} = 1\,000\text{H} \quad \ldots\ldots \text{(9)}$$

Now the general formula for an inductance can be approximated by:

$$L = \frac{\mu_o \mu_r N^2 A}{l}\,[\text{H}] \quad \ldots\ldots\ldots\ldots \text{(10)}$$

(neglecting the air-gap)

where μ_o = permeability in vacuum $\simeq 1{\cdot}25 \times 10^{-6}$ [H/m]

μ_r = permeability of core material

A = cross-sectional area $[m^2]$

l = magnetic path length [m]

N = number of turns.

The elements A and l are restricted by the dimensions of the obtainable core material.

Before the required values of N and μ_r can be estimated some other transformer imperfections must be taken into account. These are the stray capacitances between turns and windings and the leakage inductance. These elements form a resonant circuit. If a step current is applied to the transformer the resonant circuit causes a damped oscillation superimposed on the output waveform.

An equivalent circuit (approximating the resonant circuit) is shown in Fig. 9.

If $L_{leak} = (1 - k^2)L_1$ is made small, the influence of these oscillations are minimized. For given values of L_1, L_{leak} can be kept small by making k as high possible. (For $k = 1$: $L_{leak} = 0$ and $L_1 = L_o$). This can be achieved by special arrangements and by using core material with a high μ_r.

Summarizing the problem: to get a high $L_o = k^2 L_1$, L_1 must be high, but to obtain a low $L_{leak} = (1 - k^2)L_1$, L_1 must be low and μ_r high with respect to k. As $L_1 \sim \mu_r N^2$, the best compromise is the use of core material with the highest value of μ_r that can be obtained and choosing N to meet the requirement for L. This leads to the need of using Mumetal for the core material. Unfortunately, however, the permeability of Mumetal decreases rapidly in the presence of a polarizing field, and the core may be saturated. The influence on the equivalent circuit of Fig. 7 is that L_o becomes dependent on I_L, being the current through L_o, i.e., the difference between input and output current. Fig. 10 gives a typical curve for L as a function of L_1. This curve is derived by simultaneously passing a small alternating and direct current through the primary winding of the transformer. The voltage generated across the primary winding is used to calculate L_o.

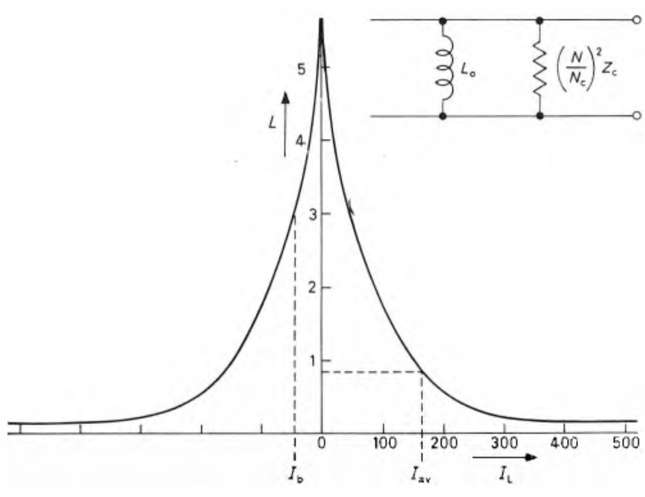
As this curve is different for different frequencies[1], it cannot be used for quantitative analysis of the output current at the application of a step input current, which can be considered as consisting of a wide range of frequencies. The required number of turns must therefore be found empirically.

Now the number of turns can be reduced considerably by leading a compensating current of opposed direction through an extra winding (N_c turns) of the transformer. To evaluate its effect, two cases must be considered:

(a) One single square current impulse with height I_o and duration d.

Fig. 10. Values of L for different polarizing fields

Fig. 11. (Inset) Equivalent circuit including impedance of source for compensating current



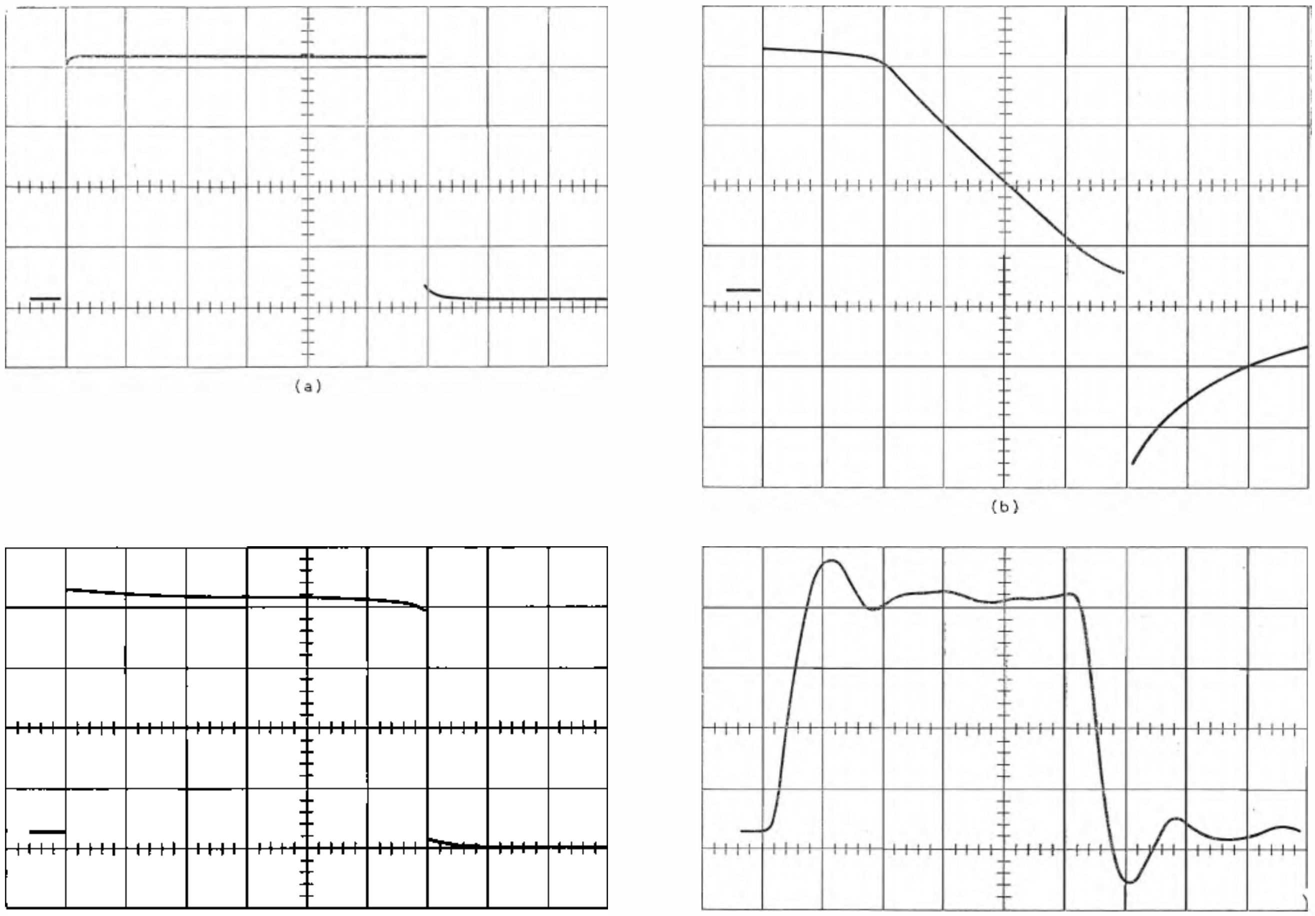


Fig. 12. Output of complete source current at $f = 2c/s$ $R_{load} = 10k\Omega$ and $I_{out} = 12mA$

(a) d = 12msec—no transformer
(b) d = 12msec—with transformer
(c) d = 12msec—with compensated transformer
(d) d = 100µsec—with transformer (expanded time-base)

(b) Square wave current impulses with height I_o, duration d and repetition frequency f.

(a) Just after the current impulse has been applied, the current through the transformer $I_L = I_o(1 - e^{-t/\tau})$ is small, L_o is near its maximum and the output current equals the expected value $I_{out} = I_o e^{t/\tau}$. But as can be seen from Fig. 10 the more I_L increases the more L_o decreases, thus causing a rapid fall of I_{out}.

This effect can be partly reduced by using a small direct current (I_b in Fig. 10) to bias L_o. This causes the inductance at zero input to be lower; but instead of decreasing when a current flows through the transformer, the inductance increases for a while, thus enabling to transform pulses at a lower value of L_o. The correct value of I_b must be found experimentally.

(b) In this case a current with an average value $I_{av} = I_o . d . f$ goes into the transformer. This current will flow through L_o as can be seen from Fig. 7. The inductance L_o at this value of L_L may be much lower (Fig. 10).

Leading a current of $N/N_c(I_{av} + I_b)$ through the extra winding will bring the inductance back around the optimum average value.

Special care must be taken that the current through the extra winding is delivered by a source having sufficiently high impedance. This source impedance (Z_s) will shunt the current source. Therefore the total output impedance of the current source with compensated transformer will be that of the circuit in Fig. 11.

Practical Example

The transformer had to have an inductance in the order of $L = 1\,000$H. Core material with a magnetic pathlength of $9{\cdot}6 \times 10^2$m could be obtained. A compromise between cross sectional area and stack with regard to copper losses resulted in a cross area of $7{\cdot}6 \times 10^{-4}\,m^2$. The manufacturer's estimation of μ_r indicated values in the order of 17 000. From equation (10) follows that with these figures N should be about 2 300. Because of the saturation effect $N = 3\,000$ was taken. This gives for L at zero current $L = 1\,600$H. N_c has been taken rather arbitrarily as 300 turns.

The windings are arranged in six sections (not including the extra winding) to obtain a high k. The parts of the secondary winding are placed in such a way so as to get equal stray capacitances to earth. Electrostatic screens of copper foils are positioned between the windings.

The transformer can be obtained from Unitran N.V. Weesp-Hallond under type number R19E-RH11.

Results

Four different cases of the output pulse form of the complete current source with $f = 2c/s$, $R_{load} = 10k\Omega$ and $I_{out} = 12mA$ are given in Fig. 12.

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A Beam-Blanking Circuit

By A. M. Andrew

A circuit is described which conveys continuously-variable signals with high accuracy between points having a large steady potential difference. The frequency response is level from zero to a very high value. It does not employ a radio-frequency oscillator, and is simple to construct.

(Voir page 133 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 140)

IN various electronic devices there is a need to convey signals between points having a large steady potential difference. In particular, it is often necessary to control the grid of a cathode-ray tube, where the control signals are generated by circuits near to earth potential, and the grid potential is near to that of the c.r.t. cathode. When the signal to be conveyed contains no slow components simple capacitance coupling can be used. When slow components must be transmitted the usual method is to

Fig. 1. Waveform which can be transmitted accurately by a capacitance coupling with restoring diode

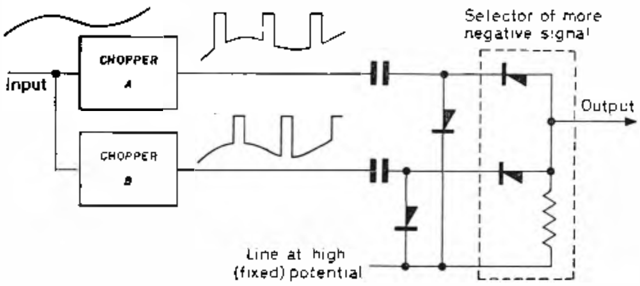

Fig. 2. Transmission of an arbitrary waveform using two choppers

let the incoming signals modulate a radio-frequency oscillator. The output of this is conveyed either capacitively or inductively to the destination of the signals, then rectified and smoothed. An alternative method is described here. First, however, a method is described which has not been put into practice but which illustrates the principles used.

Signals of the form shown in Fig. 1, which return at intervals to a fixed potential outside their normal range of variation, can be transmitted by a capacitance coupling with a restoring diode. The effect is that of d.c. coupling provided the time-constant of the coupling is long compared to the intervals between the returns to the fixed potential.

Any signal can be chopped to form a signal like that of Fig. 1. A small amount of information is, of course, lost in the chopped-out part. However, no information is lost if two separate chopped signals are produced in such a way that the choppings never overlap. Then these two signals can be conveyed to the high-potential point by capacitance couplings with restoring diodes as shown in Fig. 2. The original signal can be recovered by selecting at any instant the more negative of the two waveforms, as shown (or the more positive, if the waveform was chopped so as to return to a potential more negative than its normal range of variation).

The method represented in Fig. 2 is perfectly feasible but a more economical one has been used in practice. Only one chopped waveform is used, in conjunction with the original signal, as shown in Fig. 3. It will be assumed here that the chopping introduces a fixed potential more positive than the signal.

The two waveforms are passed through capacitors and recombined as shown in Fig. 3. Provided the time-constants of C_1 and C_2 with R are long compared to the chopping interval, the signal is conveyed to point X with, effectively, d.c. coupling.

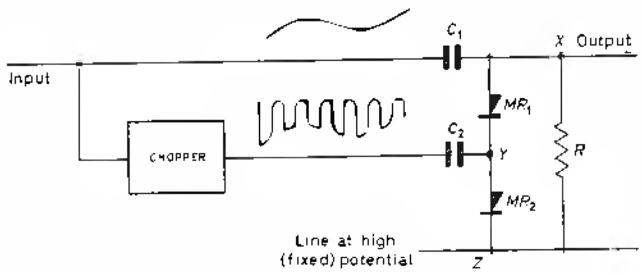

Fig. 3. Transmission of an arbitrary waveform using one chopper

Fig. 4 illustrates how the capacitor C_2 acts with MR_1 and MR_2 as a 'diode pump' to pump charge from point Z to point X to compensate for the leakage in R. When the chopped waveform goes positive, point Y is prevented by MR_2 from going more positive than point Z, except for the small transient shown. When the chopped waveform goes negative again, the charge on C_2 is shared with C_1, through MR_1, and this brings the potential of point X closer to the required signal value. The effects of leakage are greatly exaggerated in Fig. 4.

Practical Circuit

Fig. 5 shows a practical circuit for coupling to the grid of a cathode-ray tube. V_{1a} is used as a diode to eliminate positive-going excursions of the input waveform, and V_{1b} is a simple cathode-follower. The unchopped signal for transmission to the high-potential circuit is taken from V_{1b} cathode. V_3 is a free-running multivibrator whose grid waveform has negligible positive overshoot because of the high values of the anode resistors. The waveform at one of the grids of V_3 is also applied to a grid of V_2, whose other grid receives the input waveform. V_2 forms

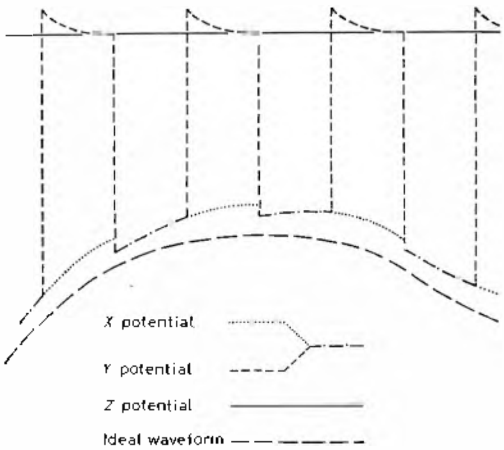

Fig. 4. Waveforms illustrating the operation of Fig. 3
Effects of leakage are greatly exaggerated

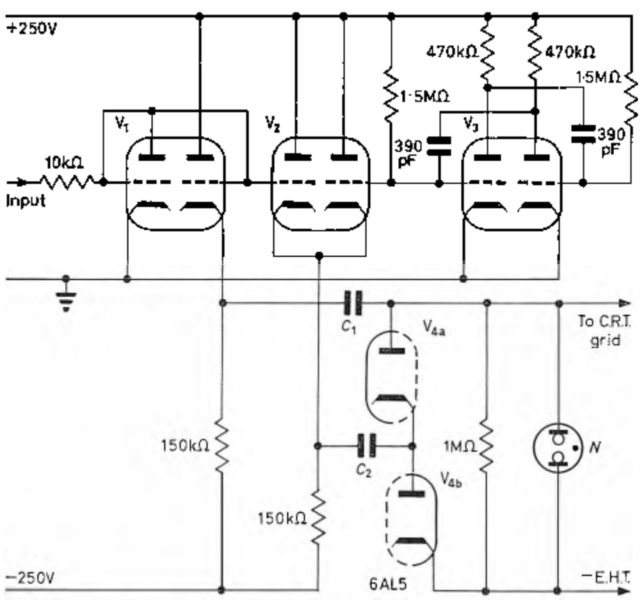

Fig. 5. Practical circuit

a pair of cathode-followers with common cathode resistor, and therefore selects at any instant the more positive of its two input waveforms. The result is to produce the required chopped waveform at its cathodes.

Thermionic diodes (6AL5) are shown in the high-potential part of the circuit. A high ratio of forward to inverse conductivity is needed. Silicon junction diodes have been used satisfactorily, but most germanium diodes are not suitable. There is usually no difficulty in providing a heater supply for thermionic diodes as this can be common with the c.r.t. heater supply.

The incorporation of a small neon tube across the output of the circuit is important. If it is not included the diodes may be damaged when the oscilloscope is switched off, since the coupling capacitors are likely to discharge more slowly than those in the c.r.t. power supply. The ignition voltage of the neon should be sufficiently high that it is inoperative during normal operation.

Performance

With the component values shown, the frequency of the multivibrator is about 2 500c/s. Operation does not depend critically on valve characteristics, and the circuit has sometimes been used with valves type 12AT7, and sometimes 12AX7. The latter are slightly preferable and the following performance figures apply to them.

The output of the circuit contains a certain amount of ripple at the multivibrator frequency. This is a combination of the sawtooth ripple represented in Fig. 4 and spikes due to effects of stray capacitance. The total peak-to-peak ripple was as follows for $C_1 = C_2 = 0{\cdot}1\mu F$:

Signal voltage (negative)	0	10	20	50	100
Ripple	0·04	0·05	0·07	0·18	0·50

For $C_1 = C_2 = 0{\cdot}5\mu F$, the total ripple was somewhat greater because the capacitors were physically larger, with higher strays, and the spiky component was increased. With $C_1 = C_2 = 0{\cdot}01\mu F$, the total ripple was again greater because the sawtooth component was increased.

The error of representation of the input voltage at the output was less than 2V over the range of negative input voltage from zero to 100. The error changes sign at 28V:

Input voltage (negative)	0	10	28	50	100
Output voltage (negative)	1·3	10·4	28·0	49·7	98·1

Measurements of output voltage were made to the middle of the ripple.

The high-frequency response of the circuit is excellent, since high frequencies are conveyed directly from input to output through the cathode-follower V_{1b} and capacitor C_1. It has been verified that a 150Mc/s square wave is transmitted with negligible distortion.

Discussion

Compared with the radio-frequency coupled circuits mentioned earlier, the circuit has several advantages and disadvantages. One advantage which the author has appreciated is that the components are all resistors and capacitors from the drawer, with no need to select or construct a radio-frequency coil. On the other hand the circuit contains more valves than a radio-frequency unit. A major disadvantage when used to modulate the brightness of a c.r.t. is that the brightness is affected by any rapid fluctuations in the e.h.t. supply voltage due to mains ripple, mains fluctuations, or variable leakage in a faulty smoothing capacitor. In practice, however, e.h.t. supplies are usually sufficiently steady to avoid trouble from this source. The most impressive features of the performance of the circuit are the extremely precise reproduction of the input signals with unity gain, and the wide frequency coverage from zero to hundreds of megacycles. Neither of these features is essential in an oscilloscope beam-blanking circuit, however, and the circuit may find more valuable applications elsewhere.

Acknowledgments

The circuit was devised when the writer was with the group headed by Dr. W. S. McCulloch at the Massachusetts Institute of Technology. This article has been prepared while a visitor at the University of Illinois, supported by grant No. 7-64 of the U.S. Air Force Office of Scientific Research, and grant GM-10718 of the U.S. National Institutes of Health.

A Precision Electronic Thermometer for Calorimetry

By R. K. Quigg*

A precision semi-automatic electronic thermometer based on an a.c. Wheatstone bridge is described. Temperature changes are followed using a high gain low noise narrow band amplifier and phase-sensitive null detector energized at 75c/s which drives an automatic counter-timer to record the instants of bridge balance. Temperature changes of 0·0001°C are accurately registered.

When used with a suitable calorimeter heat outputs of the order of 10 cal may be measured with a precision of ±0·1 per cent.

(Voir page 133 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 140)

IN the construction of a precision semi-micro calorimeter for measuring the heats of reaction of a wide range of chemical compounds it was required to measure heat outputs of the order of ten calories with a precisioin of 0·1 per cent. Since the heat capacity of the experimental system is about 120 cal.deg^{-1}, the thermometer device had to register accurately temperature changes of the order of 0·0001°C.

The method chosen was to use a temperature sensitive resistance winding in intimate contact with the aluminium calorimeter body and to connect it as one arm of a modified Wheatstone bridge.

The bridge was energized from a 75c/s oscillator and any unbalance detected with a low noise narrow band amplifier feeding a phase-sensitive detector. This has a number of advantages over the conventional d.c. bridge system: null detection by a sensitive galvanometer susceptible to mechanical vibration and zero drift is avoided, thermal e.m.f. effects are eliminated and a faster response time is achieved. In addition the use of an a.c. operated bridge facilitates the incorporation of automatic recording devices, as will be seen.

The frequency of 75c/s was chosen as a compromise to avoid interference from the 50c/s mains and its harmonics and to minimize interference from inductive and capacitive pick-up on the one hand and on the other to avoid core saturation in the transformers.

The copper resistance thermometer element, wound on the calorimeter surface in a machined thread has a resistance of about 95·7Ω and a temperature coefficient of 0·355Ω deg^{-1} at 25°C. Its resistance is measured using a three lead compensating method in the nearly symmetrical bridge shown in Fig. 1. The elements of the bridge have the values indicated and the measuring arm consists of a set of Pye precision decades in 0·01Ω to 1kΩ steps. All the fixed resistors are mounted in an aluminium block, temperature controlled to ±0·01°C by a transistorized temperature controller[1]. Under these conditions the resistance values do not change during an experiment by more than one part in 10^7. Self-heating of the resistors of the bridge, which is detectable with a bridge supply voltage of much in excess of 20V is minimized by using 10V or less.

When the bridge is energized at its normal voltage of 10V temperature change in the calorimeter of $2·7 \times 10^{-4}$°C causes a change of 10^{-4}Ω in the thermometer winding, this produces a bridge out of balance voltage of 0·1μV.

The bridge out of balance voltage is fed via an impedance matching transformer to the pre-amplifier, Fig. 1. The matching transformer must be of unusually good construction, astatically wound, well shielded magnetically and with the core and all turns, lead-out wires etc. locked in a synthetic resin impregnation.

Sir Thomas Grubb, Parsons, Newcastle upon Tyne make special transformers to order, but a standard Parmeko P.2568 was found to be very satisfactory. It is mounted in a cradle of foam-rubber to minimize vibration.

The pre-amplifier, which is mounted close to the bridge consists of one cascade low-noise input stage[2], V_1, V_2 of gain × 100, followed by two stages in cascade, V_3-V_6 made selective by asymmetrical twin-T networks[3]. The overall gain of the amplifier is about $2·6 \times 10^5$ and the bandwidth is 5c/s, symmetrical about the centre frequency of 75c/s. The conventional twin-T rejection circuit with symmetrical elements has an asymmetrical frequency response. It is in the present application advantageous to have a symmetrical, steeper sloped response. This is achieved by suitably proportioning the passive elements[3] and by arranging that each twin-T network is fed from a low-impedance source and feeds into a high impedance. To minimize hum interference from the 50c/s mains all the amplifier valve heaters are fed by 12·6V d.c. from a full-wave bridge rectifier.

To achieve high stability and low noise, metal-oxide film resistors, Mylar foil capacitors and anti-vibration p.t.f.e. valveholders were used in the construction of the pre-amplifier.

The narrow bandwidth of the pre-amplifier is necessary to reduce the effects of random noise and a.c. pick-up and prevent them from overloading the output amplifier and p.s.d. and to virtually eliminate harmonics of the 75c/s signal arising from the oscillator itself or the two bridge transformers and prevent them from reaching the p.s.d.

The pre-amplifier, which has a switched attenuator, feeds the phase-sensitive detector via V_7 a low-power driver stage. N_1 and N_2 are to limit voltage peaks when the bridge is far off balance.

The oscillator is a conventional Wien bridge circuit which delivers either 10, 3 or 1V at 75c/s to the measuring bridge via a transformer with very low inter-winding capacitance. To ensure a highly stable oscillator frequency, necessary because of the narrow (5c/s) bandwidth of the pre-amplifier, only metal-oxide resistors and polystyrene film capacitors were used in the oscillator circuits. In practice no trouble has arisen due to frequency drift in the oscillator or in the tuned circuits in the amplifier.

The phase-sensitive detector is of novel design in that equal fixed resistors are inserted in series with each diode for biasing purposes. It has been described in detail by

The Queen's University of Belfast.

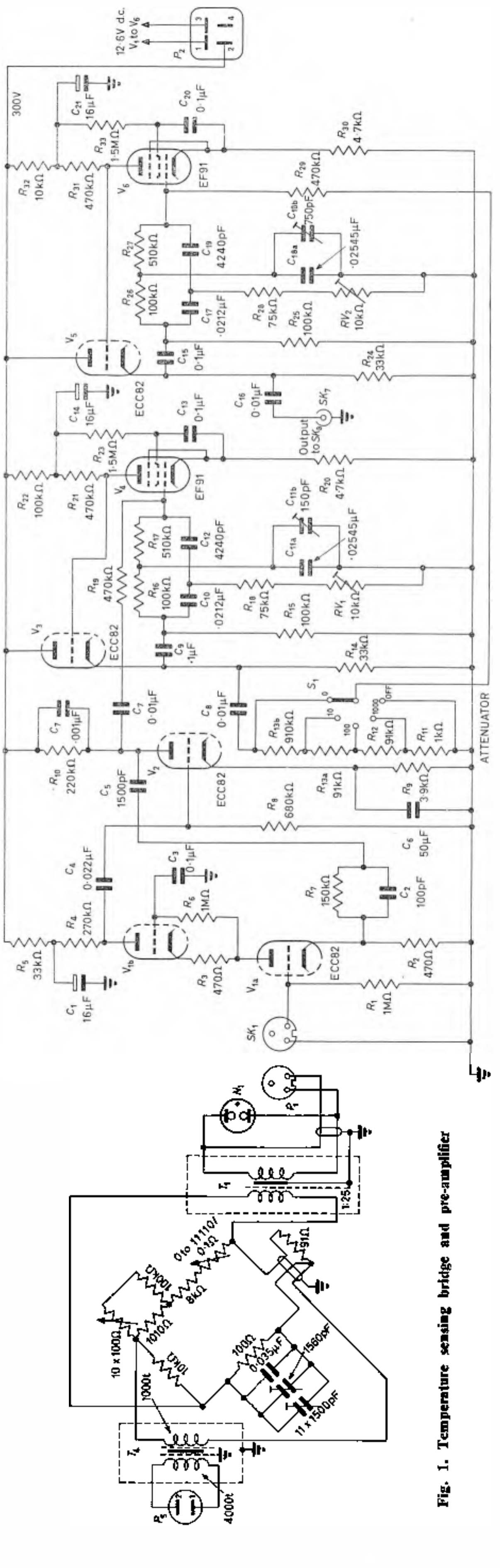

Fig. 1. Temperature sensing bridge and pre-amplifier

Tucker[4]. The reference phase of the p.s.d. comes from the oscillator via a phase-shifting circuit which is adjusted so that the p.s.d. responds only to changes in the resistive balance of the temperature-sensing bridge. This of course means that any quadrature component of the bridge output is ignored by the p.s.d. It is nevertheless necessary to maintain by occasional adjustments the capacitive balance of the bridge as otherwise the out-of-balance reactive component would overload the output stage of the pre-amplifier.

The d.c. output of the p.s.d. is displayed on a 5-0-5V centre-zero meter and is also fed to a potentiometric recorder to monitor base-line drift, bridge stability and temperature drift, in the fore-period of the calorimetric determination.

When the bridge is energized by the normal 10V, a change of $10^{-4}\Omega$ in the resistance of the thermometer (which corresponds to a change of $0{\cdot}1\mu$V in the bridge output and to a temperature change of $2{\cdot}7 \times 10^{-4}$°C) produces a p.s.d. output of 0·75V d.c. and a deflexion of 1in on the recorder. The r.m.s. variation in the p.s.d. output corresponds to less than $0{\cdot}005\mu$V (0·05in on the recorder). This is also close to the theoretical Johnson noise for a bridge of this design (c. 200Ω and a bandwidth of ~5c/s. With this arrangement a resolution about five times superior to that of the best d.c. galvanometer null detector may be achieved.

During a normal experiment the capacitive balance of the bridge drifts by 200 to 300pF and the balance must be restored from time to time to avoid overloading the amplifier. This is done by displaying on the y-axis of a miniature c.r.t. the amplifier output and on the x-axis the reference signal. A horizontal line indicates that the bridge has both resistive and reactive balance. Resistive unbalance only is indicated by a tilted line and capacitive unbalance by a horizontal ellipse. It is thus easy to adjust for capacitive balance even though the resistive component is not completely balanced. The pre-set phase shifting stage is adjusted in the following manner, after the bridge is stable and balanced: The bridge capacitance is varied a small amount either side of balance and the coarse and fine two gang phase shift controls $RV_{103a,b}$ and $RV_{104a,b}$ Fig. 2 are adjusted so that a minimum change occurs in the p.s.d. output. This also results in maximum sensitivity to resistive changes in the bridge balance. When all circuits have settled down, only occasional minor changes to this phase-angle are needed.

The power supplies of +300V d.c., −300V d.c. and 12·6V d.c. are quite conventional and are shown in Fig. 3.

It remains now to describe the method of recording the times when the bridge goes through balance, since the operational procedure follows the usual calorimetric one of timing the temperature change. This is done by pre-setting the measuring arm of the bridge at appropriate successive values and recording the instant of balance. This latter operation is done automatically in the following way. A print-out counter (English Numbering Machines Ltd.) counts down the 50c/s mains to generate a train of pulses spaced 0·1sec. The time at which a pair of contacts is closed is printed out on a 2in wide paper tape. When the bridge passes through balance (temperature of calorimeter changing during an experiment) the output of the p.s.d. changes from a negative value through zero to a positive value, this voltage drives a voltage comparator V_{12} with a two coil centre stable Carpenter relay A in the two anodes. A balance control is adjusted so that when the p.s.d. output is zero contact is just made. This is determined by the neon lamp arrangement N_3, N_4 of Fig. 4.

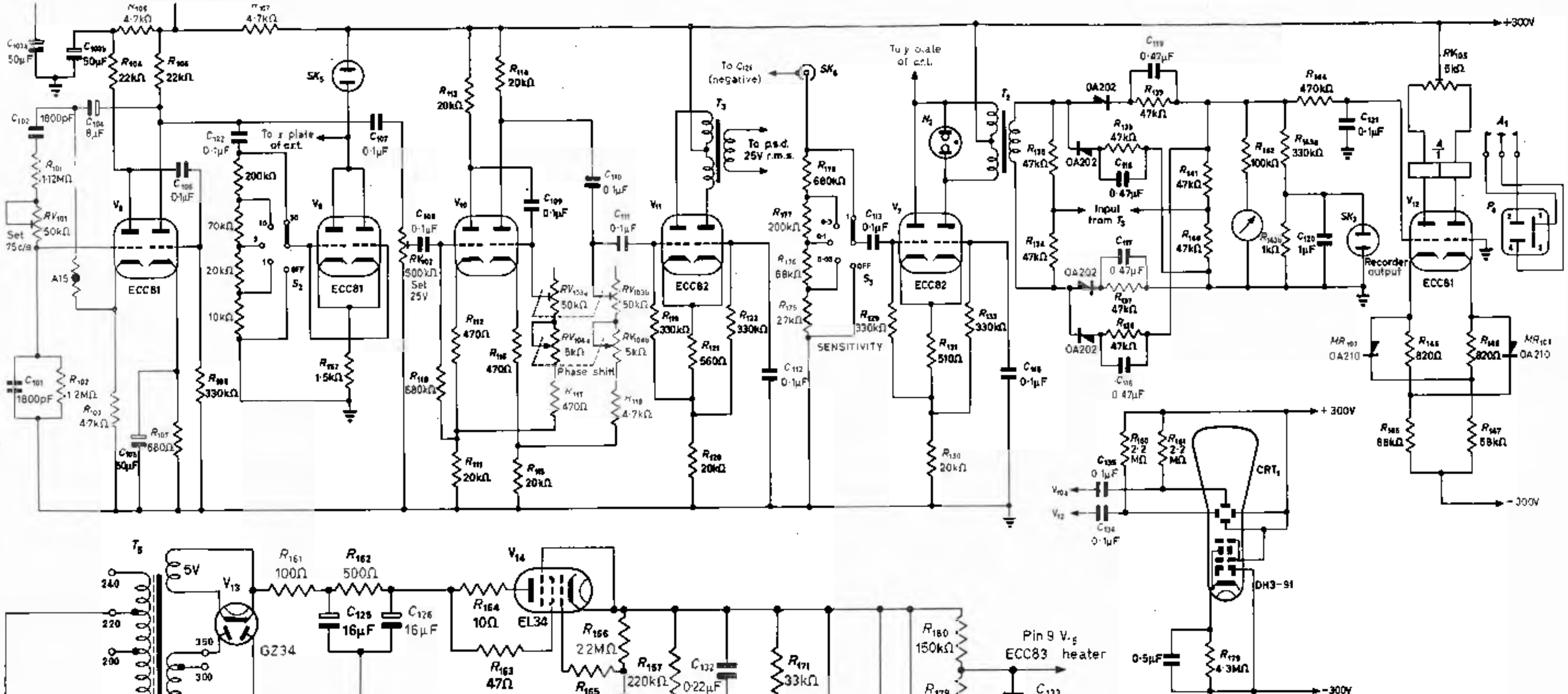

Fig. 2 (above). 75c/s oscillator, p.s.d. and circuit null-indicator

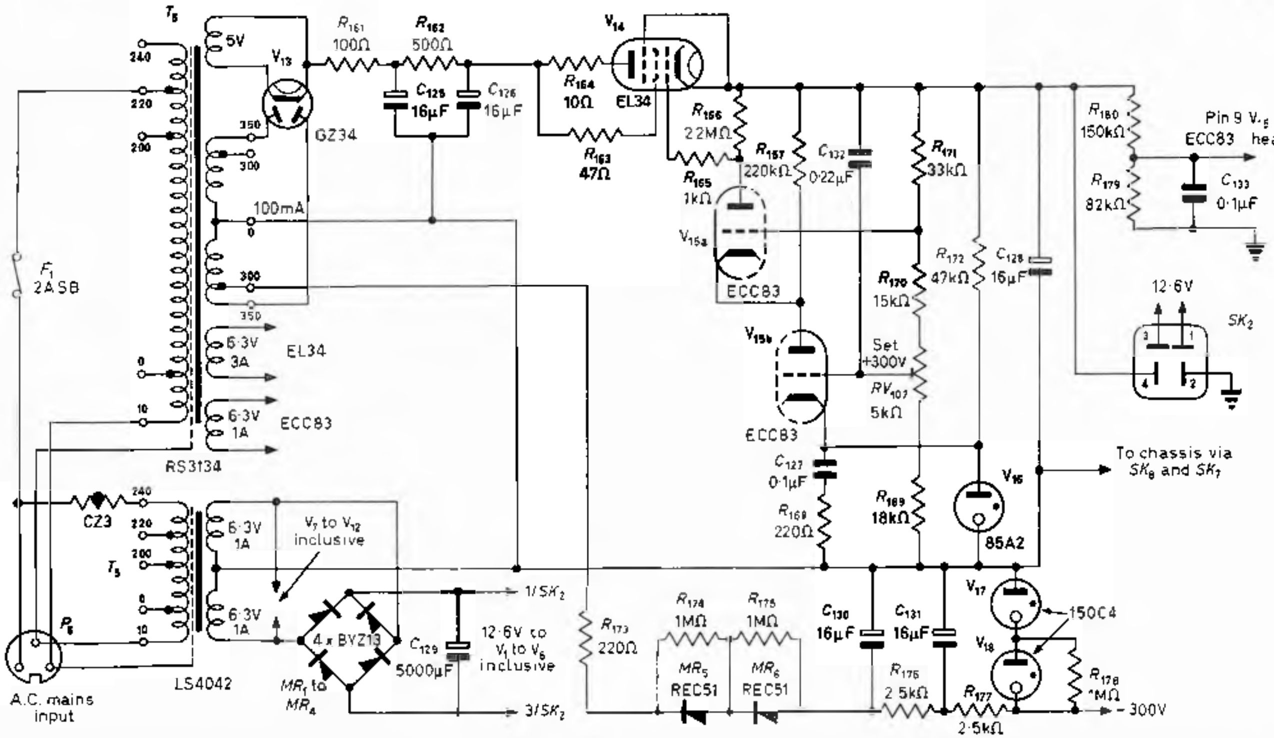

Fig. 3 (left). Power supply

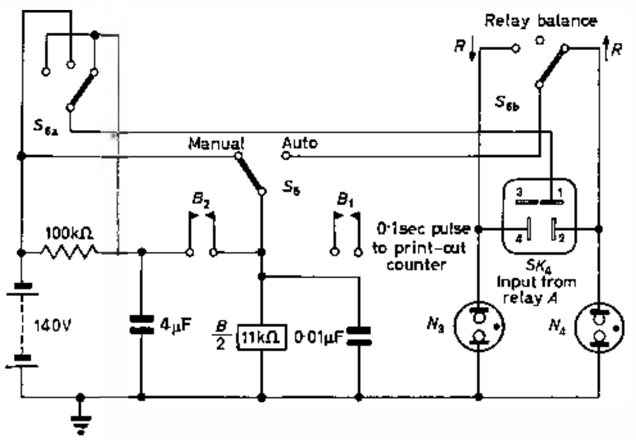

Fig. 4. Relay balance and drive to print-out counter

The contacts of the Carpenter relay are too light to energize the solenoid of the print-out counter. In addition the solenoid is required to be energized only for 0·1sec without chatter. To do this light-duty relay *B* is used which is energized by discharging a 4μF capacitor through its coil via the contacts of relay *A*. To counteract the effects of noise when going through the balance point very slowly, which would cause the contacts of relay *A* to chatter and not fire relay *B* positively a second pair o contacts on relay *B* is connected in parallel with th relay *A* contacts. This ensures a positive hold-in until th 4μF capacitor is discharged.

Conclusion

The electronic thermometer just described has been use successfully over the past 18 months in conjunction witl an aneroid semi-micro reaction calorimeter. No troubl has been experienced in maintaining stability or sensitivity The accuracy of calibration has been checked at frequen intervals (by introducing a known quantity of heat int the calorimeter) and has remained within the limit claimed.

Acknowledgments

Acknowledgment of helpful discussion of the chemistr problems involved is due to Dr. H. Mackle, Senio Lecturer and Dr. Margaret A. Frisch, Research Assistan both of the Department of Chemistry, Q.U.B. Professo C. Kemball is thanked for his permission to publish thi article.

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Power Supplies for Flight Simulators

As the speed and operating efficiency of the digital computer is constantly improved it has tended to replace the analogue computer even for applications where the latter has been most widely used. Flight simulators have been one of the largest applications of the analogue computer but the digital computer is being used increasingly for new simulators under construction.

The analogue computer has been considered specially suited to simulation problems because of its continuous response, and the ease of converting the electrical signal into mechanical movement. This last feature is of special value when building fully comprehensive simulators which provide cabin movement, vibration, and feel in the controls. These mechanical outputs from the analogue computer are initiated by two-phase servo motors. One phase provides excitation for the motor and comes direct from the power supply, and the other supply, which is 90° out of phase to the first, is the carrier used for the computer and ultimately provides the output control signal for the motors.

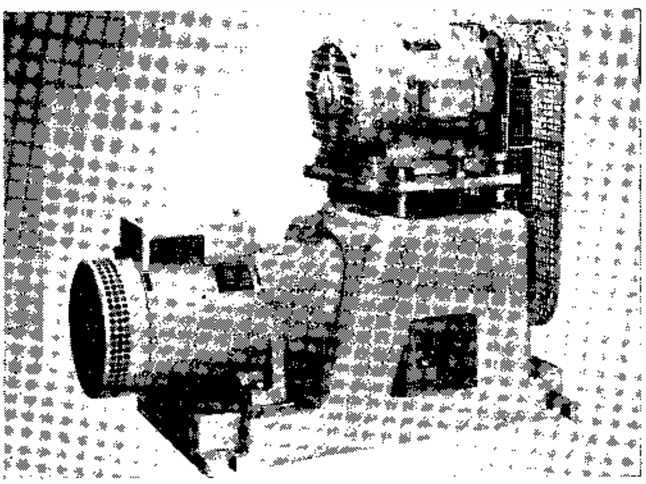
Where a sine wave carrier of this type is used for an analogue computer the frequency chosen is usually 400c/s, as this is the optimum frequency for minimum hysteresis losses in the magnetic materials used for transformers and motors in the system. It is important that the power supplies for use with an analogue computer provide a very accurate sinusoidal waveform if the amplification of the various units within the computer are to remain constant, and the final signal is to remain undistorted. A range of alternators suitable for use with flight simulators is manufactured by W. Mackie & Co. Ltd.

These motor alternators provide a basic power supply of 115V ±3%. There are strict limitations on the waveform, the requirement being for a low harmonic distortion content and in addition, no abrupt change in waveform exceeding 1 per cent is allowed. The alternator sets are required to operate over an ambient temperature range of −10°F to +120°F, and are frequently specified to run from both 50c/s and 60c/s power supplies. Depending on the size of simulator to be powered, Mackie supply a range of alternators up to 10kVA capacity.

The introduction of the real time digital computer has overcome the major difficulty of using this type of computer for simulation. Digital computers permit easy changes to the program and 'aircraft' performance, and the ability to store data offers advantages in the simulation of non-linear aero dynamic and engine characteristics. Program control of the digital computer also permits more than one cockpit to be driven from a computer complex. While the main compute complex for new simulators under construction is digital requiring d.c. power supplies, the nature of some inputs and outputs is such that analogue techniques are required fo convenient transfer of the signal to or from the digital section. 400c/s a.c. power supplies are required for many of these analogue sections and they are also required for most of the aircraft instruments that form part of the simulator.

In service, simulators must maintain a very high standard of reliability as airlines usually hire out their simulators to smaller operators using the same aircraft. This means that the equipment may be in use for as much as 18 hours a day The total reliability of the power supplies is an important factor in achieving this degree of utilization. The alternators are of the brushless type and the only servicing normally required is greasing the bearings every 5 000 hours.

A 4kVA, 115V, 400c/s motor alternator supplied for use with a 'Redifon' simulator



Second Thoughts on the Modified Z-Transform

By Y. Azar*, Dip.Eng., A.M.I.E.R.E.

The existing tables of the modified z-transform proved unsatisfactory as they give erroneous results which are always delayed by one sampling period T, and do not give the true waveform. The new method removes the artificial delay from the forward path and places it in the fictitious output sampler. It is shown that this improvement gives the expected waveforms. A formula for converting the old tables to the new ones is developed and four examples are given for illustration. The new method combined with the modified initial value theorem gives a powerful tool to the engineer as it enables him to find a single value at any time without evolving the whole series.

(Voir page 133 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 140)

THE purpose of the modified z-transform is to supply information about the system behaviour in the periods between sampling of the driving input. The unmodified transform gives the value of the output at the sampling instants only, while the modified transform enables one to obtain the values in the period between two samplings. One point must be stated clearly at this stage: when referring to 'sampling', the sampling of the driving input is meant, and it should not be confused with the theoretical imaginary sampler that is placed at the output, in order to satisfy mathematical requirements.

Fig. 1 describes how to obtain the output voltage at the sampling instants. $r(t)$ is the driving input. $G(z)$ is the z-transfer function of the system. $C(z)$ is the z-transform of the output voltage, and $C(nT)$ is the actual voltage at the sampling instants, $0, T, 2T \ldots nT$. Note that the imaginary sampler was introduced merely for convenience, while in reality the output voltage is taken directly from G.

The factor h/T was introduced to take into account the finite width of the samples[1]. The output voltage is known only at the sampling instants, although usually it exists also in the periods:

$$nT < nT + \Delta T < (n+1)T$$

where Δ is a positive number smaller than 1 ($0 < \Delta < 1$).

The following example of a simple R-C low pass filter demonstrates what has been discussed so far.

Example of a single-lag

$$\left[\ G(s) = \frac{1}{1 + sRC}\ \right]$$

Consider the circuit shown in Fig. 2(a):

Due to the intrinsic error of the z-transform[1,2] the calculated output is larger than the value measured, or calculated by the Laplace transform method. The initial step is $V_b h/RC$ instead of zero, and the final value is:

$$V_b\,.\,(h/T)\,\frac{T/RC}{1 - e^{-T/RC}} \quad \text{instead of } V_b$$

It is possible to change the definition of the z-transform such that the initial jump is eliminated, by shifting (i.e. delaying) the calculated waveform by one sampling period T. This may satisfy engineering taste better, but will not improve the magnitudes of the error, which tends to offset the enthusiasm to re-write all the existing z-transform tables.

Description of the New Modified z-transform

To obtain the output voltage in between the sampling instants, Fig. 1 is slightly modified to include a small imaginary delay ΔT, between the closing action of the input sampler and that of the imaginary output sampler (Fig. 3).

$C(z, \Delta)$ denotes that the output $C(z)$ is sampled ΔT

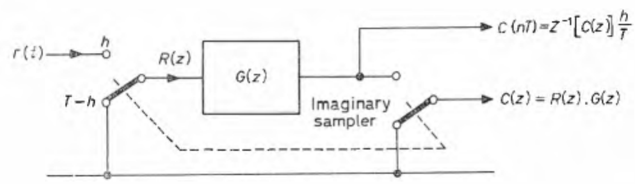


Fig. 1. Sampled system using the concept of the modified transform with synchronous switching

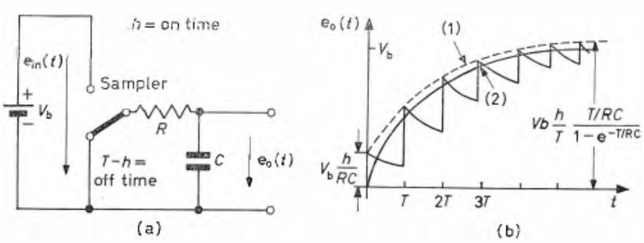


Fig. 2. Illustration of waveform calculation

Curve (1) z-transform method (approximation)
(2) Laplace transform method (exact)
Note that only the maximum values are plotted in (2)

seconds after the input $r(t)$ is sampled. $G(z, \Delta)$ is the 'modified z-transfer function' and is defined as:

$$G(z, \Delta) = \frac{C(z, \Delta)}{R(z)} \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots \ (1)$$

A sampled function in the time domain is defined as:

$$f^*(nT) = \sum_{n=0}^{\infty} f(t)\delta(t - nT) \quad \ldots\ldots\ldots\ (2)$$

where $\delta(t)$ is the impulse function.

The Laplace transform of equation (2) is:

$$F^*(s) = \sum_{n=0}^{\infty} f(nT)\,e^{-sTn} \quad \ldots\ldots\ldots\ldots\ (3)$$

From equation (2) it follows that a function sampled with a delay of ΔT is given by:

$$f^*(nT + \Delta T) = \sum_{n=0}^{\infty} f(t)\delta\,[t - (nT + \Delta T)] \quad \ldots.\ (4)$$

where $0 < \Delta < 1$

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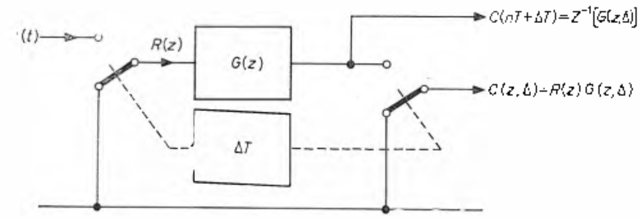


Fig. 3. Sampled system using delayed output sampling

Taking the Laplace transform of equation (4):

$$F^*(s,\Delta) = \quad [f^*(nT+\Delta T)]$$
$$= \sum_{n=0}^{\infty} f(nT+\Delta T)\,.\,e^{-sT(n+\Delta)} \quad \ldots\ldots\ldots \quad (5)$$

Substituting $z = e^{sT}$ in equations (3) and (5):

$$F(z) = \sum_{n=0}^{\infty} f(nT)\ .\ z^{-n} \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (6)$$

which is the definition of the unmodified transform, and

$$F(z,\Delta) = \sum_{n=0}^{\infty} f(nT+\Delta T)\, z^{-(n+\Delta)} \quad \ldots\ldots\ldots\ldots \quad (7)$$

which is the definition of the new modified transform.

To demonstrate the difference between equations (6) and (7) the two definitions in the case of the decaying exponential (i.e. $f(t) = e^{-at}$) will be applied.

If $f(t) = e^{-at}$, then from equation (3):

$$F^*(s) = \sum_{n=0}^{\infty} e^{-anT}\,.\,e^{-sTn}$$
$$= (1+e^{-at}\,.\,e^{-sT}+e^{-2aT}\,.\,e^{-2sT}+\ldots)\ \frac{1-e^{-aT}\,.\,e^{-sT}}{1-e^{-aT}\,.\,e^{-sT}}$$
$$= (1-e^{-aT}.e^{-sT}+e^{-aT}.e^{-sT}-\ldots+\ldots)\ \frac{1}{1-e^{-aT}\,.\,e^{-sT}}$$
$$= \frac{e^{sT}}{e^{sT}-e^{-aT}} = \frac{z}{z-e^{-aT}} \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots \quad (8)$$

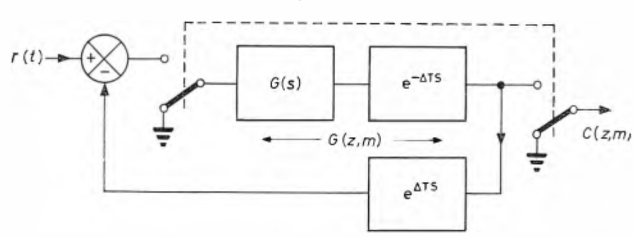
The inverse z-transform of equation (8) yields the values of $f^*(nT)$ at times 0, T, $2T$ ... etc., which are 1, e^{-aT}, e^{-2aT} etc. No information is known about values in between these periods.

To find these values one must apply equation (5), and obtain:

$$F^*(s,\Delta) = \sum_{n=0}^{\infty} e^{-aT(n+\Delta)}\,.\,e^{-sT(n+\Delta)}$$
$$= (0+e^{-\Delta Ta}\,.\,e^{-s\Delta T} + e^{-aT(1+\Delta)}\,.\,e^{-sT(1+\Delta)}+\ldots)$$
$$= e^{-\Delta T(s+a)}\,(1+e^{-T(a+s)}+e^{2T(a+s)}+\ldots)\ \frac{1-e^{-T(a+s)}}{1-e^{-T(a+s)}}$$
$$= \frac{e^{-a\Delta T}\,.\,e^{-\Delta Ts}}{1-e^{-aT}\,.\,e^{-sT}} = \frac{z}{z-e^{-aT}}\ (e^{-a\Delta T}\,.\,z^{\Delta} \ \ldots\ldots \ (9)$$

The first term of equation (9) equals to equation (8). The second term indicates that the output is delayed by ΔT

Fig. 4. Conventional way of describing the modified z-transform of feedback systems



and that the first sample is $e^{-a\Delta T}$ instead of 1 as in equation (8). Note that equation (9) is equal to equation (8) for $\Delta T = 0$.

Comparison with the Old Modified z-Transform

Description of the Old Modified z-Transform

It is unfortunate that the topic of the modified z-transform has been so obscured and confused in the literature[3,4,5,6]. That, and the intrinsic error which is due to the rectangular approximation[2] is probably the reason why so many engineers in sampled data systems are reluctant to use the z-transforms.

Firstly, the tables of the existing modified transforms yield waveforms which are always delayed by T, irrespective of the value of Δ.

Secondly, no mention is made that some theorems like the initial value theorem must be modified as all the numerators of the listed functions are one power of z smaller than the unmodified functions.

Thirdly, because of incorrect physical explanations many authors resort to a fictitious time advance in the feedback path to compensate for the artificial delay in the forward path (Fig. 4). This 'explanation' does not help to increase

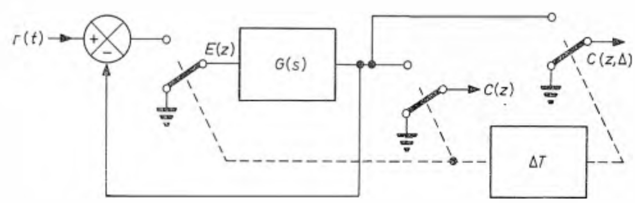


Fig. 5. New configuration for describing the modified z-transforms of feedback systems

the level of confidence among engineers. The new configuration (Fig. 5) is much more logical and gives the correct results for systems with and without feedback.

From Fig. 4 it is seen that:

$$C(z,m) = R(z)\frac{G(z,m)}{1+G(z)} \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (10)$$

The new configuration leaves the system intact but places the delay on the output sampler switching action (Fig. 5).

The modified output $C(z,\Delta)$ is found as follows:

From Fig. 5, $C(z,\Delta) = E(z)\,.\,G(z,\Delta)$, (which is also applicable to Fig. 3). Thus:

$$E(z) = \frac{R(z)}{1+G(z)} \text{ and } C(z,\Delta) = \frac{R(z)\,G(z,\Delta)}{1+G(z)} \quad \ldots\ldots \quad (11)$$

Although equations (10) and (11) are in identical form no extra 'tricks' were necessary and one can safely apply the same configuration for both open (Fig. 3) and closed loop (Fig. 5).

Mathematical Conversion from the Old to the New Modified z-Transform

The basic difference between the old and the new theory lies in definition of the modified transform. The old theory defines the modified transform as:

$$F^*(s,m) = \sum_{n=0}^{\infty} f[nT+(m-1)T]\,.\,e^{-sTn} \quad \ldots\ldots \quad (12)$$

where $m = 1-\Delta$,

while the new theory defines the modified transform as:

$$F^*(s,\Delta) = \sum_{n=0}^{\infty} f(nT+\Delta T)\, e^{-sT(n+\Delta)} \quad \ldots\ldots\ldots \quad (5)$$

TABLE 1

NO.	$f(t)$ TIME FUNCTION	$F(s)$ LAPLACE TRANSFORM	UNMODIFIED Z-TRANSFORM $F(z)$	CONVENTIONALLY MODIFIED Z-TRANSFORM $F(z, m)$ $m = 1 - \wedge$	NEW MODIFIED Z-TRANSFORM $F(z, \triangle)$ $0 < \wedge < 1$
1	$u(t)$	$\frac{1}{s}$	$\frac{z}{z-1}$	$\frac{1}{z-1}$	$\frac{z \,.\, z^{-\Delta}}{z-1}$
2	e^{-at}	$\frac{1}{s+a}$	$\frac{z}{z-e^{-aT}}$	$\frac{e^{-amT}}{z-e^{-aT}}$	$z\,\frac{z^{-\Delta} \,.\, e^{-\Delta Ta}}{z-e^{-aT}}$
3	at	$\frac{a}{s^2}$	$\frac{aTz}{(z-1)^2}$	$\frac{amT}{z-1} + \frac{aT}{(z-1)^2}$	$aTz^{-\Delta}\left[\frac{z\wedge}{z-1} + \frac{z}{(z-1)^2}\right]$
4	$\sin at$	$\frac{\omega}{s^2+\omega^2}$	$\frac{z \sin \omega T}{z^2 - 2z \cos \omega T + 1}$	$\frac{z \sin m\omega T + \sin(1-m)\omega T}{z^2 - 2z \cos \omega T + 1}$	$z \,.\, z^{-\Delta}\,\frac{z \sin \wedge\omega T + \sin(1-\wedge)\omega T}{z^2 - 2z \cos \omega T + 1}$

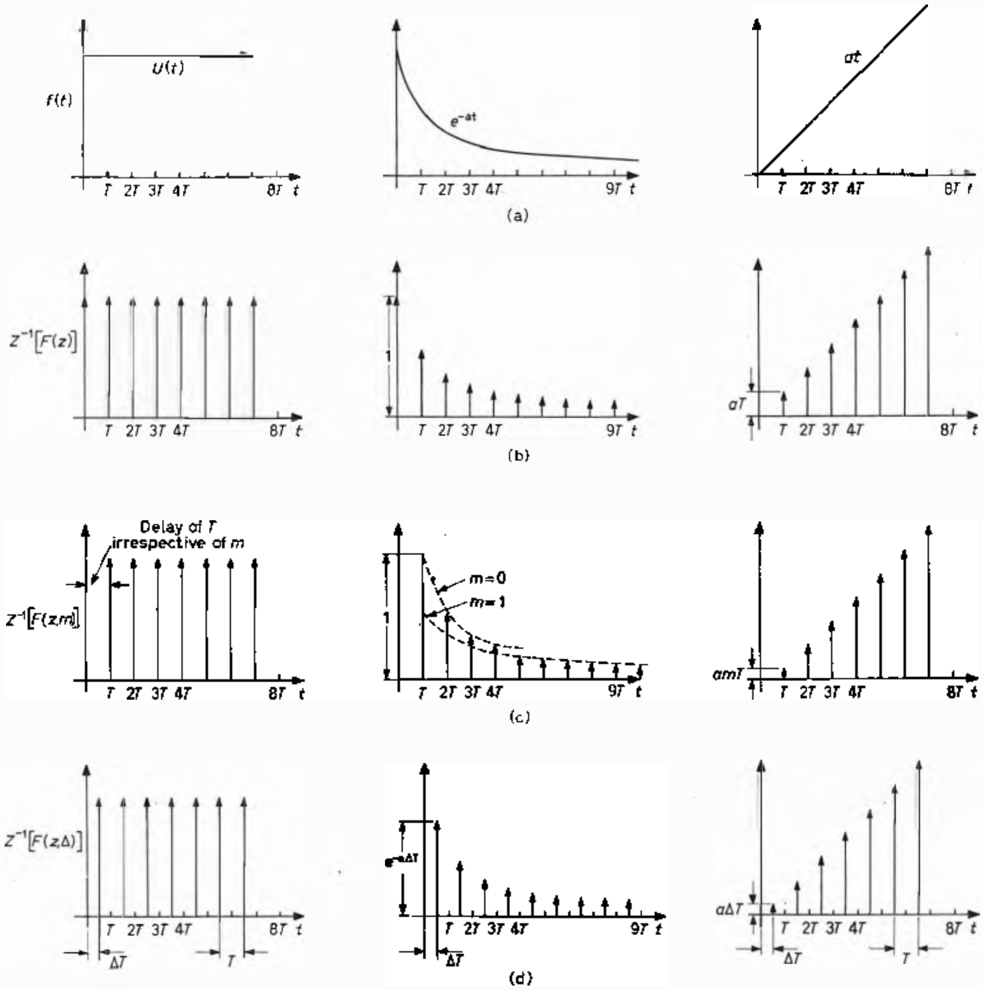


Fig. 6 (a) Original functions of t
(b) Inverse of the corresponding unmodified transforms
(c) Inverse of the corresponding conventionally modified transforms
(d) Inverse of the corresponding new modified transforms

since the delay is applied to the switching action rather than to the actual system.

Multiplying equation (12) by $e^{sT} . e^{-smT}$:

$$e^{sT} . e^{-smT} F^*(s, m) = e^{sT} \sum_{n=0}^{\infty} f[(n-1)T + mT]\, e^{-sT(n+m)} \qquad \text{.......... (13)}$$

Applying the shift theorem to equation (13) yields:

$$e^{sT} . e^{-smT} F^*(s, m) = \sum_{n=0}^{\infty} f[nT + mT]\, e^{-sT(n+m)} \qquad \text{..... (14)}$$

which is in the same form as equation (5) if mT is treated as ΔT.

In other words:

$$z \,.\, z^{-m} \,.\, F(z, m) = F(z, \triangle) \qquad \text{........... (15)}$$

where mT is treated as the delay ΔT.

With the aid of equation (15) the old tables can be modified to fit the new definition. Before proceeding further to compare the three methods, it will be shown how to derive two more modified pairs.

Derivation of the New Modified z-Transform for Ramp and Step Functions

(a) *Ramp function* $f(t) = at$

Substituting $aT = f(t)$ in equation (5) gives:

$$F^*(s, \Delta) = a\Delta T\, e^{-s\Delta T} + aT(1+\Delta)\, e^{-sT(1+\Delta)} + \ldots.$$
$$= aT\, e^{-s\Delta T}\left[\Delta + (1+\Delta)\, e^{-sT} + (2+\Delta)\, e^{-2sT} + \ldots\right]$$
$$= aT\, e^{-s\Delta T}\left[\Delta(1 + e^{-sT} + e^{-2sT} + \ldots) + e^{-sT} + 2e^{-2sT} + \ldots\right]$$
$$= aT\, e^{-s\Delta T}\left[\frac{\Delta}{1 - e^{-sT}} + \frac{e^{-sT}}{(1 - e^{-sT})^2}\right]$$

or:

$$F(z, \Delta) = aTz^{-\Delta}\left[\frac{\Delta z}{z-1} + \frac{z}{(z-1)^2}\right] \ldots\ldots\ldots\ldots (16)$$

(b) *Step function*: $f(t) = u(t)$

In this case:

$$F^*(s, \Delta) = e^{-s\Delta T} + e^{-sT(1+\Delta)} + e^{-sT(2+\Delta)} + \ldots$$

which converges to:

$$F(z, \Delta) = \frac{z \,.\, z^{-\Delta}}{z-1} \quad \ldots\ldots\ldots\ldots\ldots\ldots (17)$$

Comparison Between the Three Methods

In Table 1 are listed three time functions and their corresponding continuous Laplace transforms, unmodified, conventionally modified and the new modified z-transforms.

Fig. 6 shows the plot of the original functions and their corresponding function which were obtained by first transforming them from the time domain to the z-domain and then back to the time domain. It will be noticed that the unmodified method gives the correct waveshape but only the values at multiples of T. The conventionally modified transform gives the correct values but shifts the waveform by T and thus results in a wrong waveshape.

The new modified theory gives the correct waveshape and enables one to pick the information at any desired time.

z-Transform Theorems

Initial Value Theorem

The initial value theorem applies in this case to the first output sample, and is derived on the same line as its corresponding unmodified theorem.

It has been shown that:

$$\lim_{t\to 0} f(t) = \lim_{z\to\infty} F(z) = \lim_{n\to 0} f(nT) \quad \ldots\ldots (18)$$

Similarly, it can be shown by either using first principles or by the use of the displacement theorem (sometimes also known as the shift theorem) that:

$$\lim_{n\to 0} f[nT + (\Delta + k)T] = \lim_{z\to\infty} z^k \,.\, z^\Delta \,.\, F\,[z, (k + \Delta)] \quad \ldots\ldots\ldots\ldots (19)$$

where n and k are integers.

k was introduced here to take into account displaced functions as well.

An interesting point emerges here. As the first term in the time function is $f[T(k + \Delta)]$ and its image is $f^*[T(k + \Delta)]\, z^{\Delta+k}$ one may call $\Delta_i = \Delta + k$ and allow it to vary from 0 to ∞. This enables one to find any single value at any time t_1 of the function $f(t_1)$, by using the initial value theorem and substituting $\Delta_i = t_1/T$ in $F(z, \Delta_i)$.

Thus:

$$\lim_{t_1\to t} f(t) = \lim_{z\to\infty} z^{t_1/T}\,[F(z, t_1/T)] \quad \ldots\ldots\ldots\ldots (20)$$

Equation (20) gives one a very powerful tool in finding a single value at a particular time t_1 without having to develop the whole function from 0 to t_1, and as explained before, t_1 must not be an integer multiple of T.

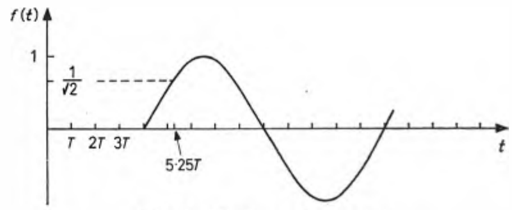


Fig. 7. Displaced sine wave

To illustrate this point the values will be calculated of a sampled displaced sine-wave (Fig. 7) at $t_1 = 5\tfrac{1}{4}T$. T was chosen such that $10f = 1/T$ where f is the frequency of the discussed sine-wave.

To avoid the tedious evaluation of the modified transform of $f(t) = \sin \omega t$ existing tables will be used (Table 1, row 4) and write with the aid of equation (15).

$$F(z, \Delta) = z.\, z^{-\Delta}\, \frac{z \sin \omega\Delta T + \sin \omega\,(1-\Delta)T}{z^2 - 2z \cos \omega T + 1} \quad \ldots\ldots\ldots (21)$$

Applying the modified initial theorem (equation (20) to equation (21):

$$\lim_{t\to t_1} f(t) = \lim_{z\to\infty} z^{-(\Delta+k)} \,.\, z^{(\Delta+k)}\, \frac{z^2 \sin \Delta\omega T + z \sin \omega\,(1-\Delta)T}{z^2 - 2z \cos \omega T + 1}$$
$$= \sin \Delta\omega T \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots (22)$$

In this example $k=4$, $\omega T = 2\pi/10$ and $T\Delta = t_1 - k = 5/4T$. Thus, $\sin \Delta\omega T = \sin(5/4\ 2\pi/10) = \sin(\pi/4) = 1/\sqrt{2}$.

Final Value Theorem

For the same reasons that were applied to equation (18) one can modify the unmodified final value theorem:

$$\lim_{t\to\infty} f(t) = \lim_{z\to 1} \frac{z-1}{z} F(z) \ldots\ldots\ldots (23)$$

to read:

$$\lim_{n\to\infty} [f(nT + \Delta T)] = \lim_{z\to 1}\left[\frac{z-1}{z} \,.\, z^\Delta F(z, \Delta)\right] \ldots\ldots (24)$$

Conclusion

The new formula for the modified z-transform has been defined, and three examples given to illustrate the difference between the new and the conventional modified transformation, It has been shown that the old method results in erroneous waveform and obscures the physical understanding of z-transforms.

The new theory also proves to give a very powerful tool, by which one can determine single values at any time t_1 without having to derive the whole series of $Z^{-1}[F(z, \Delta)]$ from 0 to t_1

The reader should be warned that this new method does not increase the overall accuracy of the z-transforms. This still depends on the sampling rate[1,2]. The modification enables one to find the values of the system's output at any time and to derive the correct waveform, which the old method of the modified z-transforms did not.

Acknowledgment

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A Low Frequency Random Signal Generator

By P. N. Nikiforuk*, R. Pronovost* and G. Squires*

A low frequency random signal generator is described. Its output is a stationary voltage having a Gaussian amplitude probability density function, a mean value of zero, and a power spectral density that is constant from zero to 50c/s. A demodulation technique is used to generate the signal. A circuit diagram and typical output characteristics are included.

(Voir page 133 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 140)

IT is often necessary to evaluate the response of an electronic circuit or an automatic control system to a low frequency random signal which is a stationary function of time and has:

(1) A Gaussian amplitude probability density function
(2) A mean value of zero.
(3) A power spectral density that is constant from zero to about 50c/s.

As is well known, random signal generators incorporating the first two requirements are easily realized and are readily available. Those which also incorporate the third requirement are more difficult to realize and are not so readily available. It is for this reason that the potential user of such equipment must frequently resort to the design and development of a suitable generator.

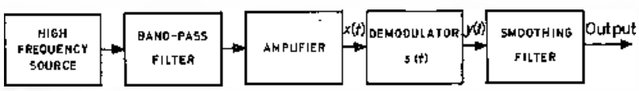

Fig. 1. Arrangement of low frequency random signal generator

Techniques for the generation of low frequency random signals are described in a number of publications[1,2,3,4,5]. However, details of actual circuit diagrams and results obtained therefrom are seldom given. In addition, some of the generators which are described tend to be complex, or unduly sensitive to component limitations, ageing and environmental changes. As a contribution to this area of work, this article describes a simple low frequency random signal generator which is constructed of readily available components. This generator has been proven to be stable and capable of producing the desired output over long periods of time. Graphs of the measured amplitude probability density function and the power spectral density are included.

Generation Technique

A demodulation technique[3] can be used for the generation of a low frequency random signal. This technique, which is illustrated in Fig. 1, requires that a high frequency stationary random signal with a Gaussian amplitude probability density function be applied to a band-pass filter. This filter has a flat response and a bandwidth which is sufficiently broad to permit the desired signal to be realized at the output of the generator. The output of this filter is amplified and applied to a demodulator which operates exactly at the centre frequency of the band-pass filter. The operating waveform of the demodulator is a square wave, whose amplitude is large compared to that of the random signal. Since demodulation, like modulation, is a multiplication process[6], the demodulator output is the product of the random signal input and the demodulator square wave signal. This results in an output that is comprised of sum and difference frequencies. One component of this output is centred about zero frequency and has a shape which is identical, except for a constant factor, to the signal at the input to the demodulator. Higher frequency components of the spectrum are also present, but these are removed by the low-pass filter which permits only the desired low frequency spectrum to appear at the generator output.

That the system of Fig. 1 produces the desired signal can be readily verified. Noting that demodulation is a multiplication process it is seen that:

$$y(t) = x(t)\ s(t) \quad \ldots\ldots\ldots\ldots\ldots \quad (1)$$

where $y(t)$ is the signal at the output of the modulator, $x(t)$ is the random signal at the input to the demodulator, and $s(t)$ is the demodulator square wave signal. The autocorrelation function of $y(t)$ is:

$$\phi_{yy}(\tau) = 1/T \int_{-T/2}^{T/2} x(t)\, s(t)\, x(t+\tau)\, s(t+\tau)\, dt \quad \ldots\ldots \quad (2)$$

but because $\phi_{yy}(\tau)$ describes an average process, this equation can be written as:

$$\phi_{yy}(\tau) = \overline{x(t)\, s(t)\, x(t+\tau)\, s(t+\tau)} \quad \ldots\ldots\ldots \quad (3)$$

where the bar indicates averaging with respect to time.

In this system $x(t)$ is independent of $s(t)$ and because of this $x(t)\, x(t+\tau)$ and $s(t)\, s(t+\tau)$ are also independent of each other. Hence, since the average of the product of two independent variables is equal to the product of their individual averages, equation (3) becomes:

$$\phi_{yy}(\tau) = \overline{x(t)\, x(t+\tau)}\ .\ \overline{s(t)\, s(t+\tau)} \quad \ldots\ldots \quad (4)$$

or:

$$\phi_{yy}(\tau) = \phi_{xx}(\tau)\, \phi_{ss}(\tau) \quad \ldots\ldots\ldots\ldots\ldots \quad (5)$$

The latter shows that the autocorrelation function of the demodulator output is the product of the autocorrelation functions of the demodulator input signal and the demodulator waveform.

To obtain the power spectral density of $y(t)$ it is necessary to obtain the Fourier transform of $\phi_{yy}(\tau)$. This is readily accomplished by noting that for a balanced square wave demodulator waveform of unit amplitude $\phi_{ss}(\tau)$ can be expressed in the Fourier series:

$$\phi_{ss}(\tau) = 8/\pi^2 \left[\cos \omega_s\tau + \frac{\cos 3\omega_s\tau}{3^2} + \frac{\cos 5\omega_s\tau}{5^2} + \ldots\ldots \right] \quad \ldots\ldots\ldots\ldots \quad (6)$$

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where $2\pi/\omega_s$ is the period of the demodulator waveform. Substituting equation (6) into equation (5), and taking the Fourier transform of the result gives the following expression for the power spectrum of $y(t)$:

$$G_y(\omega) = 8/2\pi^2\left[G_x(\omega-\omega_s)+G_x(\omega+\omega_s)+\frac{G_x(\omega-3\omega_s)+G_x(\omega+3\omega_s)}{9} + \frac{G_x(\omega-5\omega_s)+G_x(\omega+5\omega_s)}{25} + \ldots \right] \quad \ldots\ldots (7)$$

where $G_x(\omega)$ is the power spectral density of $x(t)$.

Equation (7) shows that a spectral shift of the desired type is produced by the demodulator. With the latter operating at the centre frequency of the band-pass filter one component of $G_y(\omega)$ becomes centred about zero frequency

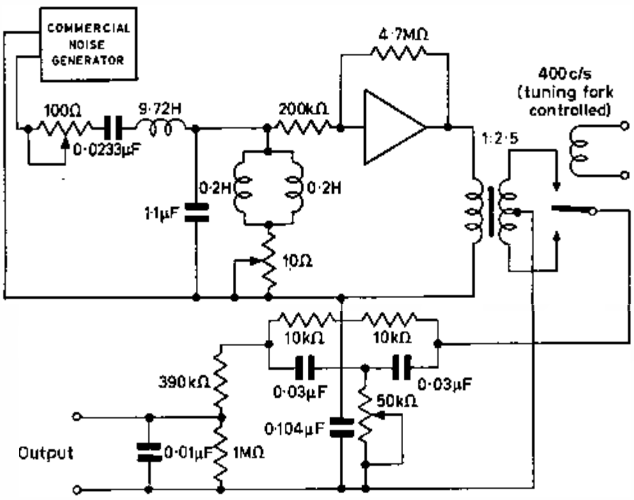

Fig. 2. Circuit of low frequency random signal generator

and has a shape which is identical to the band-pass filter characteristic, except for the factor $8/2\pi^2$. Higher frequency components are also present, but these can readily be removed by an appropriate low-pass filter. Hence, providing that the bandwidth of the filter is sufficiently broad, the desired low frequency random signal is reproduced at the output of the generator.

Generator Description and Results

A circuit diagram of the actual generator is shown in Fig. 2. A commercially available, 20 to 20 000c/s random signal generator is used as the input stage. The output of this generator has a Gaussian amplitude probability density function and a mean value of zero. A band-pass filter of the required bandwidth (approximately 100c/s) and having a flat response is difficult to realize and for this reason it is approximated by a double tuned, series parallel *RLC* network having a characteristic that peaks at approximately 340 and 460 c/s. The amplifier stage is comprised of a drift-free, operational amplifier of sufficient output current capacity to drive a 1:2·5 step-up transformer. The use of a transformer at this stage permits the removal of any residual d.c. component in the band-passed signal. The demodulator consists of a mechanical chopper whose operating frequency is maintained exactly at 400c/s by a tuning-fork controlled power supply. The low-pass filter is comprised of a twin-T rejection filter, which removes any 400c/s spikes that might appear in the output waveform, and a low-pass *RC* filter having approximately 3dB attenuation at 60c/s, which is sufficient to eliminate the peaks appearing at 60c/s due to the band-pass filter characteristic. Good quality components, shielded leads and minimum separations between component parts are used to minimize induced voltage effects. The advantage of using a mechanical detector in place of the previously used[3] diode ring demodulators is that the effects of diode mismatch, temperature fluctuations, and ageing are eliminated.

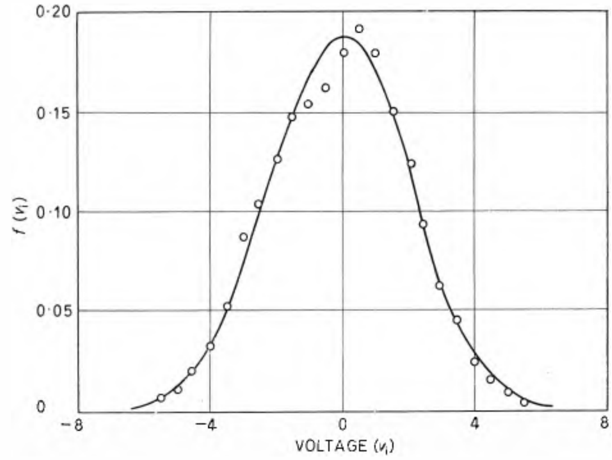

Fig. 3. Amplitude probability density function
Experimental points shown by circles
Theoretical curve shown by full line

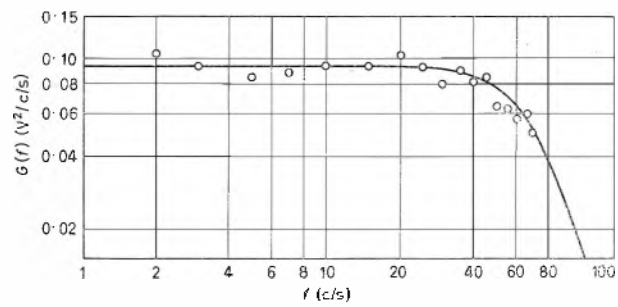

Fig. 4. Power spectral density
Experimental points shown by circles
Theoretical curve shown by full line

A considerable effort was devoted to the accurate measurement of the amplitude probability density function and the power spectral density. Typical results are shown respectively in Figs. 3 and 4. The former was obtained by means of a probability distribution analyser and the latter by means of analogue computer techniques. The application of the Chi-square test shows that the fit between the experimental points and the theoretical curves is good.

Acknowledgments

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A High Resolution Sin/Cosine Function Generator

By S. Harigovindan*

A novel function generator is described capable of generating sin/cos θ as a d.c. voltage to an accuracy better than 0·1 per cent full-scale, when θ is fed as a binary coded decimal number with a resolution of 0·1 per cent in the range of 0 to 2π. The circuit consists of high precision potential divider stages switched by relays. Two variations are described.

(Voir page 133 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 140)

THE function generator described in this article was developed for a third hybrid computer[1] to sum the series

$$F(hkl) = \sum_j f_j \sin/\cos\,(hx_g + ky_j + lz_j) \;\ldots\ldots\; (1)$$

which occurs in X-ray crystallography. The computer generates the arguments of the functions serially as binary-coded decimal numbers and the function generator is required to produce the function as a d.c. voltage to a high degree of accuracy. The function generator is simple in principle consisting of precision voltage sources and resistance potential divider (p.d.) networks, the proper taps of which are selected by multi-position switches. This class of function generator is known as digital-analogue[2], due to the fact that the argument θ is in the digital form (single or mixed base) as

$$\theta = \sum_i a_i\, b_i \;\ldots\ldots\ldots\ldots\ldots\ldots\; (2)$$

where a_i are integers (represented as switch positions) and b_i constants depending upon the bases, while the output is an analogue quantity, usually a voltage. The main advantages of this type of function-generators are their high reliability and accuracy which are determined mainly by the resistors used, and the ease with which the multiplication of any input analogue variable by the function generated can be carried out.

When the resolution in θ required is low (<2 per cent full-scale), a single stage of potential division is sufficient[3,4,5]. For higher resolution it is more convenient to use multiple stages[2,6,7] based on some relationship between $f(\theta)$ and $f(\theta)+\Delta\theta)$ such as linear interpolation[2] or Taylor's series[6]. However, in the present function generators the author has found it possible to use simple and exact generating equations, avoiding errors inherent in the approximations mentioned above. The type of switches used varies with the application, the ordinary wafer switches[3,4] being sufficient for manual operation and electronic or electromechanical switches for automatic operation. In spite of their superior speed, electronic

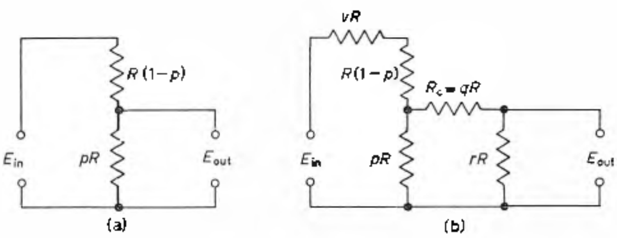
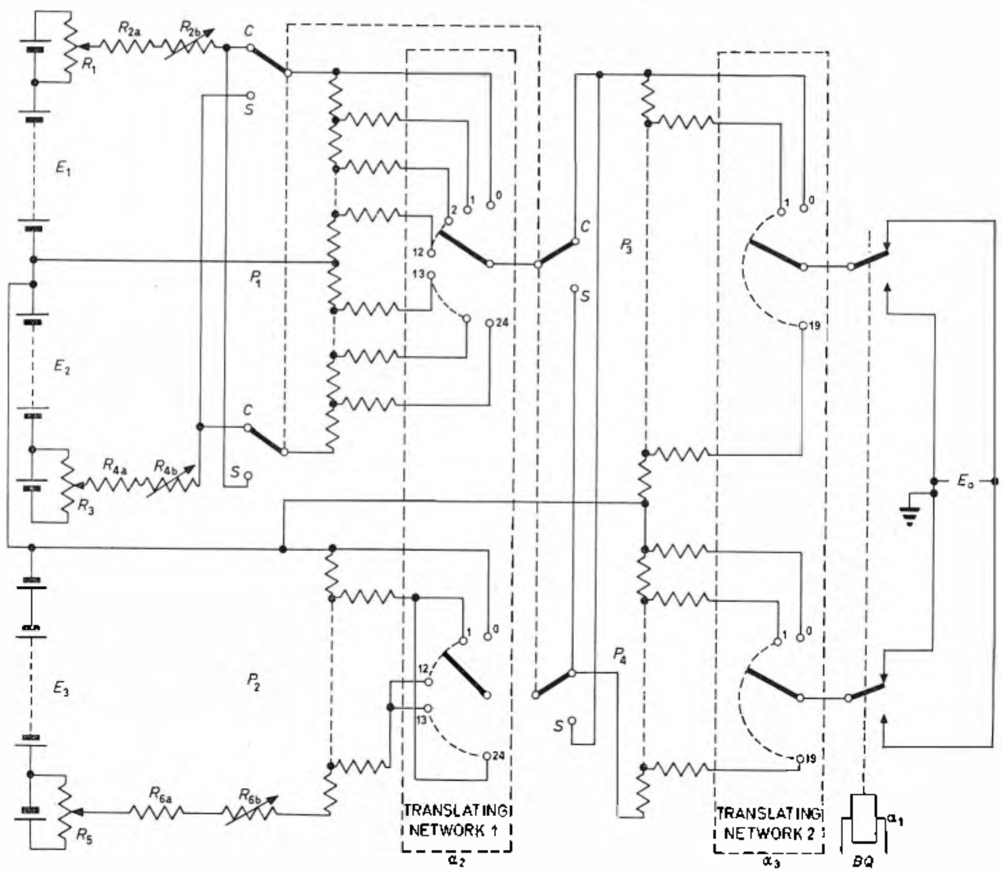


Fig. 1. (a) Simple potential divider stage
(b) Corrected stage

Fig. 2. Sine-cosine function generator



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switches such as transistors[5] are inferior to electromechanical switches for high accuracy applications due to their non-ideal switching characteristics. The uniselector (or the rotary switch) is the most commonly used electromechanical switch[6,7] and is perhaps the most economical also. The relay transfer tree network[2,6] on the other hand is faster and has the additional advantages of less complex 'argument read-in' circuits and the ability to perform translating operations. Finally special half-ladder networks are used to eliminate loading errors which become prominent in cascaded p.d. stages, thus dispensing with any active impedence matching devices such as cathode-follower in-between stages.

The function generator may be considered in detail now.

Basic Principles

The major requirements are that sin/cos θ be generated as a d.c. voltage to an accuracy of 0·1 per cent (full scale) and argument resolution of 0·1 per cent in the range 0 to 2π. Obviously, if a single stage of potential division is used this means 1 000 taps and a 1 000-way switch is required. Considerable reduction in the number of taps etc., can be effected by using multiple stages and by exploiting the symmetry of sin/cos functions, as can be done by adopting a three-digit mixed base system for θ as follows.

$$\begin{aligned} \theta &= a_1 b_1 + a_2 b_2 + a_3 b_3 \\ a_1 &= 0, 1 \qquad b_1 = \pi \\ a_2 &= 0, 1, 2 \ldots (m\text{-}1) = b_2 = \pi/m \\ a_3 &= 0, 1, 2 \ldots (n\text{-}1) = b_3 = \pi/mn \end{aligned} \qquad (3)$$

To achieve an accuracy of 0·1 per cent for θ the bases may be chosen as $m = 25$, $n = 20$ which also minimizes the number of taps. Making use of the antisymmetry relation,

$$\sin/\cos(\theta + \pi) = -\sin/\cos\theta \qquad (4)$$

the generating equations for θ given by equation (3) are seen to be

$$\cos\theta = (-1)^{a_1}\left\{\cos(a_2b_2)\cos(a_3b_3) - \sin(a_2b_2)\sin(a_3b_3)\right\} \qquad (5)$$

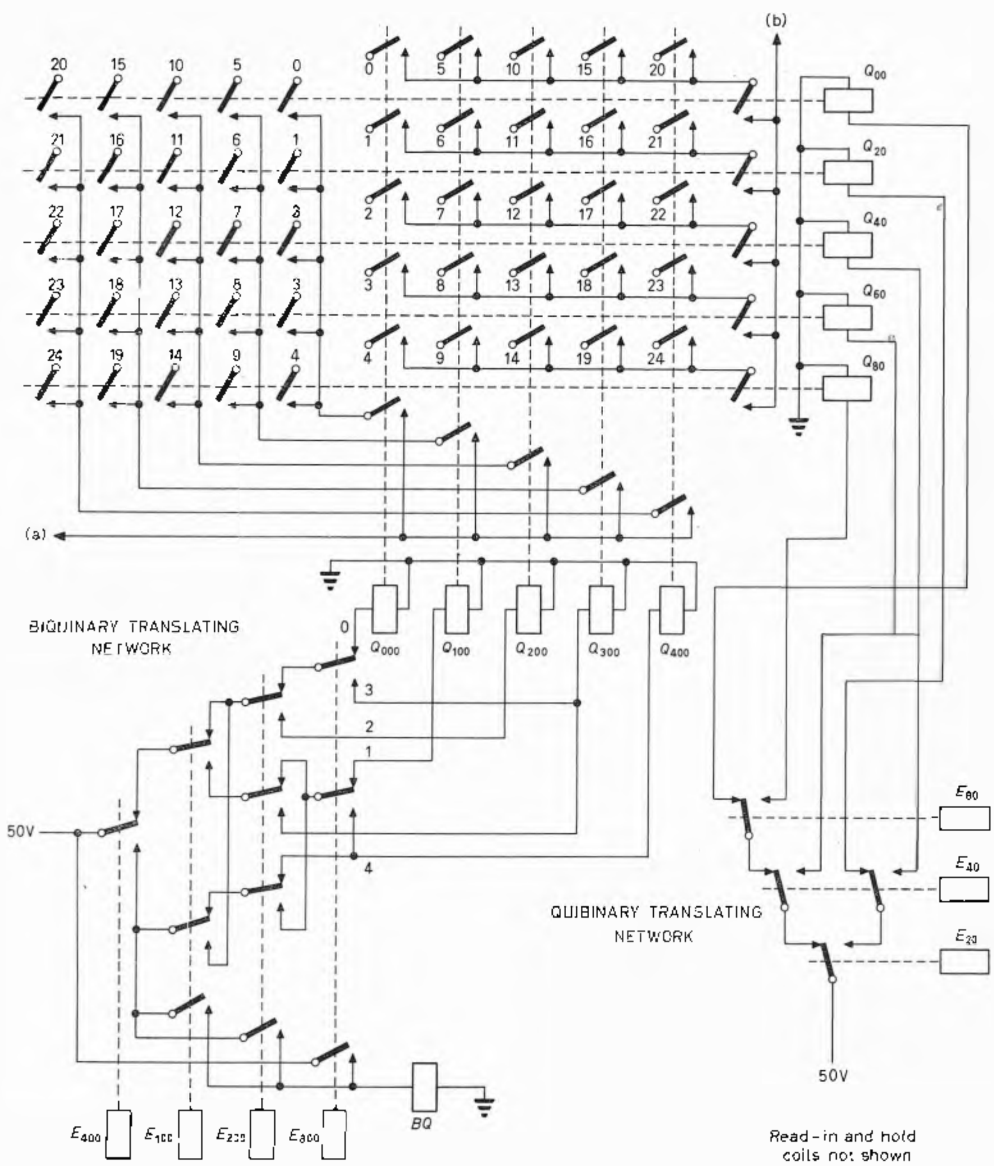
$$\sin\theta = (-1)^{a_1}\left\{\sin(a_2b_2)\cos(a_3b_3) + \cos(a_2b_2)\sin(a_3b_3)\right\} \qquad (6)$$

Since the effect of a_1 is merely to change the sign, two levels of p.d. are sufficient. The circuit used to realize equations (5 and 6) is shown in Fig. 2. It contains four p.d. stages (P_1, P_2, P_3 and P_4) each being similar to that shown in Fig. 1(b) for loading correction as discussed later. P_1 contains 25 taps with p.d. ratios proportional to cos (a_2b_2) in the range 0 to π. Its two halves corresponding to the quadrants 0 to $\pi/2$ and $\pi/2$ to π, are fed separately by two equal voltage sources of opposite signs, to provide the change of sign in the second quadrant. P_2 fed by E_3 contains 13 taps corresponding to sin (a_2b_2) in the range 0 to $\pi/2$ only since

$$\sin(a_2b_2) = \sin\left\{(25 - a_2)b_2\right\} = a_2 \leqslant 12 \qquad (7)$$

and the taps corresponding to a_2 and $(25-a_2)$ are identical. The connexions to the switch are simply repeated as shown. The taps on P_1 and P_2 are selected by the two-pole 25-way switch on which a_2 is set. Similarly, P_3 and P_4 contains 20 taps each, corresponding to cos (a_3b_3) and sin (a_3b_3) respectively in the range 0 to $19\pi/500$. The required pair of taps corresponding to a_3, are selected by a

Fig. 3. Translating network No. 1



two-pole 20-way switch. The d.p.d.t. switch operated by the relay BQ, is for changing the sign according to the value of a_1 by grounding the proper terminal which is permissible since E_1, E_2 and E_3 are floating supplies.

The four-pole two-way switch K permits the selection of sin θ or cos θ as desired. If it is assumed that the voltages across P_2 and the two halves of P_1 are equal and the loading errors are eliminated as discussed later it will be seen that E_o is proportional to cos θ if K is switched to terminals denoted C, and to sin θ if K is at S.

Loading Correction

When the function generator is loaded by a constant resistance, it can be shown by repeated application of Thevenin's theorem at each p.d. stage that the loading error can be eliminated if the output resistance of each stage remains the same at all taps. As already mentioned, each stage is similar to that shown in Fig. 1(b). The resistance $R_o = qR$ is inserted at the tap to keep the output resistance of the stage at a chosen value uR. If the p.d. ratio at the tap is p, this requires that

$$q = u - \frac{p(1 + v - p)}{(1 + v)} \quad \ldots\ldots\ldots\ldots\ldots (8)$$

where vR is the output resistance of the previous stage. Obviously it is necessary that $u \geqslant (1 + v/4)$ to ensure that q is positive. Also the values of u, v, R are the same for all p.d. stages in the same level. The formulae for p and q as well as the actual values assigned to u, v, R in a function generator constructed by the author are given in Table 1.

If the voltages at the outputs of R_1, R_3 and R_5 with P_1 and P_2 disconnected, are adjusted to be equal to E_1 and the load is rR_2, then it can be shown that

$$E_o = \frac{r E_1}{(1 + v_1)(1 + v_2)(r + 2u_2)} \sin/\cos\ (\theta)\ . \quad (9)$$

where the suffixes 1 and 2 refer to the two levels of p.d. stages (P_1, P_2) and (P_3, P_4) respectively. It is evident from equation (9) that the loading error is eliminated at the cost of constant attenuation. In the function-generator constructed by the author, E_1, E_2 and E_3 each consisted of 12 dry cells in series, so that E_o will have a full scale value of 10V. As already mentioned, the potentiometers R_1, R_3 and R_5 are for equalizing the voltages and R_2, R_4 and R_6 are for keeping the output resistance of each source at v_1R_1 at 150Ω as stated in Table 1. These operations can be carried out rapidly by means of a balancing circuit not shown in Fig. 2, for the sake of simplicity.

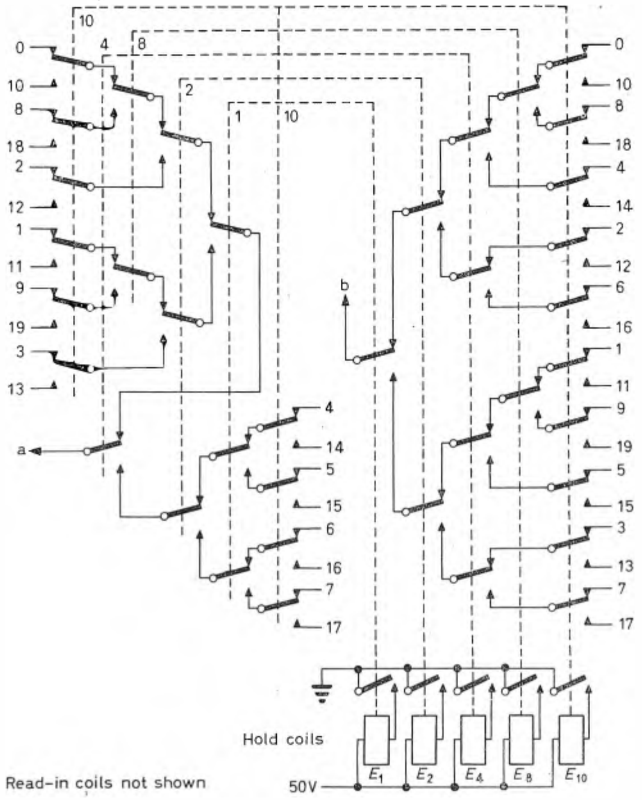


Fig. 4. Translating network No. 2

Relay Networks

The switches shown inside the dotted lines in Fig. 2, actually consist of the relay circuits shown in Figs. 3 and 4. The computer generates θ in the decimal system as

$$\left.\begin{matrix} a_1 \\ a_2 \\ a_3 \end{matrix}\right\} = 0, 1 \ldots .9 \qquad \begin{matrix} b_1 = \pi/5 \\ b_2 = \pi/50 \\ b_3 = \pi/500 \end{matrix}$$

$$\theta = a_1 b_1 + a_2 b_2 + a_3 b_3 \quad \ldots\ldots\ldots\ldots (10)$$

Hence the relays have to translate θ from the decimal system of equation (10) to the mixed base system of

TABLE 1

Values of parameters for the function generator

P.D. STAGE	R (Ω)	v	u	p	q	RANGE OF a_1
P_1	1000+ 1000	0·15	0·30	$\cos\left(\frac{a_2\pi}{25}\right)$	$0{\cdot}3 - \left\{\frac{\cos\left(\frac{a_2\pi}{25}\right)\left[1{\cdot}15 - \cos\left(\frac{a_2\pi}{25}\right)\right]}{1{\cdot}15}\right\}$	a_2 = 0 to 24
P_2	1000	0·15	0·30	$\sin\left(\frac{a_2\pi}{25}\right)$	$0{\cdot}3 - \left\{\frac{\sin\left(\frac{a_2\pi}{25}\right)\left[1{\cdot}15 - \sin\left(\frac{a_2\pi}{25}\right)\right]}{1{\cdot}15}\right\}$	a_2 = 0 to 12
P_3	2000	0·15	0·15	$\cos\left(\frac{a_3\pi}{500}\right)$	$0{\cdot}15 - \left\{\frac{\cos\left(\frac{a_3\pi}{500}\right)\left[1{\cdot}15 - \cos\left(\frac{a_3\pi}{500}\right)\right]}{1{\cdot}15}\right\}$	a_3 = 0 to 19
P_4	2000	0·15	0·15	$\sin\left(\frac{a_3\pi}{500}\right)$	$0{\cdot}15 - \left\{\frac{\sin\left(\frac{a_3\pi}{500}\right)\left[1{\cdot}15 - \sin\left(\frac{a_3\pi}{500}\right)\right]}{1{\cdot}15}\right\}$	a_3 = 0 to 19

equation (3) as follows. Since a_1, a_3, a_2 are binary coded (8421) decimal digits, four delays are required to represent each of them. Thus a_1 is set on the four relays E_{800}, E_{400}, E_{200}, E_{100} of Fig. 3, which translate the 8421 code to biquinary code, the binary part of the latter directly giving a_1 on the relay BQ (Figs. 2 and 3). Similarly, a_2 is fed to the relays E_{80}, E_{40}, E_{20} (Fig. 3) and E_{10} (Fig. 4) which carry out quibinary translation. The two quinary parts so obtained are set on the relays, Q_{000}, Q_{100}, Q_{200}, Q_{300}, Q_{400}, and Q_{00}, Q_{20}, Q_{40}, Q_{60}, Q_{80}, respectively, which form a cross-point array. The two sets of 25 terminals in this array correspond to similarly numbered terminals of the two-pole 25-way switch of Fig. 2, a and b being the two poles of the switch. The last digit a_3 is set on the relays E_8, E_4, E_2 and E_1 which together with E_{10}, form (Fig. 4) the switch for selecting a_3. Again the two sets of 20 terminals in the network of Fig. 4 correspond to the similarly numbered terminals of the two-pole two-way switch of Fig. 2. Actually E_{10} consists of four relays the coils of which are connected in parallel, since the required heavy spring load cannot be carried by a single relay. Each relay of these circuits is provided with two coils, for 'read-in' and 'hold' operations. For reading in any value of θ, proper 'read-in' coils are energized for 20msec by the electronic circuits in the computer, and the energized relays will be self-locked through their 'hold' coils (see Fig. 4). The argument can be cleared simply by breaking the common ground connexion of the self-locking paths for a short interval. In these switching circuits care has been taken to ensure minimization and uniform distribution of the spring load.

TABLE 2
Sin/cosine Function Generator Performance for Random Inputs

	Cos $(2\pi\theta)$			Sin $(2\pi\theta)$		
θ	TABLES	FUNCTION GENERATOR	FULL SCALE ERROR (%)	TABLES	FUNCTION GENERATOR	FULL SCALE ERROR (%)
0·010	0·9980	0·9976	0·04	0·0628	0·0628	0·00
0·025	0·9877	0·9866	0·11	0·1564	0·1566	0·02
0·046	0·9585	0·9577	0·08	0·2851	0·2847	0·04
0·067	0·9127	0·9126	0·01	0·4086	0·4089	−0·03
0·088	0·8510	0·8515	−0·05	0·5252	0·5256	−0·04
0·130	0·6846	0·6849	−0·03	0·7290	0·7292	−0·02
0·151	0·5826	0·5836	−0·10	0·8128	0·8129	−0·01
0·172	0·4708	0·4713	−0·05	0·8823	0·8828	−0·05
0·193	0·3505	0·3511	−0·06	0·9366	0·9367	−0·01
0·214	0·2244	0·2251	−0·07	0·9745	0·9720	0·20
0·235	0·0941	0·0951	−0·10	0·9956	0·9949	0·07
0·256	−0·0378	−0·0377	0·01	0·9993	1·0003	−0·10
Standard Deviation 0·06%				Standard Deviation 0·07%		

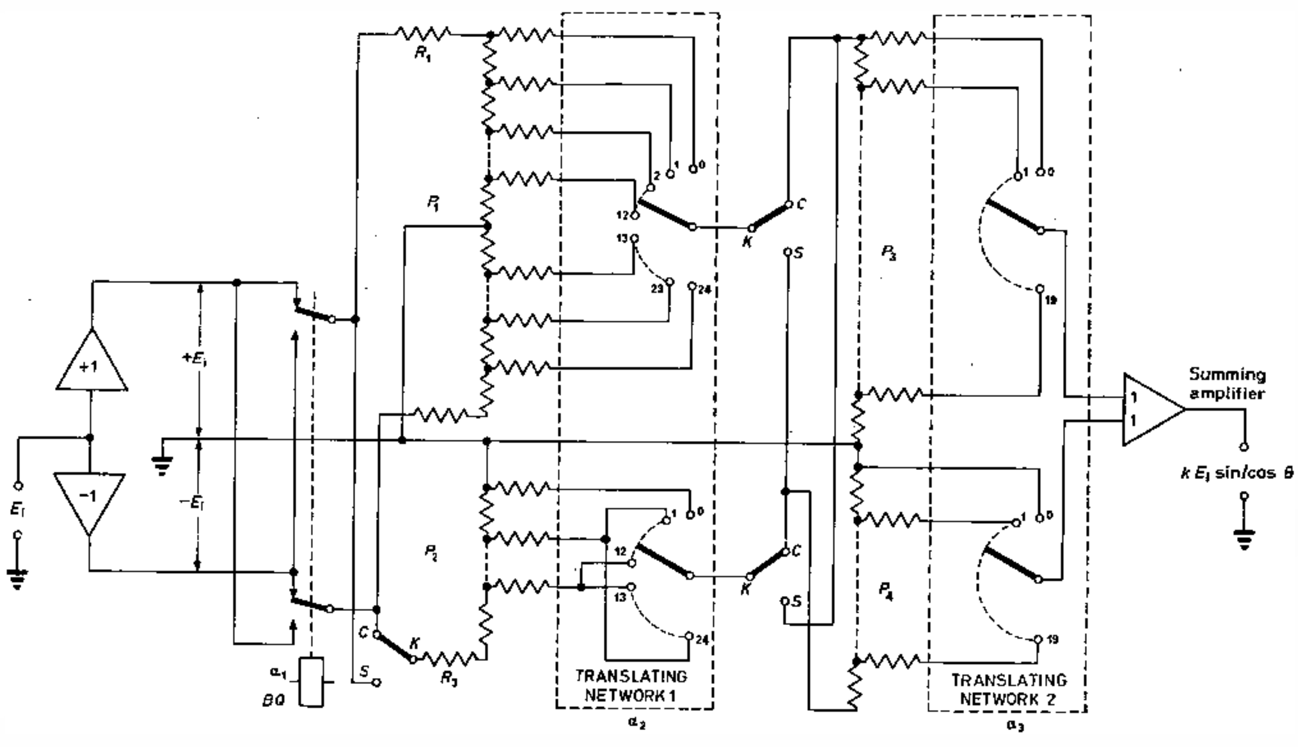
A Modified Circuit

The circuit of Fig. 2 suffers from the drawbacks of using floating supplies and the necessity of frequent adjustments of the six potentiometers (R_1 to R_6) due to the drainage of the cells. The modified circuit shown in Fig. 5 avoids these troubles by using d.c. amplifiers. It has also the advantage that any varying input voltage E_i can be multiplied by sin/cos θ. R_1, R_2 and R_3 of Fig. 5 are fixed precision resistors of value $v_1 R_1$ so that no adjustments are required. In all other respects the circuits of Figs. 2 and 5 are similar.

Error Analysis

The major sources of error are seen to be the errors in the values of the resistors and voltages. If it is assumed that all these quantities are within $\pm d$ (d is a positive

Fig. 5. Modified sin/cos function generator



fraction) of their correct values, then it can be shown that the theoretically maximum possible error ΔE_o in E_o under the worst conditions for a given value of θ can be—for cos θ generation

$$\max\left\{\frac{\Delta E_o}{E_{o(max)}}\right\} = d\left\{A\cos(a_2b_2 - a_3b_3) + B\cos(a_2b_2 + a_3b_3)\right\} \quad \ldots\ldots\ldots\ldots (11)$$

for sin θ generation

$$\max\left\{\frac{\Delta E_o}{E_{o(max)}}\right\} = d\,(A + B)\sin(a_2b_2 + a_3b_3) \ldots\ldots (12)$$

where

$$A = 3 + \frac{v_1}{1 + v_1} + \frac{v_2}{1 + v_2}$$

$$B = \frac{2\,u_2}{r + 2\,u_2}$$

$$E_{o(max)} = \frac{r\,E_1}{(1 + v_1)\,(1 + v_2)\,(r + 2\,u_2)}$$

Thus the full-scale error can have a theoretical maximum of $d(A + B)$ at $\theta = 0$ for cos θ and $\theta = \pi/2$ for sin θ. However, this is a highly pessimistic estimate and in practice, the average error will be less than one-third of the above. It may be pointed out that temperature variation should not be a major source of error if all the resistors have the same temperature coefficient as will be the case when all the resistors are wound from the same type of wire.

Conclusion

A practical circuit of the type shown in Fig. 2 has been successfully constructed by the author. The resistors were wound from 38 s.w.g. manganin wire and their values were adjusted to a tolerance of 0·05 per cent. The relays are the standard 3000 type. The performance of the function generator is highly satisfactory as is evident from Table 2, where E_o was measured by means of a Tinsley Vernier Potentiometer. The average error is seen to be about 0·05 per cent only. It may be noted that the same circuit configurations as those shown in Figs. 2 and 5 can be used for generating hyperbolic sine or cosine functions also. If a pure binary system were used for θ, the circuits could be simplified more.

Acknowledgment

The author's acknowledgments are due to Prof. R. S. Krishnan for his kind interest and the Council of Scientific & Industrial Research, India, for financial support.

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A Keyboard Morse Coder

The 'Keymaster' manufactured by 'fast-electronic', West Germany, is a device which produces faultless Morse code messages from a typewriter-like keyboard. It is intended to ease the task of a telegraphist or to enable persons not skilled in Morse to send telegraphic messages, as may be necessary in cases of emergency etc. By lightly depressing a key the Keymaster will transmit a complete Morse character with accuracy and uniformity at an adjustable speed between 4 and 40 words/min.

The main component of the 'Keymaster' is a semiconductor counting chain. In its initial position the last flip-flop of this chain is "flipped over", thus keeping a clock (multivibrator) blocked via a stop circuit. As soon as one of the keys is depressed for a short period, a counting stage is switched and the clock commences to operate: with each pulse (= a Morse dot) the counting chain jumps one step forward. The character is completed when the last flip-flop has been reached.

To obtain a dash the counting chain prevents the de-energizing of the keying relay between two dots via an appropriate tap, thus achieving a Morse dash of exactly three times the dot duration independent of the operating speed.

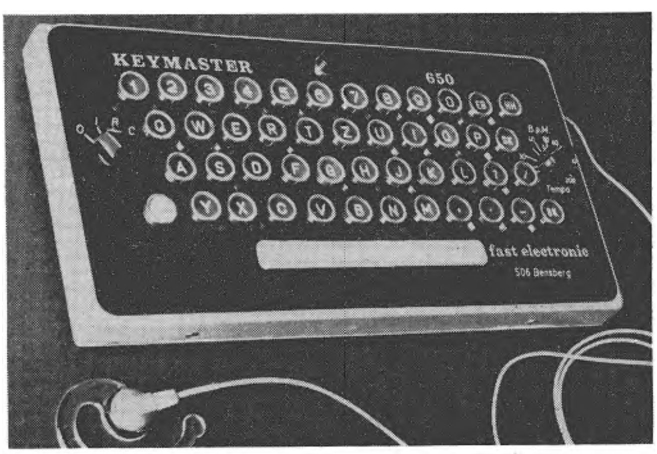
A special magnetic keyboard has been designed for the input of characters; the shortest possible operational movement ensures positive contact. The 'Keymaster' 650 contains silicon transistors and diodes, ceramic capacitors and resistors with an operating temperature range from −50°C to +100°C (−58° to +212°F). The telegraphic speed is between 20 and 200 characters/min and continuously adjustable.

The 'built-in' pre-programmed spacing between characters prevents 'writing together' of two characters. They follow each other with standard spacing although the key may have been depressed prematurely. Depressing of the spacing key extends the standard spacing in accordance with standard procedure. This ensures that no Morse message is garbled during dispatch.

Special versions can be supplied for Telex as well as for coded, automatic or computer input and the like. The overall dimensions are 5·5in by 11in by 0·8in and the weight under 4 lb. The operating voltage is 6·5V ±2V d.c., the consumption approximately 25mA.

The output is via dry reed relays rated at 12VA (up to 48V d.c. or 110V a.c.), an audio signal at 1kc/s is also provided for operation of 2kΩ headphones.

The Keymaster Morse coder



The Logic and Input Equations of Flip-Flops

N. N. Biswas*, Ph.D.

After a brief review of the logical properties of the five basic flip-flop circuits, the article analyses the map method of deriving the input equations of the flip-flops. Various rules are given in a tabular form which may prove useful and instructive to those engaged in the logical design of sequential circuits.

(Voir page 133 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 140)

In the design of a sequential circuit, flip-flops are invariably used as binary memory elements. The combinations of states of these individual memory or storage elements represent various internal states of the sequential circuit. There are five basic flip-flop circuits which can be

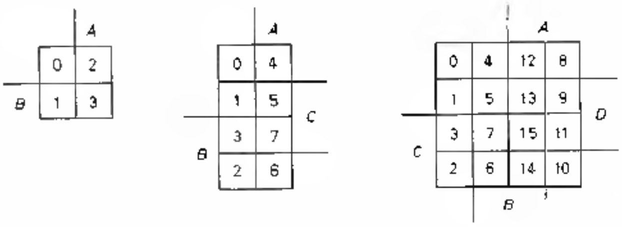
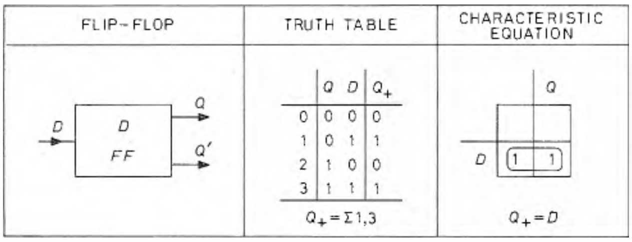


Fig. 1. Decimal numbering of cells on the Veitch-Karnaugh map



	Q	D	Q_+
0	0	0	0
1	0	1	1
2	1	0	0
3	1	1	1

$Q_+ = \Sigma 1,3$

Fig. 2. D flip-flop

used for this purpose. These are the delay (*D*), the trigger (*T*), the *S-R*, the *J-K*, and the *S-R-T* flip-flop. After the designer selects the particular type of flip-flop to be used as a memory element, it becomes necessary to derive the input equations to this flip-flop from the application equation which the flip-flop must satisfy for the given sequential circuit. This article analyses the map method[1,2] of deriving these input equations. Input equations can be derived by the truth table[1,3] and the analytic method of solving simultaneous Boolean equations[1]. Both these methods are, however, lengthy and laborious. On the other hand, the method outlined here may prove much simpler and quicker. As it is convenient to work with a four variable Veitch diagram or Karnaugh map, the analysis has been carried out for a case requiring four flip-flops, that is, for a sequential circuit whose number of internal states s is such that

$$2^3 < s \leqslant 2^4$$

It will be evident later that the analysis is valid for sequential circuits with any number of states; the functions are to be plotted on an n variable map where n is the number of flip-flops required and is the smallest integer satisfying the relation

$$n \geqslant \log_2 s$$

In order that the Veitch diagram[4], and the Karnaugh map[5] coincide, the various minterms will be plotted[6] as shown in Fig. 1. In this arrangement, the decimal numbering of cells[7] will be as shown. Since the method is based on the characteristic equations of the flip-flops, it is worth while to review the properties of the flip-flops, and determine their characteristic equations.

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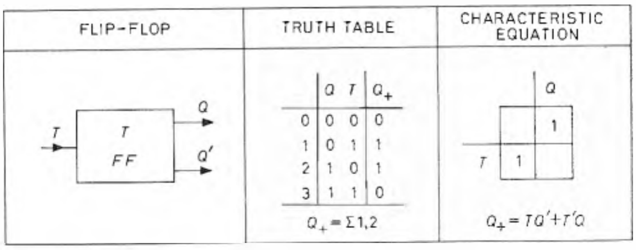
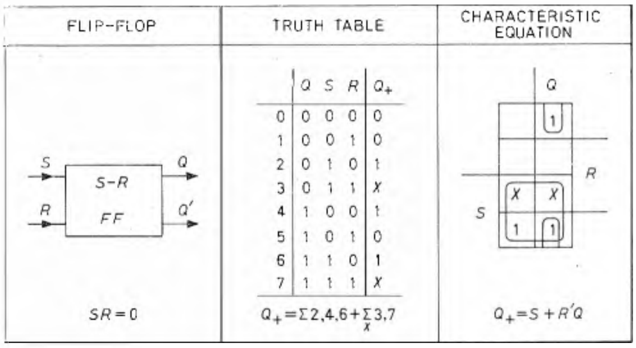


	Q	T	Q_+
0	0	0	0
1	0	1	1
2	1	0	1
3	1	1	0

$Q_+ = \Sigma 1,2$

Fig. 3. T flip-flop



	Q	S	R	Q_+
0	0	0	0	0
1	0	0	1	0
2	0	1	0	1
3	0	1	1	X
4	1	0	0	1
5	1	0	1	0
6	1	1	0	1
7	1	1	1	X

$Q_+ = \Sigma 2,4,6 + \Sigma_X 3,7$

Fig. 4. S-R flip-flop

Characteristic Equations of the Flip-Flops

Delay (*D*) flip-flop

The *D* flip-flop has only one input line and its output equals the input, one bit time earlier. Hence its truth table will be as shown in Fig. 2. In all the truth tables, the state of an *FF* or an input line at bit time *t*, is denoted by the letter as such, and the state at bit time $(t + 1)$ is denoted by putting a + sign after the letter. After plotting the truth table on the Veitch diagram and simplifying, the characteristic equation of the *D* flip-flop turns out to be

$$Q_+ = D$$

Trigger (*T*) flip-flop

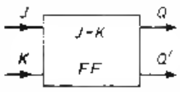
The *T* flip-flop also has only one input line. When a 1 arrives at the input, the state of the *FF* changes. For a 0 input, on the other hand, the state remains unchanged. The truth table and the derivation of the characteristic equation is shown in Fig. 3. The characteristic equation for this *FF* is

$$Q_+ = T'Q + TQ'$$

Set and Reset (*S-R*) flip-flop

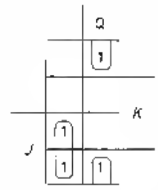
The *S-R* flip-flop has two input lines, the *S* line and the *R* line. When the *S* line is 1, the *FF* is set to 1, irrespective of

FLIP-FLOP	TRUTH TABLE	CHARACTERISTIC EQUATION



	Q	J	K	Q_+
0	0	0	0	0
1	0	0	1	0
2	0	1	0	1
3	0	1	1	1
4	1	0	0	1
5	1	0	1	0
6	1	1	0	1
7	1	1	1	0

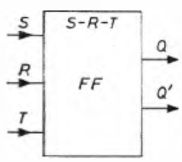
$Q_+ = \Sigma\, 2,3,4,6$



$Q_+ = JQ' + K'Q$

Fig. 5. J-K flip-flop

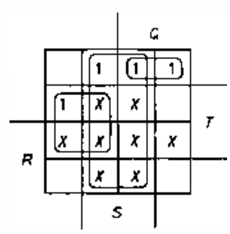
FLIP-FLOP	TRUTH TABLE	CHARACTERISTIC EQUATION



$SR = 0$
$RT = 0$
$ST = 0$

	Q	S	R	T	Q_+
0	0	0	0	0	0
1	0	0	0	1	1
2	0	0	1	0	0
3	0	0	1	1	X
4	0	1	0	0	1
5	0	1	0	1	X
6	0	1	1	0	X
7	0	1	1	1	X
8	1	0	0	0	1
9	1	0	0	1	0
10	1	0	1	0	0
11	1	0	1	1	X
12	1	1	0	0	1
13	1	1	0	1	X
14	1	1	1	0	X
15	1	1	1	1	X

$Q_+ = \Sigma\, 1,4,8,12 + \Sigma_X\, 3,5,6,7,11,13,14,15$



$Q_+ = S + TQ' + R'T'Q$

Fig. 6. S-R-T flip-flop

its previous state. When the R line is 1, the FF is set to 0, irrespective of its previous state. The state of the FF remains unchanged when both S and R are 0. Moreover, the S and R lines are not allowed to be 1 simultaneously, that is, $SR = 0$. The truth table, its plotting and simplification are shown in Fig. 4. The characteristic equation is found to be

$$Q_+ = S + R'Q$$

J-K FLIP-FLOP

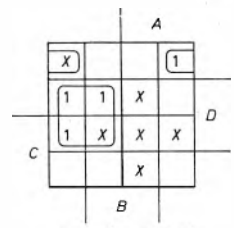
The *J-K* flip-flop has also two input lines. The J and K corresponds to the S and R input respectively of the *S-R* flip-flop, that is, a 1 on J sets the FF to 1, and a 1 on K resets it to 0, irrespective of the previous state. However, here the J and K lines are allowed to become 1 simultaneously, and under this condition the state of the FF changes. The truth table, its plotting and simplificataion is shown in Fig. 5. The characteristic equation is found to be

$$Q_+ = JQ' + K'Q$$

S-R-T FLIP-FLOP

The *S-R-T* flip-flop has three input lines, S, R, and T. When S is 1, the FF is set to 1. When R is 1, the FF is reset to 0, and when T is 1, the FF changes state. No two of the S, R, and T lines are allowed to be 1 simultaneously, which means

$$SR = RT = TS = 0$$



$D_A = A'D + B'C'D'$

Fig. 8. Input equation of D flip-flop

The truth table, and the derivation of the characteristic equation are shown in Fig. 6; the latter is

$$Q_+ = S + TQ' + R'T'Q$$

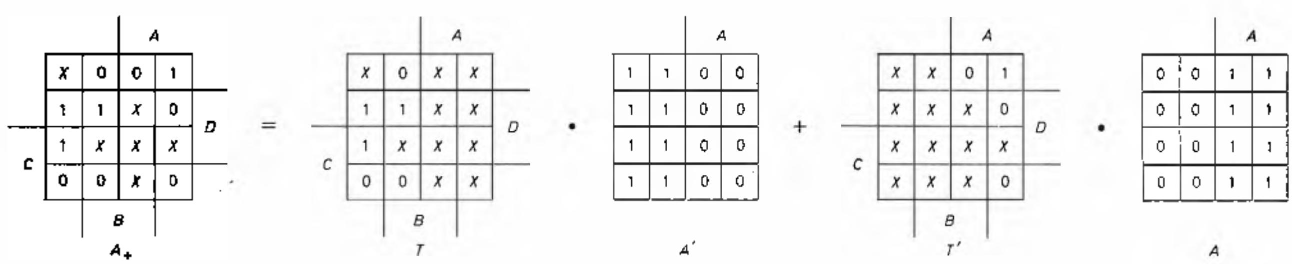
Application Equations

To illustrate the method, take the case of the decade counter solved by Lowell Amdahl[2]. Four FF's A, B, C and D are required for the ten count-states. The coding of the states and the flow table are shown in Fig. 7. It is clear that the condition of any flip-flop at bit time $(t + 1)$ is dependent on the states of all the flip-flops at bit time t. The equation which gives this relation is called the application equation for the particular FF. These equations for the FF's A, B, C and D are also shown in Fig. 7. Analysing and finding the solution for $FF\ A$ only will be

COUNT STATES	A	B	C	D	DECIMAL CELL NUMBER	A_+	B_+	C_+	D_+	APPLICATION EQUATIONS
d_0	1	1	0	0	12	0	1	1	0	
d_1	0	1	1	0	6	0	0	1	1	$A_+ = \Sigma\, 3,5,1,8$
d_2	0	0	1	1	3	1	0	0	1	$+ \Sigma_X\, 0,7,11,13,14,15$
d_3	1	0	0	1	9	0	1	0	1	
d_4	0	1	0	1	5	1	0	1	0	
d_5	1	0	1	0	10	0	1	0	0	$B_+ = \Sigma\, 12,9,10,8$
d_6	0	1	0	0	4	0	0	1	0	$+ \Sigma_X\, 0,7,11,13,14,15$
d_7	0	0	1	0	2	0	0	0	1	
d_8	0	0	0	1	1	1	0	0	0	$C_+ = \Sigma\, 12,6,5,4$
d_9	1	0	0	0	8	1	1	0	0	$+ \Sigma_X\, 0,7,11,13,14,15$
NOT USED	0	0	0	0	0	X	X	X	X	
	0	1	1	1	7	X	X	X	X	
	1	0	1	1	11	X	X	X	X	$D_+ = \Sigma\, 6,3,9,2$
	1	1	0	1	13	X	X	X	X	$+ \Sigma_X\, 0,7,11,13,14,15$
	1	1	1	0	14	X	X	X	X	
	1	1	1	1	15	X	X	X	X	

Fig. 7. Flow table and application equations for a decade counter

Fig. 9. $A_+ = TA' + T'A$ plotted on maps



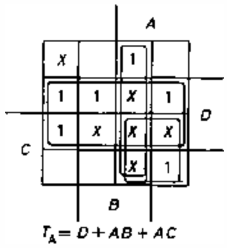


Fig. 10. Input equation of T flip-flop

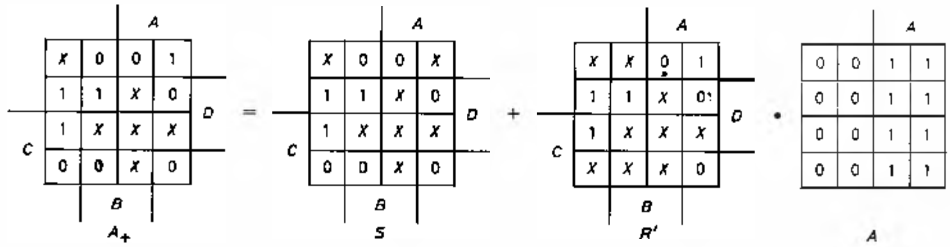


Fig. 11. $A_+ = S + R'A$ plotted on maps

adequate to frame the rules for deriving the input equations.

Input Equations

(a) *D FF*

For the *D FF*, the characteristic equation is

$$Q_+ = D. \text{ For } A \text{ flip-flop, then,}$$

$$A_+ = D_A$$

Thus the plot of A_+ is also the plot for D_A. The plotting is shown in Fig. 8. On simplification the input equation for the delay *FF* is found to be

$$D_A = A'D + B'C'D'$$

(b) *T FF*

For the *T FF*, the characteristic equation is

$$Q_+ = TQ' + T'Q. \text{ For } A\ FF, \text{ then,}$$

$$A_+ = TA' + T'A$$

One can solve for T on the map as shown in Fig. 9. As the T map is to be multiplied by A' which has zeroes on the A half, this half of the T map can be plotted with 'don't-care's.' Similarly the A' half of T' map can be plotted with 'don't care's.' Thus the 1's and 0's of the A' half of A_+ must be plotted as such on the A' half of T map, and 1's and 0's of the A half of A_+ must be plotted as such on the A half of T' map. 'Don't cares' can be plotted as such on both the halves*. Now, since T and T' are to be complementary, the final plotting (Fig. 10) has to be according to the following rules:

(1) On the A half, 1's and 0's are to be plotted as such.
(2) On the A half, 1's are plotted as 0's and 0's as 1's.
(3) On both halves X's are plotted as such.

On simplifying the finally plotted function (Fig. 10) the input equation is:

$$T_A = D + AB + AC$$

(c) *S-R FF*

The characteristic equation is

$$Q_+ = S + R'Q. \text{ For } A\ FF, \text{ then,}$$

$$A_+ = S + R'A.$$

To solve for S and R they can be plotted as shown in Fig. 11.

As R' is to be multiplied by A, which has 0's on the A' half, this half of R' may be plotted with X's. This also leads to the fact that 1's and 0's on the A' half

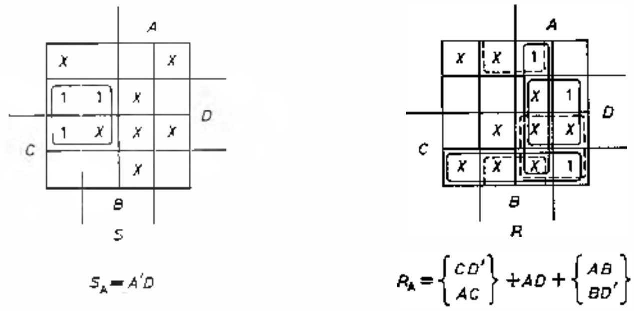


Fig. 12. Input equations of S-R flip-flop

of A_+ must be plotted as such on S. Since S and R cannot be 1 simultaneously, any 1 plotted on the S map must have 0 at the corresponding cell of the R map. But a 0 plotted on the S map, may have either a 0 or a 1, i.e. an X on the corresponding cell of the R map. Thus to take care of the 1's and 0's of A' half of A_+.

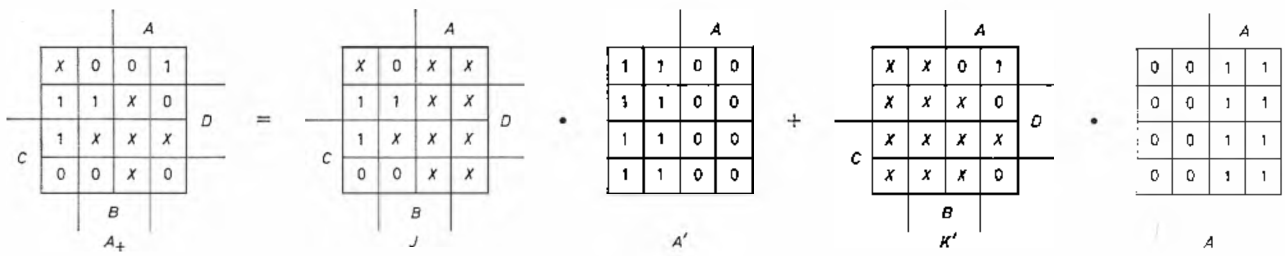
(1) On the A' half of the S map 1's and 0's are plotted as such.

(2) On the A' half of the R map, 1's are plotted as 0's and 0's as X's.

The next question concerns the 1's and 0's of the A half of A_+ map. It should be seen that they can be accounted for by plotting them on the A half of either the S map or R' map or on both. Now, any '1' plotted on the S map has to be a '0' on the R map, i.e.

* *Such a phrase as "1's of A_+ map will be plotted as 0's on the T map" will be used very frequently in this article. This means that in those cells where there are 1's on the A_+ map, there will be 0's in the corresponding cell of the T map. Thus if there are 1's in cells 2, 6, 9 and 14 (decimal notation of Fig. 1) on the A_+ map, there will be 0's in cells 2, 6, 9 and 14 on the T map. This also implies that the minterms, $A'B'CD'$, $A'BCD'$, $AB'C'D$, $ABCD$ for A_+ function are 1's, whereas these minterms for the T function are 0's.*

Fig. 13. $A_+ = JA' + KA'$ plotted on maps



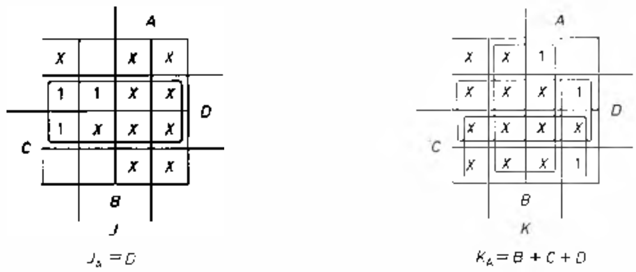


Fig. 14. Input equations of J-K flip-flop

a ' 1 ' on the R' map. Thus, if 1's are to be plotted on the S map, they are to be plotted also on the R' map. On the other hand, if they are plotted on the R' map, they become 0 on the R map, and the corresponding cells of the S may be 'don't-care's'. This will give better scope for simplifying the S map. Hence the rules for the A half of A_+ become:

(3) On A half of S map, 1's are plotted as X's and 0's as 0's.

(4) On A half of R map 1's are plotted as 0's and 0's as 1's.

The rule for X's is obviously:

(5) On both halves and on both maps X's are plotted as such.

On following these rules, the maps of S and R become as shown in Fig. 12, and the resulting input equations are:

$$S_A = A'D$$

$$R_A = \begin{Bmatrix} CD' \\ AC \end{Bmatrix} + AD + \begin{Bmatrix} AB \\ BD' \end{Bmatrix}$$

(d) *J-K FF*

For *J-K FF* the characteristic equation is

$$Q_+ = JQ' + K'Q, \text{ for } A\ FF \text{ then,}$$

$$A_+ = JA' + K'A$$

To solve for J and K, the plotting should be as shown in Fig. 13.

Since J is multiplied by A', and K by A, it is obvious that A half of J and A' half of K' can be plotted with 'don't care's'. Consequently the 1's and 0's of A_+ for the A half are plotted on A' half of J, and those on the A half are plotted on the A half of K'. Thus the rules for J and K are as follows:

(1) On the A' half of the J map, 1's and 0's are plotted as they are, but the A half is plotted with X's.

(2) On the A half of K map, 1's are plotted as 0's and 0's as 1's; but the A half is plotted with X's.

(3) The X's are plotted as they are on both maps.

Thus the final plotting is as shown in Fig. 14. The input equations are:

$$J_A = D$$

$$K_A = B + C + D$$

(e) *S-R-T FF*

Here the characteristic equation is

$$Q_+ = S + TQ' + R'T'Q, \text{ For } A\ FF \text{ then,}$$

$$A_+ = S + TA' + R'T'A$$

As S is not multiplied by any function, the 0's of A_+ must be plotted as 0's on S. Hence, the rule

(1) All 0's of A_+ are plotted as 0's on the S map.

The 1's of A_+ for the A' half must be plotted on the A' half of either S or T as the R' and T' are multiplied by A which has 0's on the A' half. Since S and T cannot be 1 simultaneously, where there is a 1 on S, there must be a 0 on the corresponding cell of T and vice-versa. This can be indicated by plotting T' on S and S' on T on the A' half of S and T. Hence the second rule is

(2) 1's of the A' half are plotted as S' on T, and as T' on S.

Since on these cells either S or T is 1 and R, S, and T, cannot be 1 at the same time, there must be 0's on the corresponding cells on R, that is, 1's on R'.

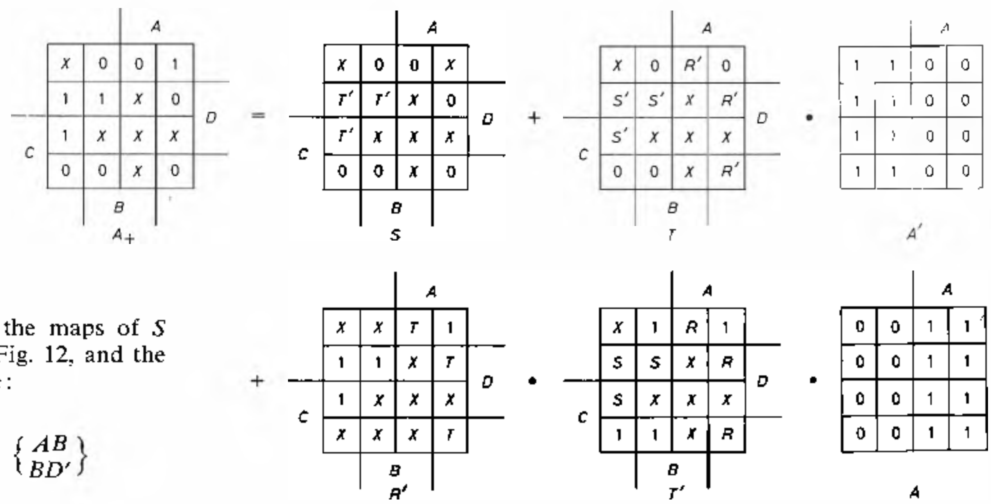
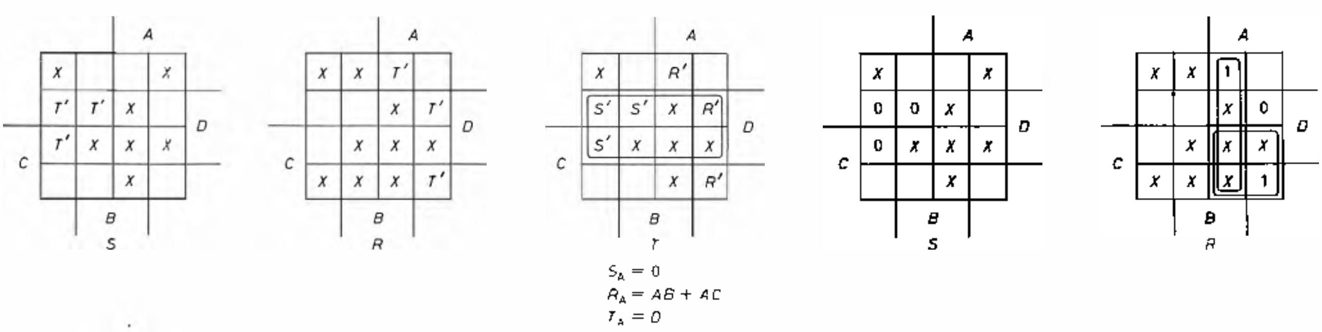


Fig. 15. $A_+ = S + TA' + R'T'A$ plotted on maps

Fig. 16. Input equations of S-R-T flip-flop



Thus:

(3) 1's of A' half are plotted as 0's on R map.

On the T' map these (1's of A' half) obviously will be S. As the T map is multiplied by A', which has 1's on the A' half, the 0's of A' half of A_+ must be plotted as 0's on T map. Since both S and T are 0's, and the R' map is multiplied by A, for 0's on A' half, R and R' can have X's.

(4) 0's of A' half are plotted as 0's on T map and as X's on R map.

By following similar reasoning as was used to derive rules (3) and (4) of S-R flip-flop, it can be seen that for 1's of the A half, both R' and T' must be 1, and S can have an X on the corresponding cell, giving the rule,

(5) 1's of the A half becomes X's on the S map, and 0's on R and T map.

0's on the A half can be plotted either on R' or T', but not on both. These can be indicated by plotting R' on T and T' on R. Thus:

(6) 0's on the A half remain 0's on the S map, and are plotted as T' and R' on R and T maps respectively.

The final plotting following the above rules is shown in Fig. 15.

It is now a matter of making S, R and T as simplified as possible. In the above case T can be made equal to D as shown. With this choice the values of T's and T''s in the S and R map become fixed, and the maps are as shown in Fig. 16. This yields the value for S and R, as follows:

$$T_A = D$$

$$S_A = 0$$

$$R_A = AB + AC$$

Conclusion

During the analysis as outlined above, some rules have been framed for deriving input equations by plotting the application equation on individual maps. For ready reference these rules are summarized in a tabular form in Fig. 17. By following these rules the maps for input lines can be obtained directly without going through the intermediate plotting such as has been done in Figs. 9, 11 13, and 15. This not only saves time, but also renders the derivation almost effortless.

FF	On map	On Q' half For 1	On Q' half For 0	On Q half For 1	On Q half For 0
D	D	1	0	1	0
T	T	1	0	0	1
$S-R$	S	1	0	X	0
	R	0	X	0	1
$J-K$	J	1	0	X	X
	K	X	X	0	1
$S-R-T$	S	T'	0	X	0
	R	0	X	0	T'
	T	S'	0	0	R'
On all maps X's are plotted as X's					

Fig. 17. Tabular listing of rules for deriving input equations for all flip-flops

It may be seen that for all FF's *except the S-R-T* there is no ambiguity, and the solutions are unique. For the *S-R-T FF*, however, there is more than one solution, and the designer has to find the best one to suit his problem.

Acknowledgment

The author is grateful to Dr. A. K. Kamal, Professor and Head of the Department, for his encouragement in the work.

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Industrial Television Cameras

A miniature television camera for the internal inspection of pipes down to 2¼in diameter has been developed by Pye Laboratories Ltd for use in the complicated pipeline networks of many major industries.

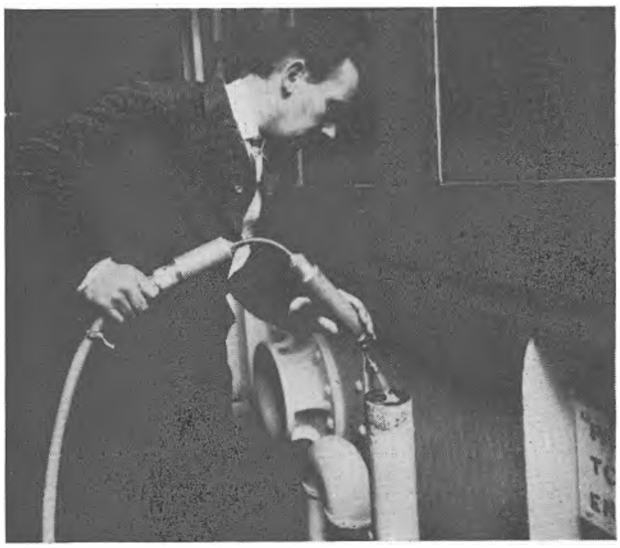
This articulated pipe inspection camera, claimed to be the smallest of its type in the world, is only 2¼in in diameter. The entire length of the camera, its amplifier and lighting system is 28in. It has been specially designed for small bore pipe inspection and will save time and cost in industries throughout the world, including oil refineries, chemical plants, power stations, gasworks and shipping companies.

Pipe inspection and maintenance has been an acute problem in industry because long lengths of pipe have so far had to be physically disconnected for periodic checks and repairs. The unique construction of this new Pye camera will help overcome these problems by enabling continuous lengths of up to 1 000ft to be inspected.

The camera itself is articulated so that it can pass easily round bends in piping. It consists of two sections, a tube assembly and amplifier unit joined by a flexible metal coupling. An articulated optical assembly with built-in lighting is fitted to the front end of the tube assembly. The camera is lightweight and completely waterproof and is supplied complete with control and monitor unit.

Another camera in the new Pye range is designed for the thorough inspection of the interiors of pipes with larger diameters from 3in to 42in. This one is supplied with lighting and centring attachments to accommodate pipes within this range. Its applications in the civil engineering field include gas, sewer and oil installations.

The articulated television camera being lowered down the ventilation pipe of a double-bottomed tank on board the Union Castle liner 'Reina del Mar'



A Combined Peak and Average Value Rectifier Tachometer

By A. Kislovsky*

A tachometer for spark ignition internal combustion engines is described. It employs a diode pump circuit as in previous designs but, due to the choice of time-constants it is inherently insensitive to transients caused by contact bounce etc.

(Voir page 133 pour le résumé en français: Zusammenfassung in deutscher Sprache auf Seite 140)

IN an internal combustion engine the ignition spark repetition frequency is directly proportional to the rotation speed of the motor. The coefficient of proportionality depends on the type of the motor. Thus, for a four cylinder, four stroke engine two sparks per main shaft rotation are produced.

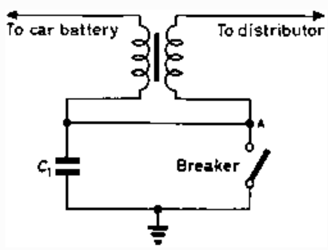


Fig. 1. Ignition circuit

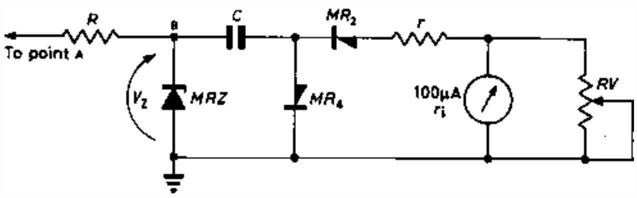


Fig. 2. Tachometer circuit

There exists a number of electronic speed-of-rotation meters based on this relationship. The basic circuit in these meters is a kind of a monostable multivibrator which, when triggered at spark repetition frequency, produces a train of normalized voltage-time pulses. The average value of this train is proportional to the trigger pulse frequency, and thus the meter deviation is proportional to the speed of rotation.

A conventional ignition circuit is represented in Fig. 1. When the circuit breaker is opened, a transient oscillatory voltage across the capacitor C_1 is generated. After the transient has finished, the voltage across the capacitor is established at car battery level. When the breaker closes, a certain current flows in the transformer primary winding and after the next opening of the breaker the whole process is repeated.

If the circuit represented in Fig. 2 is connected across the circuit-breaker, the voltage swing at point B will be limited by the breakdown voltage of diode *MRZ*, and thus a train of rectangular pulses of transient voltage frequency is produced. The resistances in the charge and in the discharge paths of the capacitor C are not equal and are chosen so that, as far as transient voltage frequency is concerned, the circuit represents a peak value rectifier.

Consequently, the charge stored in the capacitor will be $Q = C \cdot V_z$. When the circuit breaker is opened the capacitor will discharge through the diode MR_2, the frequency of the breaker motion being much lower than the transient frequency. As a result a constant charge is delivered to the resistance r and to the following network, for every period of the breaker's motion. Thus a meter deflexion proportional to the motor speed rotation is obtained. For a four cylinder, four stroke motor the total current is given by

$$i = \frac{C.V_z}{30} N$$

where V_z is the Zener diode breakdown voltage and N is the number of turns-per-minute of the motor. Only a part of this current is taken to yield a correct meter deflexion.

As far as transient voltage frequency is concerned, the circuit represents a peak value rectifier, and for the breaker motion frequency it acts as a half-wave average value rectifier.

A closer examination reveals that in its average value rectifier part the system is virtually identical to the well known diode-pump circuit frequently applied in various ratemeter types of measuring instruments.

A diode pump circuit has already been applied in an earlier electronic tachometer[1], where the filtering action was accomplished by a (saturable) choke.

It is possible, however, by a judicious choice of the time-constants to achieve a filtering action at the same time with the 'pumping' and compel the system to respond only to the modulating frequency of the pulse train. The contact bounce problem is thus eliminated. This opportunity of dual, simultaneous action of the diode pump circuit does not seem to have found any practical application up to now.

It should be pointed out that, because of this dual action, the meter deviation versus motor speed is not a perfectly straight line, not even theoretically. There is always a certain discharge of the capacitor C during the intervals between two succeeding pulses of transient repetition frequency. But, for many practical applications the resulting linearity is sufficient.

In any practical realization the total resistance in the discharge path is large enough, and thus the diode MR_2 can be eliminated.

In practice, several contradictory factors must be taken into account. In order not to overdamp the oscillatory ignition circuit, the resistance R must be large; meter current is directly proportional to the value of the capacitor, but it must not be too large so that both charge and discharge time-constants are small enough to enable complete charging and discharging of the capacitor even at the highest breaker frequencies.

A good compromise has been achieved for the follow-

* *Cie Electro-Mécanique, Paris.*

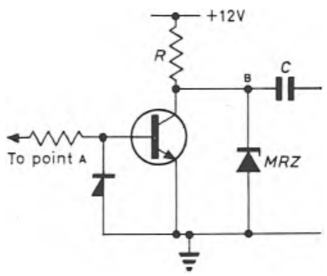


Fig. 3. Single stage pulse amplifier

ing values of elements: $R = 1k\Omega$, $C = 0{\cdot}1\mu F$, $RV = 5k\Omega$, $r = 3{\cdot}3k\Omega$, $r_1 = 1k\Omega$. A 7V breakdown diode and a 500mA silicon diode have been used.

In order to increase sensitivity, the value of the capacitor can be increased. Here, R must be decreased so that the resulting time-constants are not prohibitively long; too low a value of R might have a damping effect upon the free-oscillating circuit; to avoid this, a single-stage pulse amplifier (Fig. 3) should be added.

In some cases a four diode bridge instead of a simple rectifier circuit may be used. The basic principle, must, of course, be retained—the system must perform as a peak rectifier for transient voltage frequency and as an average value rectifier for breaker motion frequency. The time-constants in the charge and the discharge paths must be as different as practicable. This can be achieved by adding a resistance in series with one of the diodes (in the discharge path). Here the meter current is increased twice in comparison to the simple system of Fig. 2.

REFERENCE

1. WRIGHT, M. J. An Electronic Tachometer for Automobiles. *Electronic Engng. 31*, 599 (1959).

The 'Stereoscan' Electron Microscope

This new scanning electron microscope, manufactured by the Cambridge Instrument Co. Ltd., is the first in the world to be designed specifically for large scale production and has a special field of application not wholly covered by any other type of microscope. It is complementary to optical and electron transmission instruments and greatly extends the range of investigations to which these and other types of microscopes can be applied.

The instrument embodies a new technique of examining directly the microtopography of solid bulk specimens with rough surfaces or other characteristics that make them difficult or impossible to observe either directly or by extraction replica methods in a conventional transmission instrument. The displayed image, although similar in appearance to an optical micrograph, can normally be resolved to 200 Å with a depth of focus that is always 300 times greater than that of a light microscope when both are adjusted for optimum performance. It has a direct reading magnification system providing a useful range of magnifications from 50 × to 20 000 × corresponding to scanned areas of 2mm to 5μm square.

In the 'Stereoscan' a primary beam of electrons from a heated tungsten filament is focused to a fine spot on the specimen surface and made to scan across it in a square raster by deflexion coils and associated circuits. Electrons leaving the surface due to the action of the primary beam are collected and amplified, the resulting signal being used to modulate the brightness of a c.r.t. beam, which is scanned in synchronism with the primary beam but at much greater amplitude.

Contrast is produced by the variation in number of electrons emitted or reflected from different parts of the specimen, the final image having a marked three-dimensional appearance but with none of the severe shadowing or foreshortening that is characteristic of transmission instruments adapted for reflection electron microscopy. The specimen, which requires little or no preparation, is situated in a region of weak magnetic or electrostatic fields and is not subjected to strong heating from the illumination system. The specimen stage allows objects of up to 12mm in diameter and several millimetres thick to be manipulated in any required orientation while under observation.

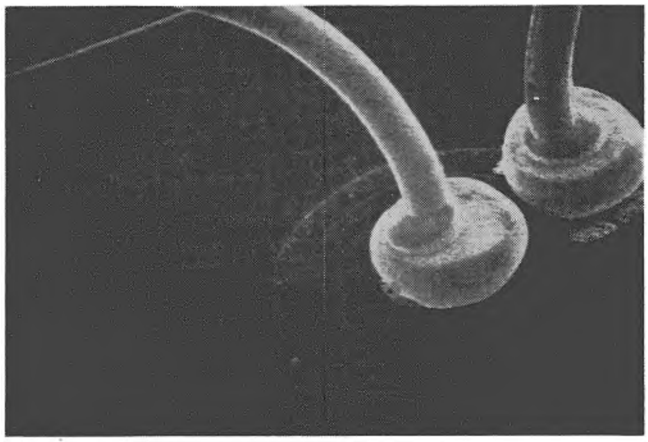
The instrument has proved extemely useful in examining metallic or ceramic fractures; the interior of holes such as small-bore tubes or capillaries; woven or unwoven natural and synthetic fibres; plastic, pulp and paper surfaces; insects; biological samples, including hairy and whiskered growths; etched surfaces; powdered or fragmented materials; etc, often providing topographical detail that can be obtained in no other way. In addition, the scanning action of the beam and the high efficiency of electron collection make it possible to examine fragile specimens, such as fine fibres and flaking substances, without any risk of damage due to overheating.

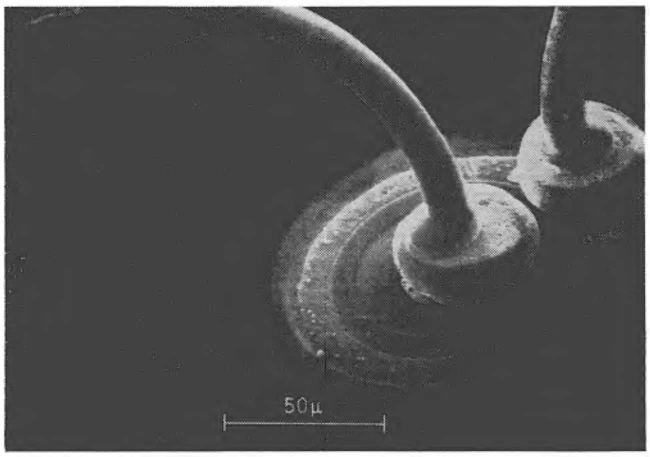
Potential differences between adjacent areas of a specimen surface alter the trajectories of low energy secondary electrons emitted, so superimposing voltage contrast on the topographical contrast of the image. This feature provides a unique method of assessing the behaviour and performance of microcircuits, such as transistors, making it possible to determine the position and magnitude of electrical and structural faults, without damage to the specimen. Even more important, it also provides a means of predicting the eventual breakdown of miniature components that might otherwise be regarded as sound.

Furthermore, low accelerating voltages down to 1kV can be used to examine insulators, such as ceramics, without the need for overlaying specimens with a conducting film. Even at such low voltages the resolving power is at least five times greater than that obtainable with the best optical microscopes.

The 'Stereoscan' is based on many of the techniques developed by Professor C. W. Oatley and his co-workers in the Engineering Laboratories of Cambridge University and is being continuously modified to take advantage of the latest developments of this and other research groups.

Microphotograph of silicon transistor: (top) transistor unbiased, (bottom) collector reverse biased





LETTERS TO THE EDITOR

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

An A.F. Power Amplifier with Crossover Feedback

Dear Sir,—Dr. Bailey's new amplifier circuit (December 1965) undoubtedly achieves a very high standard of performance. There is, however, a remarkable similarity in principle between his design and one I referred to in British Patent No. 933062, which was applied for in September 1960. The essential idea expressed in the Patent was to replace the phase retard network with an internal feedback loop which converted the amplifier into a feedback integrator. The loop from the secondary is connected in the conventional way but the feedback resistors were lowered in value to improve the stability margin when driving a reactive load. The two loop voltages are additive and permit full feedback over the entire audio band in a similar manner to Dr Bailey's circuit. The phase retard network in the 1st stage load in the conventional circuit is a profound disadvantage. Firstly the feedback at high frequencies is reduced which leaves the high frequency iner-modulation products at too high a level. The second disadvantage is that the effective drive voltage on the first stage becomes excessive at the highest frequencies when the main loop feedback exceeds 30dB.

The effective input to the first stage is

$$E_{eff} = E_{in} - \beta A E_{in}.$$

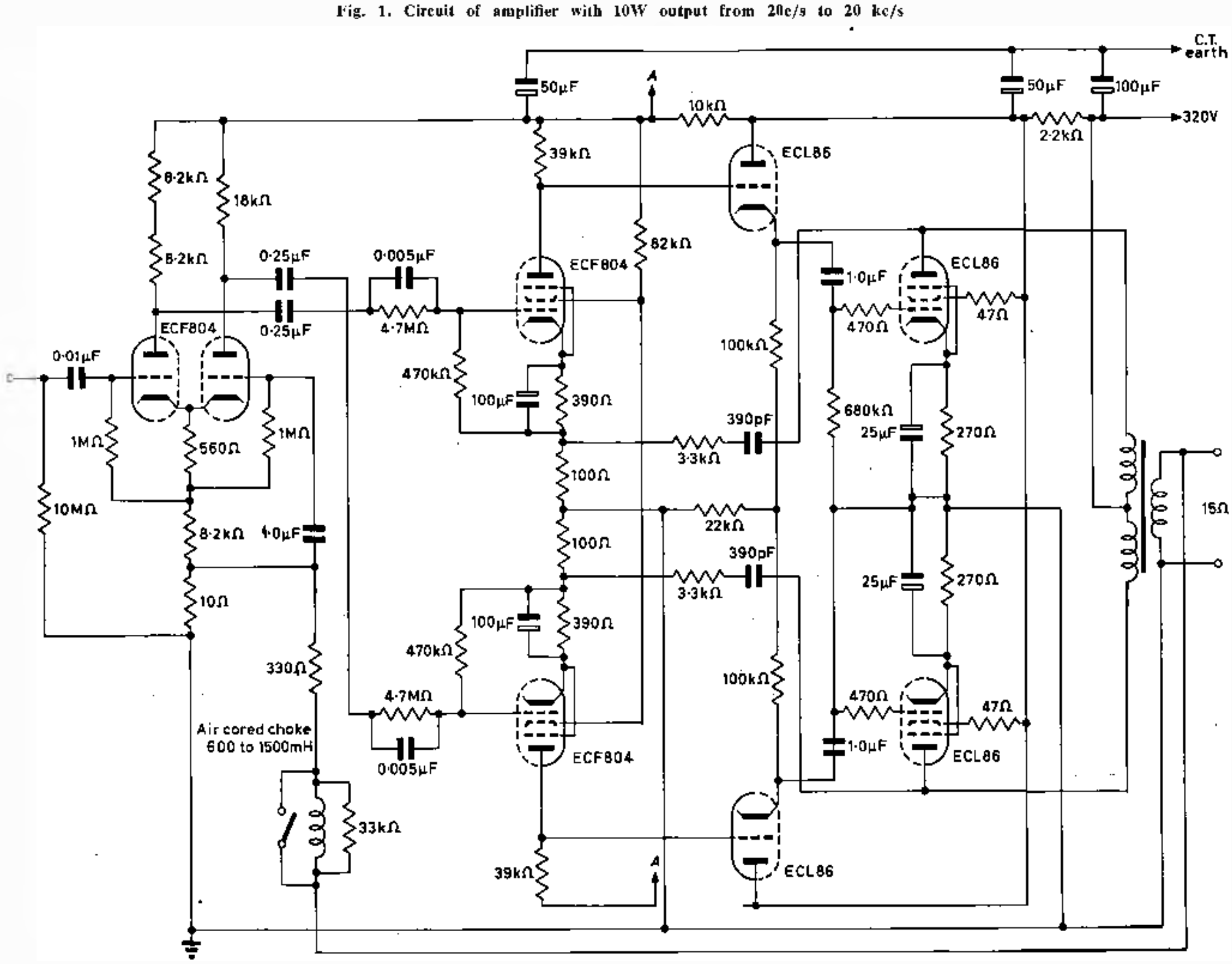
where E_{in} = input signal βA = loop gain. It is clear that at the band edges the loop gain is falling and the effective drive voltage is such as to overload the first stage since the phase retard network has a low impedance at the highest frequencies. This disadvantage is also avoided in Dr Bailey's system.

A typical circuit embodying these principles shown in Fig. 1, illustrates the essential features of the circuit. The stability margin is adequate with 40dB feedback for all moving-coil transducers with the secondary loop switch closed. However, it is desirable with some capacitive transducers to reduce feedback from the secondary winding at frequencies above 500 c/s. This is done by opening the switch in which case the choke removes the feedback at higher frequencies. The circuit may be used for power levels of up to 100W, but the circuit illustrated will deliver 10W from 20c/s to 20kc/s.

At this power level the intermodulation products are more than 65dB below the desired signal throughout the audio band. The wide bandwidth required in the pre-amplifier and driver stages is achieved by using cathode-followers to drive the output valves. This also prevents Miller effect in the output valves from introducing a lag in the desired frequency band.

The so-called ultra-linear output circuit has been discarded in favour of straight pentode operation because a worthwhile improvement in linearity is not obtained

Fig. 1. Circuit of amplifier with 10W output from 20c/s to 20 kc/s



unless the output valves are electrically matched by a differential heater balance control which is barely economic in a commercial circuit. An automatic phase balancing circuit to provide equal drive to the output valves is possible if the cost of an extra valve is acceptable. In this case the valve is used as a cathode load for the phase splitter and feedback is taken from the common resistor in the driver stages. An important feature of the circuit is that the input signal must come from a low impedance source such as a cathode-follower on the pre-amplifier. If the gain control were placed as in Dr Bailey's circuit the control would have to be of low impedance or an extra lag time-constant will appear when the control is at an intermediate position. The response of the circuit to square waves is almost identical to Dr Bailey's circuit.

Yours faithfully,

J. R. OGILVIE,
Sevenoaks, Kent.

The author replies:

DEAR SIR,—I was very interested in the comments made by Mr. Ogilvie in his letter regarding the use of conventional step networks. There is no doubt that where a step-network is used at high signal level points, the high-frequency distortion figures can be made very poor. The distortion is frequently worse than would be expected, as the load-line for the valve driving the network becomes an ellipse inside the step frequencies. Mr. Ogilvie's circuit overcomes the problem of elliptical load-lines in a very neat manner, but the first valves will still have to provide an enhanced drive level at high frequencies. As the signal level at these anodes will only be small, this should not be a serious disadvantage.

It appears however that Mr. Ogilvie's circuit was evolved as an improvement on step-networks, which it most definitely is. The circuit described in the article was evolved to improve overall feedback conditions, and step-networks will be an additional stabilizing factor. There is however a close relationship between the final circuits in spite of the different lines of approach.

I was interested to note that an inductor was advantageous when the circuit was feeding a capacitive load. The use of such an inductor gives a falling secondary feedback response, and if the values are correct will give a complementary filter action as in the amplifier described in the article. There is no doubt however that straight feedback from the output transformer secondary alone is inadequate, if the best performance is to be obtained.

Yours faithfully,

A. R. BAILEY,
Bradford Institute of Technology.

Tunnel Diode Switching Circuits

DEAR SIR,—With great interest I have read Messrs. L'Archeveque's and Sayers' article, titled 'The design of some tunnel diode switching circuits', which appeared in the October 1965 issue.

Availing myself of the opportunity I would like to describe a very simple low level detector circuit for fast pulses, which has been used as a discriminator of pulses, originating from a type 56AVP photomultiplier, with a rise time of ~5nsec and a duration of 20nsec.

The detector stage consists of a type 2N706 transistor, whose emitter is connected to a type IN3150 tunnel diode

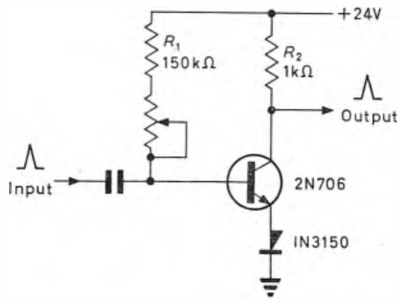


Fig. 1. Low level detector stage

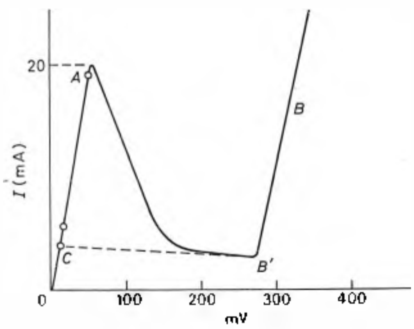


Fig. 2. Current/voltage characteristic

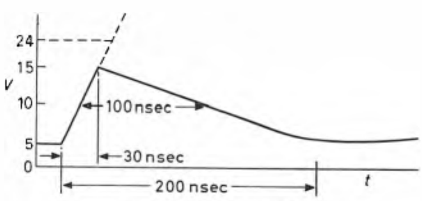


Fig. 3. Output pulse

with a peak current of 20mA, and operates as follows:

The current of the emitter is set at 19mA (Point A in Fig. 2), by means of R_1 and the series rheostat. The potential of the collector is thus set at $24 - 19 = 5V$ (Fig. 3). If now a positive going pulse is applied at the input, the emitter current exceeds the I peak of the tunnel diode and causes it to trigger (Point B in Fig. 2).

The emitter then instantaneously acquires a potential of +450mV to ground and the transistor swings to cut off. The collector voltage quickly approaches the value of 24V (Fig. 3) while the point of operation of the tunnel diode drifts from B to B' (Fig. 2).

The tunnel diode however swings back to point C, before the collector voltage reaches the value of 24V. This means that the reason which caused the transistor to swing to cut off is no longer present, whereupon the emitter current gradually returns to its initial value of 19mA and the collector voltage to 5V.

The obtained positive going output pulse, for positive going input pulses, as those mentioned, is shown in Fig. 3.

Yours faithfully,

CH. MANTAKAS,
Nuclear Research Center "Demokritus",
Athens, Greece.

A Property of the Correlation Function of Hilbert Transforms

DEAR SIR,—Let X denote a non-stationary random process and $\hat{X}$ its Hilbert transform:

$$\hat{X}(t) = 1/\pi \int_{-\infty}^{\infty} \frac{X(s)ds}{t-s}$$

Let $R_{X\hat{X}}(t, \tau)$ denote the cross-correlation function of X and $\hat{X}$. Urkowitz[1] has derived a number of properties for this function analogous to those for the stationary case (see for example Papoulis[2]). His last result (Theorem 5) that the cross-correlation is zero for zero time shift is established under the assumption that the process has a band-pass character. His proof also involves the introduction of the time-dependent spectral density function.

The purpose of this note is to establish this result assuming merely that the auto-correlation of the process X is an even function of the time shift τ. The result of Urkowitz then follows as a special case. Let $R_X(t, \tau)$ denote the auto-correlation function of X. Then (as seen from the proof of Theorem 1 of Urkowitz)

$$R_{X\hat{X}}(t, \tau) = 1/\pi \int_{-\infty}^{\infty} \frac{R_X(t, s+\tau)}{s} ds$$

so that

$$R_{X\hat{X}}(t, 0) = 1/\pi \int_{-\infty}^{\infty} \frac{R_X(t, s)}{s} ds$$

The assumption that $R_X(t, s)$ is an even function of s leads at once to the desired result

$$R_{X\hat{X}}(t, 0) = 0$$

Yours faithfully,

H. KAUFMAN,
McGill University,
Montreal.

REFERENCES

1 URKOWITZ, H. Pre-envelopes of Nonstationary Bandpass Processes. *J. Franklin Inst* 277, 31 (1964)
2 PAPOULIS, A. Probability, Random Variables, and Stochastic Processes. p. 356 (McGraw-Hill, 1965)

SHORT NEWS ITEMS

The Thirteenth International Exhibition on Modern Electronics organized by the Yugoslav Electronics, Telecommunications, Automation and Nucleonics committee (ETAN) is to be held at Ljubljana, Yugoslavia on 4 to 9 October this year.

During the Exhibition, a conference will be held dealing with electrical measurements and measuring equipment, and on component reliability.

Further details are obtainable from GOSPODARSKO RAZSTAVISCE (Ljubljana Fair), Ljubljana, Titova 50, Yugoslavia.

Pye TVT Ltd, of Cambridge, has been awarded a contract valued at £100 000 by the Swiss Post and Telegraph Administration for the supply and installation of two 10kW u.h.f. television transmitters. The transmitters will be part of the expansion of the Swiss Television Service.

This is the first order ever obtained by Britain for the supply of u.h.f. television transmitters to Europe and is the first u.h.f. television transmitter to be purchased by Swiss PTA. The equipment will be installed at Mount Rigi in central Switzerland.

EMI Electronics Ltd has completed a £40 000 contract for the supply and installation of television studio equipment for the Zambian Broadcasting Corporation.

The equipment is now in operation at the ZBC's broadcasting stations in Lusaka and Kitwe.

To meet the scheduled start of broadcasting from the new Lusaka station, EMI assembled, delivered by air and installed the equipment within 6½ weeks of receiving the order. The equipment included two vidicon cameras, telecine, picture and waveform monitors, sync pulse generators, vision mixer, amplifiers and cabling. Installation work was carried out in the record time of five days.

Equipment installed at the Kitwe station included three vidicon camera picture waveform monitors, vision mixer, slide projector and pulse equipment.

EMI Electronics Ltd has also supplied the New Zealand Broadcasting Corporation with a series of microwave links to improve and extend the country's television coverage.

The latest of the link units (type ML4C) to be despatched to NZBC are capable of transmitting colour or monochrome television signals.

The link equipment, valued at about £65 000 provides facilities for transmitting sound and vision signals over long distances and operates in the 7 000Mc/s band.

"The Use of S.I. Units" is the title of a recently published booklet (PD5686) by the British Standards Institution giving an explanation in simple terms of the British Standard B.S. 3763 "The International System (SI) Units" published in 1964, and containing the newly rationalized set of metric units known as the Système Internationale d'Unités (SI).

The new booklet outlines the historical background to the international recognition of SI metric units, explains what they are, comments on the use of the newton as the unit of force, and indicates also how the units may best be introduced in industry. The more important British Standards which provide further information on the use of SI units are listed in an appendix.

One copy of the booklet is available post free, on application to the British Standards Institution, 2 Park Street, London, W.1, additional copies price 1/- each, post free with the usual discount for bulk orders of 100 or more.

Supplementing this booklet is one published by H.M.S.O. entitled, "Changing to the Metric System, Conversion Factors, Symbols and Definitions" (price 3/6). This has been prepared by staff of the National Physical Laboratory, the responsible authority for the nation's reference standards of measurement.

The booklet covers the ordinary quantities for which British units are used in trade, technology and science, including not only weights and measures, but work in which heat, light and other units are involved. Information is given on conversions relating to the customary c.g.s. system, but the main emphasis of the booklet is laid upon the use of the new international system (SI), which was agreed at a conference in Paris in 1960.

A conference on electrical network theory and design has been planned to take place during the week commencing 11 September this year at the University of Newcastle upon Tyne. The Conference, organized jointly by the Ministry of Technology, the University, the Institution of Electrical Engineers and the Institution of Electronic and Radio Engineers, includes discussion of the following topics: analysis of linear passive and linear active networks, nonlinear and time-variant networks, the place of network theory in the design of integrated circuits, and the application of computers to network design.

Lectures will be given in the Electrical Engineering Department of the University and accommodation will be provided in one of the Halls of Residence. Bookings, details of papers offered (together with synopsis of up to 200 words), and requests for further information should be sent to Dr. A. G. J. Holt, Electrical Engineering Department, Merz Court, The University, Newcastle upon Tyne 1.

The General Electric Company Ltd (G.E.C.) and Associated Electrical Industries Ltd (A.E.I.) announce that they have agreed to divide the trading of their jointly owned Company, S.G.E. Signals Ltd, into separate road and rail businesses. The G.E.C. will own and manage the business of road signalling, and A.E.I. that of rail signalling.

S.G.E. Signals Ltd will change its name to G.E.C. Road Signals Ltd and be controlled by G.E.C. (Electronics) Ltd. It will continue to operate from the present address at East Lane, Wembley and to market its vehicle actuated traffic control systems under the trade name 'Autoflex'. The new arrangements will enable forty years of experience in road traffic signals to be combined with the latest electronic techniques including data transmission equipment and digital computers.

Associated Electrical Industries International Ltd, on behalf of AEI Electronics, is the main contractor to the Government of Saudi Arabia for the supply of the £25M ground environment radar defence system—including radar stations, data handling and communications.

The completely new radar system—known as type 40 series radar—is very similar to the equipment currently being offered in the AEI proposals as part of the ITT NADGE consortium bid for the £100M NATO integrated air defence.

It provides instantaneous three-dimensional position data on all aircraft within radar cover and is suitable for civil and military air traffic control.

The type 40 series radar system produces accurate information for the microelectronic data handling computers, enabling the computers to calculate precise flight paths for the Lightning interceptors, so that the aircraft and missile defences can be brought into action at the optimum time and efficiently controlled.

A high-capacity, nation-wide communications network passes up-to-date information to a central control.

Vital electronic units are duplicated and most of the electronic units use a continuous, automatic testing system giving instantaneous fault indication and automatic switchover to the stand-by equipment in the event of a failure.

The Czechoslovak State Commission for Technology and the Czechoslovak Scientific and Technical Society are proposing to hold an International Computer Exhibition at Prague on 12 May to 8 June this year to which British manufacturers of computers are invited.

Exhibitors will also be invited to take part in a ten-day symposium, details of which will be announced in the near future.

Further details are available in the United Kingdom from S. Black, 68 Queen Street, London, E.C.4.

Telefunken, Berlin, has received an order from the Indian Government for the installation of a precision approach radar system, type PAR-T-3, at the Indian airport of Dabolim.

The system is similar to that already installed by Telefunken at the Bombay and Calcutta airports.

Thomson-Tele-Industrie, a subsidiary of the Compagnie Française Thomson-Houston has recently equipped the international airport at Mexico with a closed circuit television system similar to that supplied by the company at the Orly and Le Bourget airports at Paris.

The system consists of ten television cameras and thirty receivers in addition to a control centre.

The cameras and receivers are divided into two groups, one of which is for use by the airport staff in the supervision of airport equipment and of aircraft landings.

The other group is for the public display of arrival and departure information throughout the airport building.

Racal (Canada) Ltd, Ottawa, has been awarded a contract valued at over \$1·6M by the Canadian Department of Defence Production for the supply of automatically tuned high-frequency single sideband radio systems known as Racal 'Speedrace' systems which are already in service with the R.A.F.

The purchase of this equipment represents the first phase in the multi-million dollar augmentation and improvement programme of ground-to-air communications for aircraft of the Transport and Maritime Commands of Canada's Defence Forces.

The majority of the units will be manufactured in the Berkshire factories of the Racal Group, the remainder will be made in North America under the responsibility of Racal (Canada) Limited.

Racal 'Speedrace' equipment has now been ordered by Government authorities of eight different countries and it is expected that by the end of this year 'Speedrace' equipments will be in use on a world-wide basis in 20 countries.

The 'Speedrace' h.f. communications family of systems incorporate full remote control of both receiver and transmitter from a single remote location, self-tuning linear amplifiers of 1 to 10kW p.e.p. with automatic matching of the transmitter output circuits to the load presented by the aerial system.

The systems are available in static, transportable or fully-mobile versions and are suitable for all spheres of military communications, point-to-point circuits, ship-to-shore installations, in meteorological services and in any medium-or-long-distance radio communications application.

Operational frequency of transmitter and receiver is controlled by synthesizers in steps of 100c/s between 2 and 30Mc/s, using a precise frequency standard as their input reference to achieve frequency stability of a few parts in 10^{10} per day in the more complex systems.

"Soviet Electronic Abbreviations" is the title of a dictionary compiled by A. J. Michaels and edited by D. I. Mitchell, GEN-TECH Research Associates, Inc., City Post Office, Box 1815, Washington 13, D.C. (USA). 170 pages, \$5·95.

This dictionary contains some 1 800 definitions of abbreviations of electronic terms appearing in Russian technical literature dealing with the different branches of electronics such as: telecommunications (radio, telegraphy, telephony, television, wire, etc.), space communications, radar, navigation, computers, X-ray, railroad applications, etc.

225 Westrex Teletype Model 35 Printers are to be used by British European Airways in the second stage of their Automatic Seat Reservation System based on twin Univac 490 real time computers.

The printers are an integral part of the Uniset complexes used by BEA telephone sales clerks at West London Air Terminal to interrogate the computer system as to seat availability on any of BEA's flights throughout the U.K. and Europe, and to subsequently insert into the system details as to passengers' names, addresses, telephone numbers, special requirements, etc.

A new British Standard, B.S. 3934—*Dimensions of semiconductor devices* has been prepared to meet the need within the electronics industry for a record of the dimensions of semiconductor devices in general use.

The primary object of the drawings of outlines, bases and gauges is to indicate the space to be allowed for semiconductor devices in an item of equipment, together with other dimensional characteristics needed to ensure mechanical interchangeability. Electrical and thermal requirements for full interchangeability are not covered by the standard. Also included in the standard is a section giving guidance on the interpretation and use of the drawings and an index for quick cross reference.

Copies of B.S. 3934 may be obtained from the B.S.I. Sales Branch, 2 Park Street, London, W.1. Price £4 10s. 0d. each. (Postage 3s. will be charged extra to non-subscribers.)

The Inner London Education Authority is proposing to have installed in London what is considered to be the world's largest closed circuit television education network.

Over a thousand primary, junior and secondary schools as well as colleges of further education of the Inner London Education Authority are on a network planned by the Post Office. Subject to the Authority's acceptance of the GPO quotation, the service will be in operation to some schools by September, 1968, and to 80 per cent of the schools a year later.

The education network will be hired from the GPO in much the same way as telephones, the rental including provision for maintenance. Its electronic distribution equipment, using transistors, will be capable of putting out at least six simultaneous programmes over coaxial cables using frequency division multiplex techniques. This means that the teacher can choose his programme by switching a simple selector. The Post Office has made use of many years' experience providing high-quality television links for the BBC and ITA, in designing the London schools network, which is equally effective for carrying 405 or 625-line vision signals in monochrome or colour. Sound signals associated with 405-line transmissions will be amplitude-modulated and those associated with 625-line signals will be frequency-modulated. This arrangement is similar to that used in standard broadcast practice, allowing for a reasonably wide choice of receivers.

The Post Office is to instal an experimental press-button telephone system for a trial period of three months to 300 subscribers at the Langham Telephone Exchange, London.

The press-button telephone will have ten buttons instead of the ten finger holes in the dial of the ordinary telephone.

Two different experimental systems will be put on trial at the Langham exchange. One uses voice frequency tone signalling generators inside the telephone working into special equipment in the exchange.

In the other system the press-button telephone is connected to a mains operated convertor, which can be placed under the table or near the telephone and does not need any special equipment at the exchange. This can be used, therefore, for internal calls on existing types of private automatic branch exchange, as well as to other numbers on the public network.

NEW BOOKS

Principles and Design of Linear Active Circuits

By M. S. Ghausi. 621 pp. Royal 8vo. McGraw-Hill Book Co. 1965. Price 132s.

THIS book is the latest addition to the well-known McGraw-Hill Electrical and Electronic Engineering Series, and the author, an Associate Professor of Electrical Engineering at New York University, maintains the high standard set by authors such as Terman, Seely, and Millman.

The first three chapters give a review of the basic concepts involved in the analysis of linear circuits, and introduce the required mathematical techniques. Analysis by the up-to-date state-space and scattering parameters techniques are included, and are explained particularly well.

Chapter 4 is devoted to filter design, and in chapter 5 the characteristics of vacuum-tube devices are considered. In chapters 6, 7, and 8 semiconductor electronics, the theory of p-n junction diodes, and junction transistors are discussed. Chapter 9 is an especially lucid development of transistor small-signal equivalent circuits, including a theory and equivalent circuit for the drift transistor. In chapter 10 biasing techniques for both vacuum-tubes and transistors are given, and selection and stabilization of the operating point are considered.

The next five chapters are concerned with the design of amplifiers. Wideband cascaded and additive amplifiers, feedback amplifiers, and band-pass amplifiers are included in the treatment. Chapter 16 deals with the analysis and improvement of the transient response of linear circuits. In chapter 17 the synthesis of active *RC* transfer functions is considered. Linvill's, Yanagisawa's, and Kuh's synthesis procedures are explained. The final chapter is devoted to noise in linear circuits.

There are five appendices, which usefully supplement the treatment in the text, and to a large extent make the book self-contained. Problems and a bibliography are given at the end of each chapter.

This is a useful well-written book and is highly recommended. It complements the standard approach to linear circuit design with information on the more modern methods. It will be suitable both as an introductory work for students, and as a reference, particularly on the more modern topics for practising engineers.

G. D. BERGMAN.

Fundamental Analogue Techniques

By R. J. A. Paul. 216 pp. Demy 8vo. Souvenir Press. 1965. Price 35s.

OVER the past few years this reviewer in these columns has criticized several books on analogue computing. A common fault has been the fact that interspersed between chapters on the main subject would be chapters on electronic theory and techniques not strictly essential to an understanding of analogue computing. An added distraction would be the fact that such chapters would be almost wholly in terms of hard-valve techniques and thus obsolescent if not obsolete.

The publishers of the book now reviewed have avoided this situation by making Professor Paul's book one of an 'Electronic User Series' thus allowing him to devote the entire book to fundamental analogue computing techniques. An Editor's Preface makes reference to two other volumes which cover electronic devices and circuit techniques. The result is a book attractive in text layout and eminently readable.

Chapter 1 is an introduction to the subject dealing with the historical background to computing aids. It includes one section on analogue techniques and for comparative purposes one on digital techniques. The chapter ends with a distinction between the analogue simulator and the differential analyser.

Professor Paul selects the latter as being more relevant to the needs of the average computing laboratory and Chapter 2 shows how the machine has been developed from the concept of Lord Kelvin in 1876. It is explained in terms of solutions for linear and non-linear differential equations and there is a discussion on coefficients and function generation.

Chapter 3 deals with programming, scaling and checking procedures. The emphasis is on dynamic systems and there are sections on determination of time scale factor and amplitude scale factor. Both are illustrated with worked examples. The last half of the chapter covers, first, symbolic notation for computing units the final program and some notes on preliminary runs.

Chapter 4 is a comprehensive treatment of dynamic analogies and network element transforms. Examples are given of a car suspension system and a suspension system with three degrees of freedom which tends to emphasize the use of the computer as a simulator, slightly at variance with the author's declared intentions of concentrating on the differential analyser.

Chapter 5 deals with some of the more complex modes of operation such as iterative operation, use of the integrator as a memory device, time-scale expansion, multiple integration and function generation. Each topic is illustrated by examples.

The final chapter deals with the simulation of rational transfer functions, two and three terminal networks and networks with the minimum number of components. These same topics are extended in three appendices dealing with Laplace transforms, generalized two-part networks and symmetrical lattice networks.

T. R. H. SIZER

Silicon Semiconductor Technology

By W. R. Runyan. 277 pp. Crown 4to. McGraw-Hill Book Co. 1965. Price 132s.

W. R. RUNYAN has written a book that will be a useful work of reference for anyone working on silicon technology. It contains a great deal of experimental data on a wide range of technologies such as material preparation, casting processes, crystal growth and diffusion, and also on the electrical, properties of silicon.

However, there are parts of the book where the emphasis on experimental data and useful formulae seems excessive in relation to the description of the mechanisms involved. For example, in the chapter on diffusion it is a pity that there is no real discussion of interstition versus substitutional diffusion nor of interactions between impurities. The impression that in some ways this is more a handbook than a textbook is strengthened by the occasional repetition (for example on segregation coefficients in the chapters on Crystal Growth and Doping Procedures) and by certain omissions of cross-references (such as to the table of energy levels of impurities when discussing the effects of doping on resistivity). This does not imply that the book suffers seriously in utility since it is entirely suitable for a reader with a grasp of semiconductor physics or for one who merely wants detailed information on a particular topic. Such readers will find this a valuable work, not the least for its comprehensive lists of references.

J. R. A. BEALE.

Electronic Components Tubes and Transistors

By G. W. A. Dummer. 166 pp. Crown 8vo. Pergamon Press. 1965. Price 21s.

The essential characteristics of each main type of component, tube and transistor are summarized in this book which is intended to bridge the gap between the basic measurement theory of resistance, capacitance and inductance and the practical application of electronic components.

Oscilloscope Measuring Technique
By J. Czech. 620 pp. Med. 8vo. Philips Technical Library. 1965. Price 105s.

This book is a revised and enlarged version of the author's book published originally in 1959 in German and deals with the theory and design of oscilloscope and methods of measurement.

Eighteen chapters are devoted to detailed descriptions of measuring techniques in different fields.

Electronics
A Bibliographical Guide—2
By C. K. Moore and K. J. Spencer. 369 pp. Crown 4to. Macdonald & Co. 1965. Price 85s.

Volume 2 of this guide lists the major works in electronics literature, text books, survey articles and papers presented at the major conventions and conferences during the period July 1959 to December 1964.

Volume 1 of this guide covered the period 1945 to June 1959.

Mechanising Laboratories
By E. A. Smith. 204 pp. Demy 8vo. Iliffe Books Ltd. 1965. Price 63s.

The purpose of this book is to survey the general range of mechanical aids and instruments available to enable the scientists or technologists to choose those most suitable for this purpose and to provide management with information on the best ways of mechanizing laboratories.

Transistor Pocket Book
By R. G. Hibberd. 304 pp. Crown 8vo. George Newnes Ltd. 1965. Price 25s.

This book is intended mainly for students and technicians and the treatment therefore is largely non-mathematical.

It describes the principles of operation of transistors, their characteristics, equivalent circuits and parameters.

A number of practical circuits is given, complete with transistor types and component values.

Practical Vacuum Techniques
By W. F. Brunner and T. H. Batzer. 198 pp. Med. 8vo. Chapman & Hall. 1965. Price 66s.

Although not primarily concerned with theory and design, this book contains enough theory to provide a basic understanding of vacuum systems to enable the technician to assemble, install, test and operate modern vacuum installations.

Analogue Computing Methods
By D. Welbourne. 140 pp. Crown 8vo. Pergamon Press. 1965. Price 17s. 6d.

This book is an attempt to present the field of analogue computation and simulation in a compact form. The reader is assumed to have the necessary knowledge of mathematics so that having read the elements described in this book and examples given, he can then proceed to the more comprehensive books on analogue computing.

Automatic Digital Calculators
By A. D. Booth and K. H. V. Booth. 263 pp. Demy 8vo. Chapman & Hall. 1965. Price 52s. 6d.

The third edition of this book, originally published in 1953, now includes information on transistor and tunnel diode circuits, thin film magnetic and cryogenic storage and modern auto-coding systems with particular reference to FORTRAN.

Reports on Progress in Physics
By A. C. Strickland. 518 pp. Crown 8vo. The Institute of Physics and the Physical Society. 1965.

The 28th volume of this series contains 13 separate articles covering various aspects of physics including photoconductivity electron beam spectroscopy and the thermomagnetic effects in semiconductors and semimetals.

Dictionary of Electrical Engineering
By K. G. Jackson. 373 pp. Crown 8vo. George Newnes Ltd. 1965. Price 45s.

Terms that are of general use to electrical engineers are included in this illustrated dictionary. Included also are terms used in electronics, lighting and constructional materials.

Circuits Using Direct Current Relays
By A. H. Bruinsma. 86 pp. Demy 8vo. Iliffe Books Ltd. 1965. Price 13s. 6d.

Intended for the technician, this book deals with a number of uncommon circuits which have proved practical for the application of d.c. relays in commanded or programmed sequence circuits.

Switch-in and switch-out phenomena are described, and the book shows how a number of problems of automatic or commanded sequence circuits, with either fixed or variable programmes, can be solved by the use of d.c. relays.

Elements of Theoretical Mechanics for Electronic Engineers
By F. Bultot. 260 pp. Med. 8vo. Pergamon Press. 1965. Price 45s.

The English edition of this book is intended for students at Colleges of Technology studying electronics. Subjects dealt with include vectors and vector systems, vibrations and waves, scalar and vector fields.

Linear Sequential Switching Circuits
By W. H. Kautz. 234 pp. Med. 8vo. Holden-Day Inc. 1965. Price $6.75.

This book consists of 10 papers previously published on the theory of linear sequential switching circuits.

Elektrische Messgeräte und Messverfahren
(Electrical Measuring Instruments and Methods)
By P. M. Pflier and H. Jahn. 317 pp. Royal 8vo. 3rd. edition, completely revised by Dr.-Ing. Hans Jahn. Springer Verlag. 1965. Price DM.45.-

The author aims at a sweeping survey rather than specialized knowledge and has the engineer in mind who has to keep up with rapidly changing techniques. Starting with the principles of measurements, the book deals with pointer indication and digital presentation, pre-amplifiers and transducers, bridges and potentiometric recorders, telemetry, insulation and earth measurements. It ends with a chapter on the electrical measurement of non-electric quantities.

Regelungstechnik für Praktiker Elementare Einführung in die Prozessregelung
(Control Engineering for the Practical Engineer Basic Introduction into Process Control)
By F. Piwinger. 132 pp. Demy 8vo. VDI-Verlag, Düsseldorf. 1965. Price DM.19.80

This book is based on lectures, courses and publications by the author and addresses itself clearly to the practical man who wishes to know something about process control without delving too deeply into the subject.

Die Anwendung der linearen Programmierung in Industriebetrieben
(The Application of Linear Programming in Industrial Plants)
By J. Vokuhl. 176 pp. Med. 8vo. Walter de Gruyter & Co., Berlin. 1965. Price DM.28.-

Considerable space is given to the introduction to linear programming, followed by chapters on decision making and linear programming, feasibility, practical application to specific problems, and the limits of application set by the mathematical model, economic conditions and the problem itself.

NEW EQUIPMENT

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

(Voir page 127 pour la traduction en français; Deutsche Übersetzung Seite 134)

OPERATIONAL AMPLIFIER

Tektronix UK Ltd, Beaverton House, Station Approach, Harpenden, Hertfordshire

(Illustrated below)

The type 3A8 wide-band operational amplifier plug-in unit was designed for use in any Tektronix 560-series oscilloscope that accepts 3-series plug-ins.

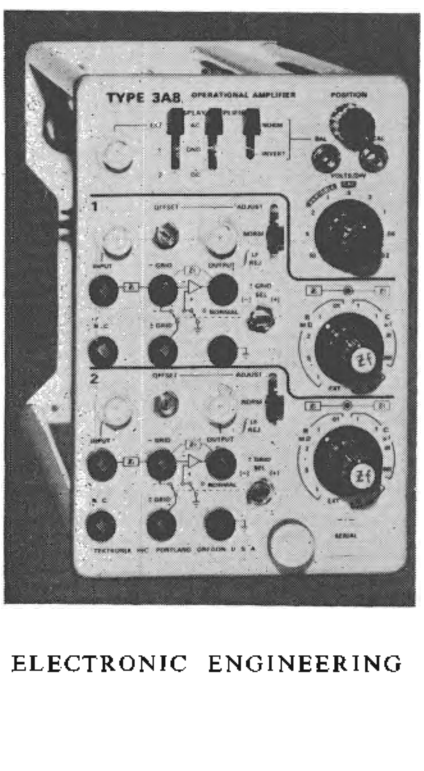
When used in a 560-series oscilliscope, or a type 129 power supply if a display is not required, the type 3A8 can perform precise operations of integration, differentiation, function generation, linear and non-linear amplification.

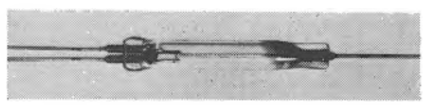
Each type 3A8 unit provides two versatile operational amplifiers and an oscilloscope deflexion amplifier. Gain-bandwidth product of each operational amplifier is 10Mc/s or higher and open-loop gain at d.c. is greater than 15 000. Front-panel controls provide wide-range selection of input and feedback components for each operational amplifier.

Pass-band of the vertical amplifier is d.c. to 3·5Mc/s; sensitivity is 20mV/div to 10V/div. Switching facilities permit the vertical amplifier to be used independently or to monitor the output of either operational amplifier.

Used with a storage oscilloscope, such as the Tektronix type 564, optimum utilization of the capabilities of the type 3A8 can be obtained. Typical applications are semiconductor leakage testing, where high junction capacitance sometimes requires sweeps in the 2sec/div range to avoid displacement-current effects, and in single-event studies where photography was formerly used.

EE 90 751 *for further details*





DRY REED INSERTS

Hivac Ltd, Stonefield Way, South Ruislip, Middlesex

(Illustrated above)

A new group of dry reed inserts with changeover contacts is available from Hivac Ltd. Similar in general construction to other Hivac inserts, the new changeover components have a diameter of 0·2in and a glass length of 1·375in.

Changeover inserts are available with two types of contact plating: partially diffused gold, designated XS11; and a fully diffused gold, designated XS13. In adidtion these types are each available in two separate sensitivity ranges, XS11/1 and XS11/2, and XS13/1 and XS13/2.

Typical operating times of an insert are less than 0·5msec for the normally closed contacts and less than 2·5msec for the normally open pair. Both contact pairs present less than 90mΩ closed resistance, and the insert has a breakdown voltage greater than 300V d.c. Maximum recommended contact ratings for long life are 200mA at 50V d.c.

EE 90 752 *for further details*

DIGITAL CHRONOMETER

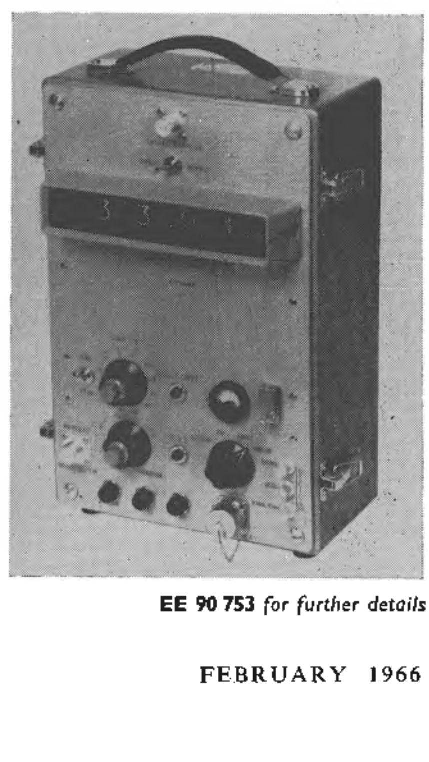
Distributed by: Britec Ltd, 17 Charing Cross Road, London, W.C.2

(Illustrated on right)

The 'SERCEL Chronoseis' is a fully-transistorized deci-millisecond chronometer with digital read-out. Designed primarily for measurements in the field, it is contained in a portable duraluminium case, 16in by $10\frac{1}{2}$in by $7\frac{3}{4}$in, with detachable lid, and has an internal nickel-cadmium battery which provides eight hours of continuous operation; recharging can be effected directly from a.c. mains or from an external 6V accumulator.

The Chronoseis measures the time interval between two control pulses, within the range of 0·1 to 199·9msec with an accuracy of ±0·1msec from −10°C to +60°C. It incorporates a high sensitivity amplifier with adjustable gain making it possible to work with pulses ranging in amplitude from a few microvolts to several volts. A built-in time delay, graduated in 10msec steps, prevents premature triggering of the control circuit by parasitic interference or unwanted signals. Coaxial outlets are provided on the control panel enabling the control pulses and a 10msec marker to be fed to an oscilloscope, if desired.

The main application is for seismic measurements. When used in seismic prospecting, the Chronoseis measures the time elapsed between the instant of impact, of a hammer or detonator, on the ground to be explored and arrival of the vibration signal detected by a geophone placed at a distance from the point of impact. Using simple formulae, it is then easy to determine the depths and characteristics of various strata (sand, clay, chalk, granite, etc.) beneath the survey line, down to about 165ft.



EE 90 753 *for further details*

GALVANOMETER

Airmec Ltd, High Wycombe, Buckinghamshire

(Illustrated below)

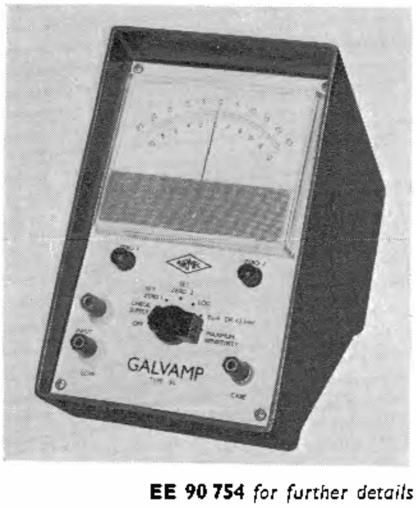
The Airmec Galvamp type 391 is a new solid-state galvanometer of high sensitivity designed to withstand hard treatment both electrically and mechanically.

It consists essentially of a transistorized d.c. chopper-type amplifier feeding a large centre-zero meter. Excellent stability is achieved by the use of negative feedback, and a special circuit is employed to ensure that the zero is independent of the source impedance of the input signal. The amplifier operates from an internal 9V battery or from an external 7V to 9V d.c. supply.

The Galvamp has a sensitivity of 2μV/mm, but overload voltages up to 20V can be tolerated without damage, and it can be used in conditions of vibration and shock that would render a conventional galvanometer inoperative.

The instrument has three overlapping voltage ranges, the highest being a 1V-0-1V logarithmic range which when used for bridge detection enables approximate balance to be achieved easily. Final balance to a high degree of accuracy can then be obtained by switching to the other two linear ranges (2·5mV and 100μV f.s.d.) in turn. This feature together with the fast response time of the meter movement allows measurements to be made very quickly.

The instrument is contained in a small cast case with the meter panel set at the most convenient angle for use on a bench. Alternatively, with the case removed the Galvamp can, if desired, be mounted through a suitable aperture in the front panel of another equipment.



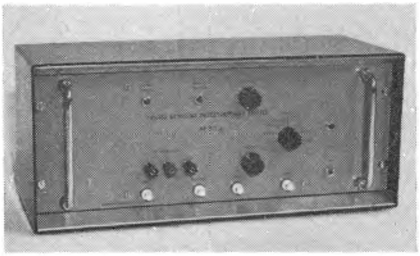
EE 90 754 *for further details*

PHASE SENSITIVE DETECTOR AND PHASE SHIFTER

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

(Illustrated above right)

The Brookdeal model PP313A combines in a single unit a phase-sensitive detector and phase shifter using high frequency silicon planar transistors. New design techniques give both sections a



frequency response up to 300kc/s and have also resulted in setting up operations being reduced to frequency range selection and coarse adjustment of reference voltage amplitude.

The phase-sensitive detector features very low d.c. drift and a high degree of linearity. The low value of the d.c. drift allows the instrument to be used at very low signal levels thus reducing the risk of overload by noise signals and therefore in most applications the signal amplifier which is used in conjunction with the PP313A need not be tuned. Overload indication is given by a neon indicator on the front panel and a similar indicator shows the correct adjustment of the reference voltage amplitude.

The output from the phase-sensitive detector is suitable for driving a pen-recorder or external meter; the terminals provided allow simple connexion of capacitors on the output so that the required time-constant may be obtained.

The phase shifter is designed to provide phase correction of signals applied to the reference channel of the phase-sensitive detector, 180° phase control being available. For a given input voltage the output of the phase shifter does not vary by more than 1dB over the whole phase range and this combined with the reference channel limiter removes the need for checking the phase detector output zeroing after re-setting the phase.

The unit can be supplied for mounting in a 19in rack or in a bench case. It is also suitable for mounting with any of the Brookdeal range of amplifiers in a dual unit bench case.

EE 90 755 *for further details*

PROGRAMMABLE PULSE GENERATOR

Tektronix UK Ltd, Beaverton House, Station Approach, Harpenden, Hertfordshire

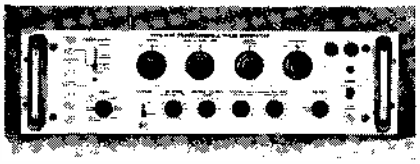
(Illustrated above right)

The type R116 10Mc/s, 10V programmable pulse generator, is primarily intended for applications where a variety of pulse amplitudes, polarities, shapes and other parameters are required in rapid sequence, as in systems and production testing. All functions are programmable and, in addition, the type R116 can be operated from calibrated front-panel controls for applications which do not require external programming.

Calibrated and programmable parameters include trigger source, period, delay or burst time, width, amplitude, mode, d.c.-offset, rise-time, fall-time, and polarity. Switch-range as well as variable control is programmable for all functions.

Rise-time and fall-time ranges from 10nsec minimum to 110μsec maximum, amplitude from 0·4V to 10V, width from 50nsec to 550μsec, delay or burst time from 50nsec to 550μsec, period from 100nsec to 11msec, and d.c. offset from plus 5 to minus 5V—all continuously variable. Rise-time and fall-time are independently variable when on the same range.

Special programmable operating modes include delayed single pulse that provides normal pulse output after the selected delay time, double pulse output for two pulses per pre-trigger output—with one pulse delayed in time by a variable amount with respect to the other, burst output with a burst of output pulses for each applied external trigger, and gated output that provides pulse outputs for the duration of the applied gate.



Complete programming capability requires 21 bits and 7 analogue lines. Characteristics of the type R116 make it suitable for semiconductor and general-purpose applications.

The type R116 mounts in a standard 19in rack and is only 4½in high and 18½in deep.

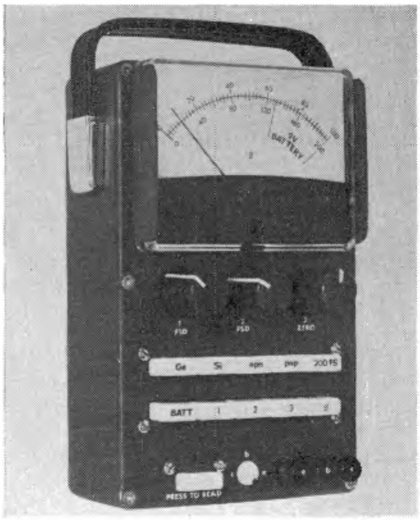
EE 90 756 *for further details*

TRANSISTOR TEST SET

Standard Telephones & Cables Ltd, STC House, 190 Strand, London, W.C.2

(Illustrated below)

Portability and robustness, coupled with simplicity of operation, are features of the 74163-B transistor test set recently introduced by Standard Telephones and Cables Ltd. It will measure the dynamic current gain of both silicon and germanium transistors of the pnp and npn types. Gains up to 200 may be measured on a dual-range meter, with an emitter current of approximately 1mA and a



collector voltage of approximately 5V. The meter is calibrated directly in current gain.

Two dry cells are housed in the light-alloy case which has no lid, but has side flanges to protect the meter. Push-button switching facilitates setting-up the appropriate circuit arrangements, and the voltages of the two dry cells can be checked on the meter. A three-pin socket and three terminals are fitted, so that the transistor to be tested can be connected to whichever is the more convenient.

The overall dimensions including the handle are 7¾in × 5¼in × 3⅝in, and the weight is 3¼ lb.

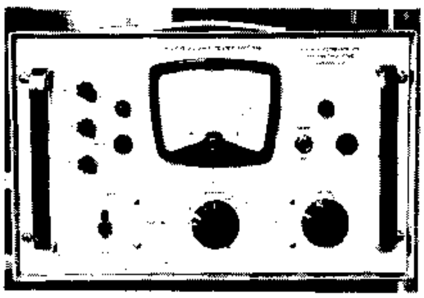
EE 90 757 *for further details*

HIGH RESISTANCE TESTER

K.S.M. Electronics Ltd, 139-149 Fonthill Road, Finsbury Park, London, N.4

(Illustrated below)

The model HR1 high resistance meter is a portable instrument designed to measure high resistance in the range



50kΩ to 200 000MΩ and is particularly useful for measuring insulation resistance. The voltages available are 100V and 500V. The 100V range is used to measure resistance in the range 50kΩ to 20 000MΩ and the 500V range measures resistance from 500kΩ to 200 000MΩ. The instrument is also calibrated in conductance from 20 microhos down to 5 picomhos.

The two supply voltages are stabilized to enable stable resistance measurements to be made. The instrument is also designed to measure the resistance of highly capacitive components such as capacitors or long lengths of cable.

EE 90 758 *for further details*

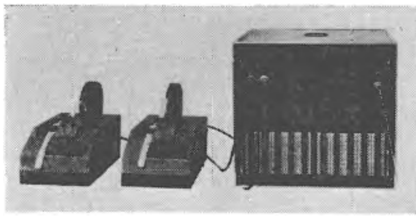
DATA LOGGER

Deakin Phillips Electronics Ltd, Tilly's Lane, High Street, Staines, Middlesex

(Illustrated above right)

Developed under contract to the Building Research Station, a digital logging device is now being marketed by the designers, Deakin Phillips Electronics Ltd, for the economical logging of events from a number of sources. The information record of each source identifies the event by its allotted channel number, together with the duration of its 'on' time, the instant in real time at which it occurs and its sequence of occurrance in relation to the other sources being recorded.

The device described has a maximum



channel capacity of 127 sources and records the start and finish of an event as two identifiable signals on a punched tape. A time marker is also punched on the tape so that when read through suitable measuring equipment, the duration of the event, and the identification of the source, can both be reconstituted.

The source of the event is coded in simple binary numbers and to distinguish between its start and finish, the appropriate binary code has an extra hole punched in a unique position to identify the 'finish' event. The time record is designed to punch a further unique code and in the equipment supplied to the Building Research Station, an interval of 15sec is used, although other time intervals can be arranged. Although the system described is designed for recording events in relation to time, it will be appreciated that any parameters can be recorded with respect to one another, allowing correlation records to be made.

To eliminate the possible loss of record which may arise when the start or stop of an event occurs during the interval in which the time marker is being punched, an arrangement of interlocking buffer stores is used in the system. These stores also ensure that there is no complementary loss of time record when an event start or finish is being punched.

EE 90 759 *for further details*

CONTAMINATION MONITOR

EMI Electronics Ltd, Hayes, Middlesex

(Illustrated below)

A compact and versatile new radiation contamination monitor, the PCM 3, has been introduced by EMI Electronics Ltd.

The new monitor, which weighs only 6 lb, can be used with suitable detector probes for monitoring all kinds of radiation. It is fully transistorized and powered by its own internal battery.

The PCM 3, primarily designed for use with EMI dual phosphor probes, types DP2 and DP2R, permits simul-

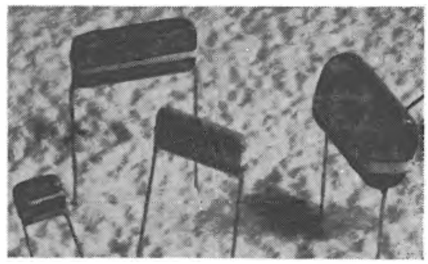
taneous monitoring of alpha and beta/gamma contamination. It can also be used with other EMI nuclear radiation probes as well as probes made by other manufacturers for alpha, beta, gamma and X-ray monitoring.

The monitor provides visual and audible indication of radiation levels. The visual indication is on a highly accurate 5in longscale meter; the audible indication is emitted by an internal loudspeaker with separate and distinctive tones for alpha and beta/gamma radiation. Provision is also made for connexion of an external relay to operate an alarm at a pre-set radiation level.

A 'fail-safe' circuit is incorporated so that if a probe is damaged or subjected to excessively heavy radiation a continuous high-pitched warning note is emitted from the speaker.

The instrument's transistor circuits are mounted on printed boards for maximum reliability. The meter and all controls are conveniently grouped on the top face in front of the handle.

EE 90 760 *for further details*



POLYESTER CAPACITORS

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

(Illustrated above)

The Mullard C280 series of foil capacitors for use in radio and television applications now covers a capacitance range from 0·01μF to 2·2μF in 15 preferred values, at a d.c. working voltage of 250V. Previously the upper limit of capacitance was 0·22μF and the d.c. working voltage 30V.

This increase in working voltage enables the capacitors to be used in valve circuits operating from h.t. lines of up to 250V. If necessary the working voltage can be exceeded to accommodate the normal increase in h.t. voltage which occurs during the valve's heating-up period.

Because of their small physical size, rectangular shape and protective coating the new series of capacitors, designated C280AE, gives maximum space utilization when used on printed wiring boards.

EE 90 761 *for further details*

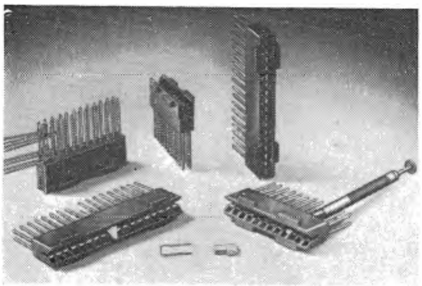
WIRE WRAP EDGE CONNECTORS

Ultra Electronics (Components) Ltd, Western Avenue, London, W.3

(Illustrated on page 123)

Two new wire wrap edge connectors have been added to the large range of miniature, sub-miniature and micro-miniature connectors produced by Ultra Electronics (Components) Ltd.

The two connectors 12 + 12 and



18 + 18 way have a contact pitch of 0·2in and terminations of 0·046in². The contact form is of bellows action, and the contact/termination assemblies have the added facility of being removable.

The connectors are open-ended, but perfect alignment of the printed wiring board is catered for by the use of a friction fit Nylon reference key.

The connectors have the additional advantage of being able to be stacked on the building brick principle, giving any number of ways required in multiples of 12 or 18.

EE 90 962 *for further details*

RANDOM NOISE CONTROL EQUIPMENT

Derritron Ltd, 24 Upper Brook Street, London, W.1

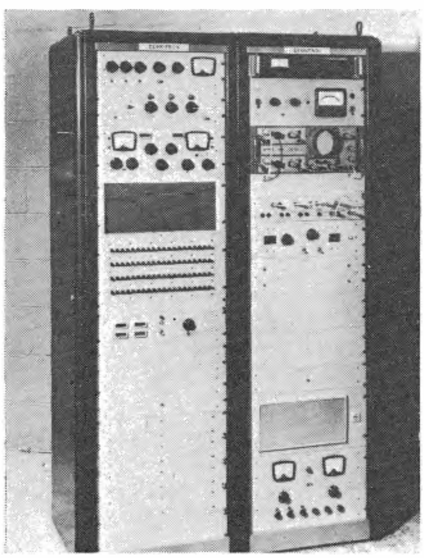
(Illustrated below)

The Derritron automatic random noise control consoles have been designed to satisfy the need for British equipment of this type to meet the growing demand both in the United Kingdom, Europe and elsewhere.

The advanced circuits feature all solid state design and plug-in printed boards, resulting in a very low temperature rise inside the cabinet, ensuring a high degree of reliability and stability.

There are three basic consoles in the Derritron range:—

Type ARN.1 40-channels
Frequency range 5 to 2 000c/s
Type ARN.2 80-channels
Frequency range 5 to 20 000c/s
Type ARN.3 160-channels
Frequency range 5 to 5 000c/s



A panel switch control immediately changes the range of the ARN.1 and ARN.2 to 2kc/s to 4kc/s, 4kc/s to 6kc/s, 6kc/s to 8kc/s, and 8kc/s to 10kc/s, and the ARN.3 to 5kc/s to 10kc/s.

A brief specification of ARN.2 is given below. The other consoles conform in general to this specification.

Number of analyser/equalizer channels: 80.
Frequency range of operation: 5c/s to 2 000c/s or 2-4kc/s, 4-6kc/s, 6-8kc/s, 8-10kc/s by operation of front panel switch.
Dynamic range (auto or manual): 50dB.
Operating modes: Automatic and manual by operation of front panel switches.
Type of filters: Magnetostrictive.
Filter operating frequency: 100kc/s nominal.
Hum and noise: 60dB below full output.
Programming: By means of front panel sliders —one slider per channel, i.e. 80 total.
Programming range: 50dB.
Ripple factor: ±1·5dB.
Filter bandwidths: as below.

Filter Bandwidth	*Frequency Range*	*Number of Filters*
12·5c/s	0-400c/s	32
25c/s	400-1 200c/s	32
50c/s	1 200-2 000c/s	16

Each console is supplied complete with a random noise generator, oscilloscope, true r.m.s. valve-voltmeter/spectral density meter, displacement, velocity and acceleration limiters, emitter-follower, accelerometer input and all controls necessary for the operation of the equipment. An *XY* plotter and logarithmic convertor can be supplied as an optional extra.

The photograph illustrates the Derritron ARN.2 80-channel console, as delivered to a British Government Research Establishment.

Desk consoles can also be supplied complete with all controls and instruments for remote operation of complete vibrator systems.

EE 90 963 *for further details*

MINIATURE MECHANICAL FILTERS

Telefunken A.G., Ernst Reuter Platz 7, 1 Berlin 10, Germany

Miniature mechanical filters, rated for 455kc/s and measuring 22·4 × 11·7 × 4mm, have been designed specifically for s.s.b. equipment. Using torsional resonators particularly steep response curves are obtained so that the 60dB values of the filters designed for bandwidths (3dB) of 1·5kc/s, 3kc/s and 6kc/s are only 4kc/s, 6·5kc/s and 40kc/s. The ripple is below 3dB in all cases, and is only 1·5 to 2·5 dB for the 6kc/s filters. In consequence this filter is very suitable for v.h.f. radiotelephones. The external resonant circuit capacitances are incorporated in the filter, and the resonant circuit coils must be connected externally. Accurate matching is thus possible, especially in transistor circuits.

EE 90 964 *for further details*

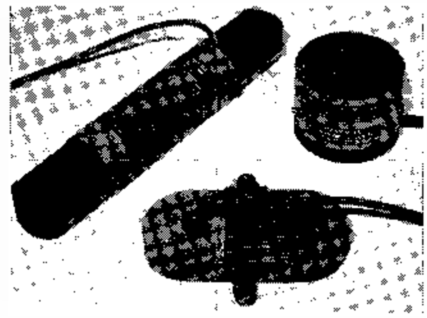
LOAD CELLS

Westland Aircraft Ltd, Saunders-Roe Division, Osborne Works, East Cowes, Isle of Wight

(Illustrated above right)

The Saunders-Roe Division of Westland Aircraft Ltd has announced the completion of a new range of load cells.

All of these utilize the well known Saunders-Roe bonded foil strain gauges



in a full bridge configuration, giving good accuracy and reliability in a compact, rugged and inexpensive assembly.

Three different types cover both tensile and compressive loads in the ranges 0 to 25 lb up to 0 to 60 tons. Full scale outputs of between 20 and 30mV are delivered from a bridge resistance of 250Ω energized at 10V supply.

For weighing installations where the load is supported on more than one point, cells may be supplied in half bridge form to enable the interconnexion of two, four or more cells in one composite circuit to obtain accurate readings of total or average loads.

EE 90 965 *for further details*

ULTRASONIC CLEANERS

Radyne Ltd, Wokingham, Berkshire

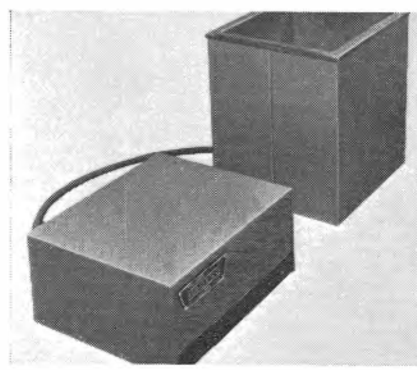
(Illustrated below)

A range of ultrasonic cleaners is now available and combines advanced circuit design with new developments in transducer techniques.

The generator is based on a silicon controlled rectifier used as an invertor and automatic frequency tracking eliminates the necessity for frequency adjustment. The circuit simplicity reduces the number of components to less than 50 per cent of those required for conventional generators.

Self-contained units are produced to provide 300/600W, 600/1200W and 1kW c.w. outputs.

The cleaning tanks are of stainless steel, shrouded and heat insulated. A new design of magnetostrictive transducer provides a mechanically robust source of ultrasonic power that covers the whole of the tank bottom. Completely insulated energizing coils operatng from low voltage eliminates electrical breakdown due to moisture. Three standard sizes of tanks are available with 3½ gallons, 7½ gallons and 20 gallons capa-



city. A feature is the ability to use any tank size with all generators, giving various power densities.

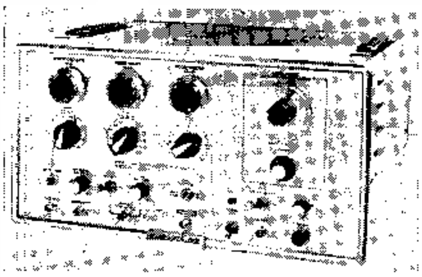
EE 90 766 *for further details*

PRECISION PULSE GENERATOR

Dynamco Instruments Ltd, Salisbury Grove, Mytchett, Aldershot, Hampshire

(Illustrated below)

Pulse delay, width and repetition rate accuracies better than 1 per cent over 0 to 50°C ambient, direct reading ampli-



tude control, solid state circuits and wide range general purpose capabilities are features of the 'Datapulse' 107 precision pulse generator.

Decade switching and 10-turn vernier dials allow resolution better than 1 part in 10^3. The direct reading amplitude control speeds set-up and ensures repeatable tests. The temperature compensated, solid state circuits are highly reliable and pulse waveform is exceptionally clean. The 107 offers both single or double pulse operation, repetition rates from 0·5c/s to 1Mc/s, 50V output into 50Ω, rise times of less than 10nsec (with degradation control), pulse delays to 1sec and pulse durations from 50nsec to 10msec.

Direct reading, high resolution controls and drift-free operation make the 107 a suitable pulse generator for production test and quality control applications. An oscilloscope is not required for test set up or stability monitoring.

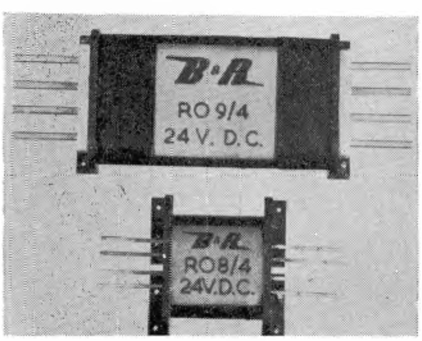
EE 90 767 *for further details*

DRY REED RELAYS

B. & R. Relays Ltd, Temple Fields, Harlow, Essex

(Illustrated below)

B. & R. Relays Ltd has introduced two new ranges of dry reed relays to be called the RO8 and RO9, providing a low cost comprehensive range of switching devices for a wide range of applications. Both have a basic construction of a coil wound on a phenolic bobbin into



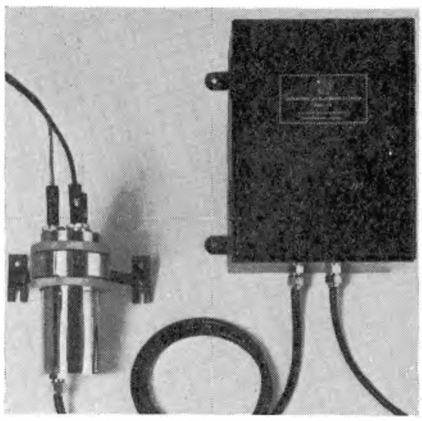
which the reeds are fitted. The reed switches are retained within the bobbin by push fit inserts. This method of consruction permits the reed switches to be easily removed.

The RO8 range comprises four sizes of dry reed relays designed to incorporate one, two, three or four miniature dry reed switches. The MR 300 and MR 306, normally open switches, and the MR 203 changeover switch may be fitted to the RO8 range.

The RO9 range is also available in four versions suitable for one, two, three or four dry reed switches of the standard size. The MR 706, normally open, MR 840 high load molybdenum contact reed and the MR 201 changeover switch are recommended for use with the RO9.

Standard coils are available of both types for 3, 6, 12, 18 and 24V d.c.

EE 90 768 *for further details*



ULTRASONIC ELECTRODE CLEANER

Electronic Instruments Ltd, Lower Mortlake Road, Richmond, Surrey

(Illustrated above)

Electronic Instruments Ltd has produced a new unit based on the latest ultrasonic cleaning techniques as an optional attachment to E.I.L. Industrial pH electrode systems. Adaptor kits are available to enable the cleaning unit to be fitted to different pH and redox electrode systems.

The main advantage of this system is that it can overcome difficult and stubborn electrode coating problems and give trouble-free performance with a negligible increase in running costs. An installation in a paper mill for example enabled the electrode maintenance interval to be extended from two hours to eight weeks.

EE 90 769 *for further details*

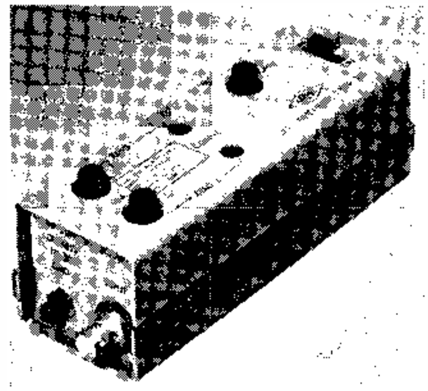
D.C. AMPLIFIER

Airmec Ltd, High Wycombe, Buckinghamshire

(Illustrated above right)

Small, fully-transistorized, and powered by two internal mercury cells with a normal life of 250 hours, this amplifier accepts inputs in the frequency range d.c. to 200kc/s and provides steplessly variable amplification from ×10 to ×100.

It is particularly useful as a pre-



amplifier in oscilloscope work and is fully compatible with the Airmec display oscilloscope type 279.

Preset zero controls are arranged to prevent unwanted currents being fed back into the source and ensure that the output zero is unaffected by changes in source resistance.

Input resistance at 1kc/s is greater than 400kΩ. Output resistance is less than 1kΩ at the same frequency, and maximum output level is at least ±1V peak at all frequencies into loads greater than 50kΩ. The amplifier will withstand input overloads up to ±30V peak without damage

EE 90 770 *for further details*

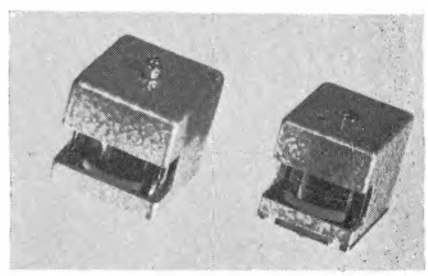
FERRITE RECORDING HEADS

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

(Illustrated below)

A range of ferrite magnetic heads for professional audio recording equipment is now available from Mullard Ltd.

Unlike laminated-metal heads the electrical properties of ferrite record and replay heads remain unchanged throughout their long working life. Thus amplifier circuits associated with ferrite heads do not need the frequent compen-



satory adjustments which have to be made throughout the much shorter life of a metal head. Tests showed that after 300 hours running time at a tape speed of 7½in/sec the output of a ferrite head measured at 8kc/s was reduced by only 1·5dB and remained substantially constant after a further 5 000 hours.

Because ferrite heads have a lower h.f. core loss and a lower eddy current loss than metal heads much lower bias currents can be used. This enables a lower drive current to be used so simplifying recorder circuits and reducing head noise. A further useful reduction in drive current can be obtained by increasing the bias frequency—a technique which may only give a marginal advantage with metal heads.

The frequency response of a ferrite head is generally much better than that of a metal head because the effective electrical gap length corresponds closely to the physical gap length. This is due to the stress-free nature of ferrite which allows it to be machined accurately during manufacture without altering its electrical characteristics—an advantage not shared by metal heads.

Both record and replay heads are being produced. Half, full, stereo-twin and two types of quarter-track twin heads are available. Replay heads have 3μm gaps. Record heads are available with 25μm (full track only), 12μm or 7μm gaps. All are colour coded.

The heads can be supplied as either fixed or adjustable assemblies. The adjustable heads are housed in a Mumetal screened assembly. Facilities are provided for pre-set adjustment of 'height' and 'tilt', and for final azimuth adjustment with the Mumetal screen in position. A magnetic or non-magnetic outer screen can also be supplied. With the fixed versions, shims are available to adjust the height of the head; mounting adaptors can also be supplied. Single and double Mumetal screening cans are available.

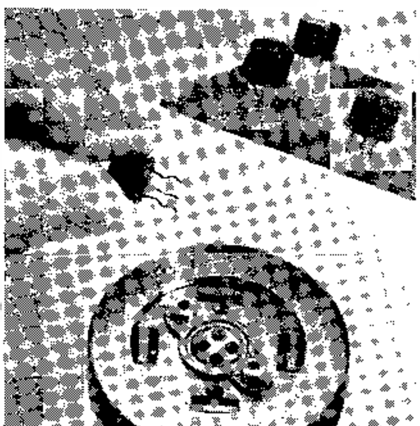
EE 90 771 *for further details*

LEAD CUTTING AND JOGGLING DIE

Kingham Electronics Ltd, 17 Briary Wood Lane, Welwyn Heath, Welwyn, Hertfordshire

(Illustrated below)

This new transistor lead cutting and joggling die eliminates the need for transistor pads between transistor and printed circuit board. The die cuts three or four leads to predetermined lengths and joggles in a single operation. The transistor can be snapped into standard hole patterns and remains in place with $^1/_8$ to $^5/_{32}$in air gap, regardless of printed circuit board position. It permits easy cleaning of the board without entrapment problems normally associated with pads. Transistor leads can be hand or wave soldered without the necessity of the extra operation of trimming after soldering. Dies are also available for: trimming to length only and for trimming and joggling of cut-of end only thus permitting use of pads. The die head is interchangeable with all crimping die heads used with Pico model 300, 300B or 300BT power units.



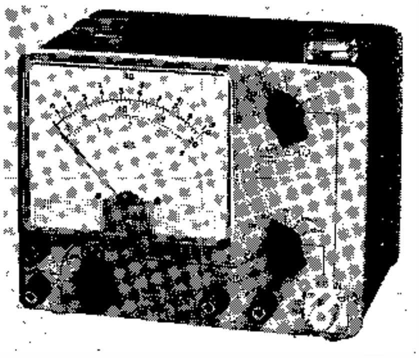
EE 90 772 *for further details*

WIDE RANGE VOLTMETER

Linstead Electronics Ltd, 35c Newington Green, London, N.16

(Illustrated below)

This transistorized instrument measures a.c. signals on fifteen ranges with full scale deflexions from 10mV to 500V. The input impedance is 10MΩ except on the three most sensitive ranges where it is between 6 and 10MΩ. The frequency range is 10c/s to 100kc/s. D.C. voltages can be measured on three ranges of 0 to 4V, 0 to 40V and 0 to 400V at an impedance of 20kΩ/V. The instrument may also be used as an amplifier with a low



output impedance and a gain of about 80 on the most sensitive range.

The instrument is operated from an internal 9V battery which has a large capacity compared with the current drain. The indicating meter is of the 4in clear-fronted type.

EE 90 773 *for further details*

DEWPOINT METER

Shaw Moisture Meters, Rawson Road, Westgate, Bradford

(Illustrated above right)

The 'Lamplighter' for which a provisional patent has been granted, is a simple automatic instrument to indicate dewpoint temperatures in the 0°C to 100°C range.

A special electric lamp, fitted with a gap over which conduction takes place when condensation forms, is the basis of this new instrument. A thermistor indicates on a dial the temperature of the lamp, which lights in the presence of condensation, thus heating until this disappears. The reading of dewpoint takes only a few seconds and no skill or manual adjustment is required. In common with several recent instruments for moisture measurement, announced by this company, the "LAMPLIGHTER" has no controls.

The method gives absolute readings;

also the dewpoint meter is portable and does not contain any thermionic valves.

EE 90 774 *for further details*

SUB-MILLIMETRE WAVE MASER

G. & E. Bradley Ltd, Electrical House, Neasden Lane, London, N.W.10

The 'Teratron' is a source of pulsed, coherent power at a wavelength of 0·337 mm which has been developed by G. & E. Bradley Ltd in collaboration with the National Physical Laboratory. It provides peak output powers up to several watts in this remote part of the spectrum. This is already proving to be of use in far infra-red spectoscopy, transmission and dispersion measurements, and shows potential in the easy demonstration of wave optics in education as well as for distance measurement and for communications.

The system consists of a 4ft long vacuum tube with plane or confocal metal mirrors mounted at its end, one mirror being adjustable to allow change of cavity length.

The active material may be one of a number of organic gases or vapours containing the CN radical, such as methyl cyanide or propyl cyanide. The vapour is leaked into one end of the vacuum tube where a discharge of approximately 10kV dissociates the molecules. This produces an inversion of populations in the CN transition concerned, necessary for stimulated emission of radiation. Maser action then occurs along the length of the tube in the usual way, while waste products are pumped away continuously from the tube and may be disposed of safely.

Output is obtained through a Melinex window at one end of the tube, either by diffraction past one mirror or by beam splitting within the cavity.

The CN emission line is situated in one of the known 'windows' in the absorption curve of water vapour in the atmosphere, so there are numerous possibilities for using the Teratron for transmission over distances, even through fog. This would overcome one of the main objections to the use of optical masers for communication.

EE 90 775 *for further details*

Meetings THIS Month

THE INSTITUTION OF ELECTRICAL ENGINEERS

All meetings will be held at Savoy Place, commencing at 5.30 p.m. unless otherwise stated.

Ordinary Meetings

Date: 3 February.
Lecture: Electromagnetic Levitation.
By: E. R. Laithwaite.
Date: 17 February.
Discussion: The "Literature Explosion" what can be done about it?
Opened by: J. A. Ratcliffe.
Date: 22 February.
Discussion: Research in Electrical and Electronic Engineering.
(Joint with the I.E.R.E.)

Control and Automation

Date: 1 February.
Lecture: Pattern Processing in the Human Visual System and Some Engineering Implications.
By: C. R. Evans and J. R. Parks.
Date: 9 February.
Lecture: The Manchester Digital Traffic Simulator.
By: F. G. Heath, M. G. Hartley, D. Green and S. Saleeb.
(Joint meeting with the I.E.R.E.)
Date: 9 February.
Discussion: Computer Service Training.
Opened by: G. M. Bolton and S. E. Williamson.
Date: 15 February.
Lecture: A Control Engineering Approach to Magnetohydrodynamic Stability.
By: W. H. Cooper, A. B. Gillespie, E. E. Aldridge and J. Contos.
Date: 18 February.
Discussion: The Production and Properties of Epitaxial Films.
Opened by: B. A. Joyce.
Date: 22 February.
Lecture: Muscular Control.
By: P. A. Merton.
(Joint meeting with the I.E.R.E.)

Electronics

Date: 1 February.
Discussion: Common-mode Problems.
Opened by: R. B. D. Knight.
Date: 2 February.
Lecture: Modern Applications of Closed-circuit Television.
By: V. J. Cooper.
(Joint meeting with the I.E.R.E., the Television Society and the I.E.E.T.E.)
Date: 4 February.
Discussion: The Physics of Circuit Breaking.
Opened by: D. T. Swift-Hook and H. Edels.
Date: 16 February.
Lecture: Post Office Towers and Trunks.
By: J. H. H. Merriman.
Date: 21 February.
Conference: Delay Devices for Pulse Compression.
(All wishing to attend must register: forms available from the Secretary, I.E.E.)
Date: 25 February.
Lecture: Distributed Network Synthesis.
By: J. O. Scanlan and R. Levy.

Power

Date: 2 February.
Discussion: Aspects of Power-station Layout that Facilitate Control of Operation.
Opened by: C. E. Pugh.
Date: 8 February.
Lecture: The Origins of Load Losses in Induction Motors with Cast Aluminium Rotors.
By: N. Christofides.
Date: 28 February.
Lecture: Induced Voltages in the Sheaths of Cross-bonded A.C. Cables.
By: D. J. Rhodes and A. Wright.

INSTITUTION OF ELECTRONIC AND RADIO ENGINEERS

Electro Acoustics Group

Held at: I.E.R.E. Lecture Room, 9 Bedford Square, London W.C.1.
Date: 9 February. Time: 6 p.m.
Lecture: Trutrack—A New Approach to Low Side Thrust Transcription Arms.
By: A. R. Rangabe.
Held at: I.E.R.E. Lecture Room, 9 Bedford Square, London W.C.1.
Date: 2 February. Time: 6 p.m.
Lecture: The Electron Beam in Electronics Research and Production.
By: M. A. Champney.

Joint I.E.E.-I.E.R.E. Computer Groups

Held at: Institution of Electrical Engineers, Savoy Place, London W.C.2.
Date: 9 February. Time: 5.30 p.m.
Lecture: The Manchester Digital Traffic Simulator.
By: F. G. Heath, M. G. Hartley, D. Green and S. Saleeb.

Joint I.E.E.-I.E.R.E. Medical Electronics

Held at: Institution of Electrical Engineers, Savoy Place, London W.C.2.
Date: 22 February. Time: 5.30 p.m.
Lecture: Muscular Control.
By: P. A. Marton.
Held at: I.E.R.E. Room, 9 Bedford Square, London W.C.1.
Date: 23 February. Time: 6 p.m.
Discussion: Mental Stresses Induced by Audio and Visual Hazards.
Opened by: B. Shackel.

Southern Section

Held at: Basingstoke Technical College.
Date: 3 February. Time: 7.30 p.m.
Lecture: Instrumentation in Forest Soil Research.
By: R. Kitching.

Yorkshire Section

Held at: Department of Electrical Engineering, University of Sheffield.
Date: 3 February. Time: 7 p.m.
Lecture: Semiconductors in Industrial Instrumentation.
By: F. W. Addams.

THE INSTITUTION OF POST OFFICE ELECTRICAL ENGINEERS

Held at: Conference Room, Fleet Building, 70 Farringdon Street, E.C.4.
Date: 1 February. Time: 5 p.m.
Lecture: Provision of Service to Subscribers.
By: E. Croft.

THE TELEVISION SOCIETY

Meetings are held in the Conference Suite, I.T.A., 70 Brompton Road, London, S.W.3, and commence at 7 p.m.

Date: 10 February.
Lecture: The Independent Television News Service.
By: G. Cox.
Date: 25 February.
Lecture: The Use of Test Line Signals For Remote Monitoring of Picture Quality at Unattended Transmitters.
By: D. J. Parkyn.

Publications Received

THORN SIGNAL LAMPS AND LAMPHOLDERS is a new coloured and fully illustrated catalogue which gives details and dimensions of a large range of signal lamps and lampholders manufactured by Thorn Special Products Ltd. It includes lamps and lampholders for a.c. mains use, SGF light duty signal lamps for a.c. or d.c. use and 50 to 200 pulse/min blinker lamps and neon indicators.

Copies of this catalogue are available from Thorn Special Products Ltd, Great Cambridge Road, Enfield, Middlesex.

THE GAS LASER AND ITS EXPERIMENTAL APPLICATIONS is a 31 page booklet written to illustrate some of the many experiments which can be carried out with the Scientifica gas laser. Some of the experiments serve as a simple illustration of the properties of the laser while others are standard experiments concerned with basic optical phenomena. Instructions are also given for the alignment and operation of laser together with the precautions which are necessary when handling the instrument.

Copies of this booklet are available, price 12s. 6d. each from, Scientifica, 148, St. Dunstan's Avenue, Acton, London, W.3.

DRIVER AND CONTROLLER UNITS. Technical Publication, 26-23. is intended to help thyristor users by describing a range of Drivers for supplying gate firing pulses of suitable polarity and amplitude. Controllers for sampling the output and controlling it to a required value and also a range of supply transformers and mounting accessories.

Copies of this publication may be obtained from the Westinghouse Brake and Signal Co. Ltd., 82 York Way, Kings Cross, London, N.1.

CATALOGUE No. 136 is dated January 1966 and gives details of the extensive range of switches, neon indicators and signal lampholders manufactured by Arcolectric Switches Ltd. Copies of this catalogue are available from Arcolectric Switches Ltd, Central Avenue, West Molesey, Surrey.

SEMICONDUCTOR HEATSINKS is a collection of data sheets which list and give brief details of the large range of semiconductor heatsinks available in the U.K. Copies, price 7s. 6d. each are available from DE Data Services, The Information Service Division of the National Trade Press Group, 33-39, Bowling Green Lane, London E.C.1.

DC SOLENOIDS is a new four page leaflet which gives details of the Westool series DK range of solenoids. It contains dimensional data and force-stroke curves in British and metric units. Copies may be obtained from Westool Ltd. St Helen's Auckland, Bishop Auckland, Co. Durham.

PAPER DIELECTRIC CAPACITORS IN RECTANGULAR METAL CASES is a new technical bulletin, No. 107, issued by TCC.

The capacitors described in this Technical Bulletin form a complete range suitable for use over a temperature range from −40°C to +100°C in both Services and professional electronic equipment. Their principal use is in d.c. applications where large blocks of capacitance are required, as in smoothing circuits. Each capacitor is given a d.c. voltage rating at 70°C and 100°C as well as an a.c. voltage rating at 70°C. Capacitance values range from 0·05μF to 12μF and d.c. rated voltages at 70°C from 200V to 10kV. Dimensions and ratings are in accordance with Styles CP10 and CP11 of British Defence Specification DEF-5131.

Copies of this bulletin can be obtained from The Telegraph Condenser Co. Ltd, North Acton, London, W.3.

QUICK-HEATING V.H.F. TETRODES AND DOUBLE TETRODES is a new booklet which provides designers with information on the techniques associated with the use of quick-heating v.h.f. tetrodes and double tetrodes. It gives abridged data on each type, a selection of practical circuits and recommended valve line-ups for different types of transmitter.

Requests for copies should be made to the following address: Industrial Markets Division, Mullard Ltd, Mullard House, Torrington Place, London, W.C.1.

NOUVELLES Réalisations

Traduction des pages 120 à 125

Une description basée sur des renseignements fournis par les fabricants de nouveaux organes, accessoires et instruments d'essai

AMPLIFICATEUR OPÉRATIONNEL

Tektronix UK Ltd, Beaverton House, Station Approach, Harpenden, Hertfordshire

(Illustration à la page 120)

L'amplificateur opérationnel interchangeable à large bande, type 3A8, a été conçu pour l'emploi dans un oscilloscope Tektronix, série 560, pouvant recevoir des fiches à trois séries.

Lorsqu'il est utilisé dans un oscilloscope série 550 ou un bloc d'alimentation type 129, si une image n'est pas exigée, le type 3A8 peut effectuer des opérations précises d'integration, de différenciation, de production de fonctions d'amplification linéaire et non-linéaire.

Chaque élément du type 3A8 fournit deux amplificateurs opérationnels d'une grande souplesse d'emploie ainsi qu'un amplificateur de déviation oscilloscopique. Le produit de la largeur de bande de gain de chaque amplificateur opérationnel est de 10 MHz ou davantage, et le gain sur circuit ouvert à une tension continue est supérieur à 15 000. Les commandes du panneau frontal assurent la sélection sur une gamme étendue de composants d'entrée et de réaction pour chaque amplificateur opérationnel.

Le courant passe-bande de l'amplificateur vertical va de la tension continue à 3,5 MHz; la sensibilité est de 20 mV/division jusqu'à 10 V/division. La commutation permet d'utiliser l'amplificateur vertical indépendamment ou de contrôler la sortie de chacun des amplificateurs opérationnels.

Lorsque le type 3A8 est utilisé avec un oscilloscope d'emmagasinage, comme le Tektronix type 564, on obtient un emploi optimum de ses capacités. Ces applications typiques sont le contrôle des fuites de semi-conducteurs, où la capacité de jonction élevée exige parfois des balayages dans la gamme de 2 sec/division pour éviter les effets de courant de déplacement, et dans les études de phénomènes uniques pour lesquels on employait la photographie dans le passé.

Le type 3A8 est également très utile dans les oscilloscopes qui reçoivent tant des éléments à fiche horizontaux que des éléments verticaux pour le fonctionnement X-Y. L'un des types 3A8 peut être utilisé pour produire et effectuer le report de la variable indépendante; l'autre type 3A8 peut être utilisé pour afficher une fonction de la variable indépendante.

EE 90 751 *pour plus amples renseignements*

ELÉMENTS À LAME VIBRANTE

Hivac Ltd, Stonefield Way, South Ruislip, Middlesex

(Illustration à la page 120)

La société Hivac Ltd produit une nouvelle série d'éléments à lame vibrante avec contact de permutation. D'une construction générale analogue aux autres éléments à lame vibrante Hivac, les nouveaux composants à permutation ont un diamètre de 0,2 pouce et une longueur de verre de 1,375 pouce.

Les éléments à permutation à lame vibrante sont fournis avec deux types de revêtement de contact: le revêtement en or à diffusion partielle, portant la référence XS11, et le revêtement en or à diffusion complète, portant la référence XS13. De plus, ces types sont livrables en deux gammes de sensibilités de fonctionnement: XS11/1 et XS11/2, et XS13/1 et XS13/2.

Les temps de fonctionnement caractéristiques d'un élément d'insertion à lame vibrante sont: moins de 0,5 msec pour les contacts normalement fermés et moins de 2,5 msec pour la paire normalement ouverte. Les deux paires de contacts offrent une résistance fermée inférieure à 90 mΩ, et l'élément a une tension de rupture supérieure à 300 V c.c. Les indices de contact maximum recommandés pour une longue durée de vie sont de 200 mA à 50 V c.c.

EE 90 752 *pour plus amples renseignements*

ELECTRONIC ENGINEERING occupera stand 101 Allée 10 au SALON INTERNATIONAL DES COMPOSANTS ÉLECTRONIQUES qui se tiendra à Paris du 3 au 8 fevrier 1966. Les visiteurs seront les bienvenus.

CHRONOMÈTRE NUMÉRIQUE

Distributeurs: Britec Ltd, 17 Charing Cross Road, London, W.C.2

(Illustration à la page 120)

Le "SERCEL Chronoseis" est un chronomètre à deci-millisecondes entièrement transistorisé avec indication numérique. Conçu principalement pour les mesures sur place il est enfermé dans un coffret portatif en duraluminium mesurant 40 cm × 26,67 cm × 19,68 cm muni d'un couvercle amovible et il comprend une batterie interne au nickel cadmié qui assure huit heures de fonctionnement continu; la recharge peut s'effectuer directement à partir du courant alternatif secteur ou à partir d'un accumulateur externe de 6 V.

Le Chronoseis mesure l'intervalle de temps entre deux impulsions de commande, dans la gamme de 0,1 à 199,9 msec, avec une précision de ±0,1 msec, de −10°C à +60°C. Il comprend un amplificateur à sensibilité élevée et à gain réglable ce qui lui permet de fonctionner avec des impulsions s'étendant en amplitude de quelques microvolts à plusieurs volts. Un dispositif incorporé de retard de temps, gradué en plots de 10 msec, empêche le déclenchement prématuré du circuit de commande par des interférences parasitiques ou des signaux indésirables. Des sorties coaxiales sont prévues sur le panneau de commande et permettent d'injecter les impulsions de commande et un marquage de 10 msec à un oscilloscope.

Le nouveau chronomètre a été conçu principalement pour la mesure des phénomènes sismiques. Lorsqu'il est employé à cette fin, il mesure le temps écoulé entre le moment d'impact d'un marteau ou d'un détonateur sur le sol devant être exploré et l'arrivée du signal de vibration détecté par un géophone placé à une certaine distance du point d'impact. En utilisant des formules simples, il est ensuite aisé de déterminer les profondeurs et les caractéristiques de diverses couches (sable, argile, craie, granite, etc) sous la ligne d'exploration, jusqu'à une profondeur d'environ 50 m.

Le Chronoseis peut être utilisé également pour déterminer l'emplacement d'un point d'impact sur une surface dont les paramètres sont connus, par exemple l'éclatement d'obus sur une zone de cibles sur terre ou sur mer, ou l'atterrissage d'un avion à l'intérieur ou à l'extérieur de la zone de sécurité d'une piste d'envol.

EE 90 753 *pour plus amples renseignements*

GALVANOMÈTRE

Airmec Ltd, High Wycombe, Buckinghamshire

(Illustration á la page 121)

Le galvanomètre Airmec Galvamp type 391 est un nouvel instrument constitué de corps solides et d'une sensibilité élevée. Il a été étudié pour pouvoir résister à un traitement dur, tant électrique que mécanique.

Il se compose essentiellement d'un amplificateur transistorisé, du type modulateur a tension continue, qui alimente un grand instrument de mesure à zéro central. Une excellente stabilité est obtenue par l'emploi de la contre-réaction, et un circuit spécial est employé pour veiller à ce que le zéro soit indépendant de l'impédance de source du signal d'entrée. L'amplificateur fonctionne sur batterie interne de 9 V ou sur tension continue externe de 7 à 9 V.

Le Galvamp a une sensibilité de 2 μV/mm mais des tensions de surcharge jusqu'à 20 V peuvent être tolérées sans risque d'endommagement. Il peut être utilisé dans des conditions de vibrations et de chocs qui rendraient un galvanomètre ordinaire inopérant.

Le Galvamp comporte trois gammes de tension qui se chevauchent, la plus élevée étant une gamme logarithmique de 1 V-0·1 V qui permet d'obtenir rapidement un équilibre approximatif lorsqu'elle est utilisée pour la détection de pont. L'équilibre final à un degré élevé de précision peut alors être obtenu par commutation aux deux autres gammes linéaires (2,5 mV et 100 μV de déviation totale). Cette facilité ainsi que le temps de réponse rapide du mouvement de l'instrument de mesure permet d'effectuer des mesures très rapides.

L'instrument est enfermé dans un petit coffret, le panneau de l'instrument de mesure étant réglé à l'angle le plus approprié pour l'emploi sur un banc d'essai. En variante et sans le coffret, le Galvamp peut être monté si nécessaire à travers une ouverture dans le panneau frontal d'un autre équipement.

EE 90 754 *pour plus amples renseignements*

DÉTECTEUR SENSIBLE AUX PHASES ET DÉPHASEUR

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

(Illustration á la page 121)

L'appareil Brookdeal, modèle PP313A, réunit en un même équipement un détecteur sensible aux phases et un déphaseur utilisant des transistors planaires au silicium à haute fréquence. Grâce à une nouvelle méthode de réalisation, les deux éléments qui composent cet appareil ont une réponse de fréquence allant jusqu'à 300 kHz et permettent de réduire les opérations de réglage au choix de la gamme de fréquence et au réglage approximatif de l'amplitude de la tension de référence.

Le détecteur sensible aux phases se caractérise par une dérive réduite de tension continue et un degré élevé de linéarité. La dérive réduite de tension continue permet d'utiliser l'instrument à des niveaux de signal très bas, réduisant ainsi le risque de surcharge due aux signaux de bruit. Il n'est plus nécessaire, par conséquent, de devoir accorder, dans la plupart des applications, l'amplificateur de signaux utilisé en liaison avec l'équipement PP313A. Les surcharges sont indiquées par un indicateur au néon se trouvant sur le panneau frontal, et un dispositif analogue indique le réglage exact de l'amplitude de la tension de référence.

La sortie du détecteur sensible aux phases permet d'actionner un enregistreur à plume ou un instrument de mesure extérieur; les bornes permettent de relier des condensateurs à la sortie de manière à pouvoir obtenir la constante de temps voulue.

Le déphaseur a été étudié pour assurer la correction des phases de signaux appliqués au canal de référenece du détecteur de phase; un contrôle de phases de 180 degrés peut être effectué. Pour une tension d'entrée déterminée, la sortie du déphaseur ne varie pas plus de 1 dB sur l'ensemble de la gamme de phases; cet avantage, allié au limiteur du canal de référence, obvie à la nécessité de vérifier le retour à zéro de la sortie du détecteur de phases après rétablissement de la phase.

L'appareil peut être fourni pour montage sur bâti de 48,2 cm ou en un coffret pour banc d'essai. Il peut être également monté, avec n'importe lequel des amplificateurs de la gamme Brookdeal, dans un coffret de banc d'essai à deux éléments.

EE 90 755 *pour plus amples renseignements*

GÉNÉRATEUR D'IMPULSIONS À PROGRAMME

Tektronix UK Ltd, Beaverton House, Station Approach, Harpenden, Hertfordshire

(Illustration á la page 121)

Le générateur d'impulsions à programme, type R116, de 10 mHz, 10 V, a été principalement conçu pour les applications exigeant une variété d'amplitudes d'impulsion, de polarité, de forme et d'autres paramètres en séquences rapides, comme dans les systèmes et le contrôle de production. Toutes les fonctions peuvent être établies par programme et, de plus, le type R116 peut être mis en oeuvre à partir de commandes étalonnées sur le panneau frontal pour les applications n'exigeant pas de programmation extérieure.

Les paramètres étalonnés et à programme comprennent la source de déclenchement, la période, le temps de retard, la largeur, l'amplitude, le mode, la compensation de tension continue, le temps de montée, le temps de chute et la polarité. La gamme de commutation, ainsi que la commande variable, peuvent être controlées sur programme pour toutes les fonctions.

Le temps de montée et le temps de chute s'étendent de 10 nsec au minimum à 100 μsec au maximum, l'amplitude s'étend de 0,4 V, à 10 V, la largeur va de 50 nsec à 550 μsec, le temps de retard ou d'éclatement va de 50 msec à 550 μsec, la période sétend de 100 nsec à 11 msec, et la compensation de tension continue s'étend de +5 à −5 V, toutes ces valeurs étant à variation continue. Le temps de montée et le temps de chute sont à variation indépendante lorsqu'ils sont sur la même gamme.

Les modes de fonctionnement spéciaux à programmation comprennent l'impulsion unique retardée qui assure une sortie d'impulsions normale après le temps de retard choisi, la sortie à double impulsion pour deux impulsions par sortie prédéclenchée, l'une des impulsions étant retardée d'une durée variable par rapport à l'autre, la sortie d'éclatement avec un éclatement d'impulsion de sortie pour chaque déclenchement externe appliqué, et la sortie controlée par porte qui fournit des sorties d'impulsions pendant la durée du déclenchement.

La programmation complète exige 21 chiffres binaires et 7 lignes analogiques. Les caractéristiques du type R116 permettent de l'utiliser pour les applications à semi-conducteurs et universels.

Le type R116 est monté dans un bâti standard de 48,2 cm et il ne mesure que 10,79 cm de haut × 46,35 cm de profondeur.

EE 90 756 *pour plus amples renseignements*

CONTRÔLEUR DE TRANSISTORS

Standard Telephones & Cables Ltd, STC House, 190 Strand, London, W.C.2

(Illustration á la page 121)

Le contrôleur de transistors 74163-B que vient de réaliser la société Standard Telephones & Cables Ltd est un instrument portatif robuste et d'une grande simplicité de fonctionnement. Il peut effectuer la mesure du gain de courant dynamique de transistors au silicium et au germanium des types pnp et npn. Des gains allant jusqu'à 200 peuvent être mesurés sur un instrument de mesure à deux gammes à courant émetteur d'environ 1 mA et à tension au collecteur d'environ 5 V. L'instrument de mesure est étalonné directement en gain de courant.

Deux piles sèches sont logées dans le boîtier en alliage léger; ce dernier ne comporte pas de couvercle mais il a, par contre, des brides latérales pour protéger l'instrument de mesure. La commutation par bouton-poussoir facilite le montage des circuits appropriés et les tensions des deux piles sèches peuvent être vérifiées sur l'instrument de mesure. Une prise à trois branches et trois bornes sont prévues sur le contrôleur, de sorte que le transistor soumis au contrôle peut être relié à la prise ou à la borne voulue.

Les dimensions hors-tout du contrôleur, poignée comprise, sont de 197 mm × 133 mm × 92 mm et son poids est de 1,5 kg.

EE 90 757 *pour plus amples renseignements*

CONTRÔLEUR DE HAUTE RÉSISTANCE

K.S.M. Electronics Ltd, 139-149 Fonthill Road, Finsbury Park, London, N.4

(Illustration à la page 122)

Le contrôleur de haute résistance, modèle HR1, est un instrument portatif conçu pour mesurer des résistances élevées dans la gamme de 50 kΩ à 200 mΩ et il est particulièrement utile pour mesurer la résistance d'isolement. Les tensions prévues sont de 100 V et 500 V. La gamme de 100 V est utilisée pour mesurer la résistance dans la gamme de 50 kΩ à 20 000 MΩ et la gamme de 500 V est utilisée pour mesurer la résistance de 500 kΩ à 200 000 MΩ. L'instrument est également étalonné en conductance de 20 micromhos jusqu'à 5 picomhos.

Les deux tensions d'alimentation sont stabilisées pour permettre d'effectuer la mesure de résistances stables. L'instrument a été étudié, en outre, pour mesurer la résistance de composants à capacité élevée, tels que les condensateurs ou les grandes longueurs de câble.

EE 90 758 *pour plus amples renseignements*

APPAREIL DE CONSIGNATION D'INFORMATIONS

Deakin Phillips Electronics Ltd, Tilly's Lane, High Street, Staines, Middlesex

(Illustration à la page 122)

Réalisé sous contrat de la Building Research Station, cet appareil de consignation de données numériques est maintenant commercialisé par le constructeur, la société Deakin Phillips Electronics Ltd, pour la consignation économique de données provenant d'un certain nombre de sources. L'enregistrement de données de chaque source identifie l'événement par le numéro du canal qui lui est attribué. Il enregistre également la durée de son temps de marche, l'instant effectif auquel il se produit et sa séquence par rapport aux autres sources enregistrées.

Ce dispositif a une capacité maxima de canaux de 127 sources et il enregistre le commencement et la fin d'un événement sous forme de deux signaux identifiables sur une carte perforée. Une marque de temps est également perforée sue la carte, de sorte que la durée de l'événement et l'identification de la source sont enregistrés sur un instrument de mesure approprié; elles peuvent toutes deux être reconstituées.

La source de l'événement est codée en chiffres binaires simples et pour distinguer entre le commencement et la fin, le code binaire approprié comporte une perforation supplémentaire dans une position permettant d'identifier l'événement final. L'enregistrement de temps fait l'objet d'un autre code et, dans l'appareil fourni à la Building Research Station, un intervalle de temps de 15 sec est utilisé, bien que d'autres intervalles de temps puissent être prévus. Bien que le système décrit soit étudié pour l'enregistrement d'événements par rapport au temps, n'importe quels paramètres peuvent être enregistrés par rapport l'un à l'autre, ce qui permet d'effectuer des enregistrements de corrélation.

Pour obvier à la perte possible d'enregistrements pouvant se produire lorsque le commencement ou l'arrêt d'un événement a lieu durant l'intervalle de temps au cours duquel la marque de temps est enregistrée, le système comprend un dispositif de réservoirs intermédiaires à inter-verrouillage. Ces réservoirs empêchent la perte supplémentaire de tout enregistrement de temps lorsque le commencement ou la fin d'un événement sont portés sur la carte perforée.

EE 90 759 *pour plus amples renseignements*

CONTRÔLEUR DE CONTAMINATION

EMI Electronics Ltd, Hayes, Middlesex

(Illustration à la page 122)

Un nouvel instrument compact et versatile pour le contrôle de la contamination due aux radiations, le PCM 3, a été réalisé par la société EMI Electronics Ltd.

Le nouveau contrôleur, qui pèse moins de 3 kg, peut être utilisé avec des sondes de détection appropriées pour le contrôle de toutes les formes de radiations. Il est entièrement transistorisé et il est actionné par sa propre batterie interne.

Le PCM 3 a été principalement étudié pour pouvoir être utilisé avec les sondes EMI à substances phosphorescentes, type DP2 et DP2R, et il permet le contrôle simultané de la contamination alpha et beta/gamme. Il peut être utilisé en outre avec d'autres sondes de radiations nucléaires, ainsi qu'avec des sondes d'autres constructeurs de contrôleurs de contamination alpha, beta, gamma et de rayons X.

Le contrôleur fournit une indication visuelle et auditive des niveaux de rayonnement. L'indication visuelle est donnée sur un instrument de mesure de grande précision à échelle de 12,7 cm. L'indication audible est émise par un haut parleur intérieur et au moyen de tonalités distinctes et séparées pour les radiations alpha et beta/gamma. On peut également brancher à l'appareil un relais extérieur qui actionne un dispositif d'alarme à un niveau de rayonnement préréglé.

Un circuit à l'épreuve des pannes est incorporé au contrôleur de sorte que si une sonde est endommagée ou soumise à une radiation excessive, un signal d'alarme sonore et puissant est émis par le haute parleur.

Les circuits à transistors du contrôleur sont montés sur plaquettes imprimées pour assurer une sécurité de fonctionnement maxima. L'instrument de mesure et toutes les commandes sont groupées sur la face supérieure devant la poignée.

EE 90 760 *pour plus amples renseignements*

CONDENSATEURS AU POLYESTER

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

(Illustration à la page 122)

La série de condensateurs à feuille, Mullard C280, pour la radio et la télévision, couvre maintenant une gamme de capacité de 0,01 μF à 2,2 μF en 15 valeurs préférées, à une tension de régime continue de 250 V. Précédemment, la limite capacitive maxima était de 0,22 μF et la tension continue de régime était de 30 V.

L'augmentation de la tension de régime permet d'utiliser les condensateurs dans des circuits à lampes fonctionnant sur ligne de haute tension allant jusqu'à 250 V. Si nécessaire, la tension de régime peut être dépassée pour recevoir l'augmentation normale de haute tension qui se produit durant la période de chauffe de la lampe.

En raison de leur encombrement réduit, de leur forme rectangulaire et de leur revêtement protecteur, les nouveaux condensateurs C280AE assurent une utilisation d'espace maxima lorsqu'ils sont employés sur plaquettes de circuit imprimé.

EE 90 761 *pour plus amples renseignements*

CONNECTEURS À FIL MÉTALLIQUE

Ultra Electronics (Components) Ltd, Western Avenue, London, W.3

(Illustration à la page 122)

Deux nouveaux connecteurs à fil métallique ont été ajoutés à la gamme déjà étendue de connecteurs miniature, sub-miniature et micro-miniature produits par la société Ultra Electronics (Components) Ltd.

Ces deux connecteurs, 12 + 12 et 18 + 18, ont un pas de contact de 5,08 mm et des terminaisons de 1,168 mm². Le contact s'effectue par action à soufflet, et les assemblages de contacts/terminaisons ont l'avantage suplémentaire d'être amovibles.

Les connecteurs sont à extrémités ouvertes, mais l'alignement parfait de la plaquette de circuit imprimé s'effectue par l'emploi d'une clef de référence en nylon.

Les connecteurs ont en outre l'avantage de pouvoir être empilés suivant le principe dit des "briques de construction", ce qui donne n'importe quel nombre de directions voulues en multiples de 12 ou de 18.

EE 90 762 *pour plus amples renseignements*

CONTRÔLEUR DE BRUITS ALÉATOIRES

Derritron Ltd, 24 Upper Brook Street, London, W.1

(Illustration à la page 123)

Le contrôleur automatique de bruits aléatoires Derritron a été conçu pour répondre à la demande croissante d'appareils de ce genre tant au Royaume Uni qu'à l'étranger.

Les circuits perfectionnés du nouveau contrôleur sont constitués de corps solides et comprennent des plaquettes de circuits imprimés à fiches; il en résulte une montée de température très faible à l'intérieur du coffret et un degré élevé de fiabilité et de stabilité.

La gamme Derritron comprend trois consoles de base:

Type ARN.1, 40 canaux
Gamme de fréquence de 5 à 2000Hz
Type ARN.2, 80 canaux
Gamme de fréquence de 5 à 20 000 Hz
Type ARN.3, 160 canaux
Gamme de fréquence de 5 à 5000 Hz

Une commande de panneau assure le changement immédiat de la gamme du ARN.1 et du ARN.2 de 2 kHz à 4 kHz, de 4 kHz à 6 kHz, de 6 kHz à 8 kHz, et de 8 kHz à 10 kHz, et pour le ARN.3, de 5 kHz à 10 kHz.

Nous donnons ci-après une spécification sommaire du ARN.2. Les autres consoles se conforment en général à cette spécification.

Nombre de canaux analyseurs/égaliseurs: 80.
Gamme de fréquences de fonctionnement: 5 Hz à 2000 Hz ou 2-4 Hz, 4-6 kHz, 6-8 kHz, 8-10 kHz par commande sur le panneau frontal.
Gamme dynamique (automatique ou manuelle): 50 dB.
Modes de fonctionnement: Automatique et manuel par commande sur le panneau frontal.
Type de filtre: Magnétostrictif.
Fréquence de fonctionnement des filtres: Fréquence nominale de 100 kHz.
Bourdonnement et bruit: 60 dB au-dessous de la sortie totale.
Programmation: Au moyen de glisseurs sur le panneau frontal; un glisseur par canal, c'est à dire 80 au total.
Gamme de programmation: 50 dB.
Facteur d'ondulation: ±1,5 dB.
Largeur de bande des filtres: Comme ci-dessous.

Largeur de bande des filtres	*Gamme de fréquences*	*Nombre de filtres*
12,5 Hz	0-400 Hz	32
25 Hz	400-1200 Hz	32
50 Hz	1200-2000 Hz	16

Chaque console est fournie avec un générateur de bruits aléatoires, un oscilloscope, un densimètre spectral/voltmètre électronique, des limiteurs de déplacement, de vitesse et d'accélération, un circuit de sortie à charge cathodique par transistors, une entrée d'accéléromètre ainsi que toutes les commandes nécessaires pour le fonctionnement de l'appareil. Un instrument de report XY et un convertisseur logarithmique peuvent être fournis comme suppléments facultatifs.

Notre gravure montre la console Derritron ARN.2 à 89 canaux fournie à un établissement de recherches du gouvernement britannique.

Des consoles de pupitre peuvent être également fournis. Elles sont complètes avec toutes les commandes et tous les instruments pour la mise en oeuvre à distance de systèmes vibrateurs complets.

EE 90 763 *pour plus amples renseignements*

FILTRE MÉCANIQUE MINIATURE

Telefunken A.G., Ernst Reuter Platz 7, 1 Berlin 10, Germany

Ce filtre mécanique miniature, d'une puissance nominale de 455 kHz et mesurant 22,4 × 11,7 × 4 mm, a été spécialement réalisé pour le matériel à bande latérale unique. Grâce à l'emploi de résonateurs à torsion, on obtient des courbes de réponse particulièrement accentuées de sorte que les valeurs de 60 dB des filtres, qui ont été conçus pour des largeurs de bande (3 dB) de 1,5 kHz, 3 kHz et 6 kHz ne sont que de 4 kHz, 6,5 kHz et 40 kHz. L'ondulation est inférieure à 3 dB dans tous les cas, et elle n'est que de 1,5 à 2,5 dB pour les filtres de 6 kHz. Par conséquent, ce filtre est particulièrement indiqué pour les radio-téléphones hyperfréquences. Les capacités à circuit résonant extérieur sont incorporés aux filtre et les bobines à circuit résonnant doivent être branchées extérieurement. On peut ainsi réaliser un équilibrage précis, particulièrement dans les circuits à transistors.

EE 90 764 *pour plus amples renseignements*

PILES DE CHARGE

Westland Aircraft Ltd, Saunders-Roe Division, Osborne Works, East Cowes, Isle of Wight

(Illustration à la page 123)

La Division Saunders-Roe de la société Westland Aircraft Ltd vient d'annoncer l'achèvement d'une nouvelle gamme de piles de charge.

Toutes ces piles utilisent les extensomètres à feuilles agglomérées Saunders-Roe dans un montage en pont complet qui assure une grande précision et une fiabilité parfaite dans un assemblage compact, robuste et peu coûteux.

Trois modèles différents peuvent être utilisés pour les charges de tension et de compression dans les gammes de 0 à 25 livres et jusqu'à 0 à 60 tonnes. Des sorties totales de 20 à 30 mV sont fournies par une résistance en pont de 250 Ω excités par une alimentation de 10 V.

Pour les installations de pesage où la charge est soutenue à plus d'un point, des piles peuvent être fournies sous forme de demi-pont pour permettre l'inter-connexion de 2 ou de 4 ou même de plusieurs cellules dans un circuit composé pour obtenir des lectures précises de charges totales ou moyennes.

EE 90 765 *pour plus amples renseignements*

SYSTÈMES DE NETTOYAGE À ULTRASONS

Radyne Ltd, Wokingham, Berkshire

(Illustration à la page 123)

Cette série de systèmes de nettoyage à ultrasons allie des circuits perfectionnés aux nouvelles méthodes en matière de transducteurs.

Le générateur est basé sur un redresseur piloté au silicium, utilisé comme inverseur, et l'alignement de frequence automatique obvie à la nécessité d'un réglage de fréquence. La simplicité du circuit réduit le nombre des composants à 50% de moins que le nombre requis par les générateurs classiques.

Des éléments indépendants sont produits pour fournir des sorties de 300/600 W, 600/1200 W et 1 kW d'ondes entretenues.

Les bacs de nettoyage sont en acier inoxydable et isolés contre la chaleur. Un transducteur magnétostrictif d'un nouveau modèle fournit une source mécaniquement robuste de puissance à ultrasons qui couvre l'ensemble du fond du bac. Des bobines d'excitation entièrement isolées fonctionnant à une faible tension éliminent les pannes électriques dues à l'humidité. Trois grandeurs standard de bac sont prévues, leur capacité est de 3,5 gallons, 7,25 gallons et 20 gallons. On peut, en outre, utiliser n'importe quelle grandeur de bac avec tous les générateurs et produire ainsi diverses densités de puissance.

EE 90 766 *pour plus amples renseignements*

GÉNÉRATEUR D'IMPULSIONS DE PRÉCISION

Dynamco Instruments Ltd, Salisbury Grove, Mytchett, Aldershot, Hampshire

(Illustration à la page 124)

Le générateur d'impulsions de précision, Datapulse 107, se caractérise par une précision du retard, de la largeur et de la répétition d'impulsions supérieure à 1% dans une température ambiante de 0 à 50° C, par le contrôle d'amplitude à lecture directe, par des circuits constitués de corps solides et par une gamme étendue d'autres fonctions générales.

La commutation par décades et les cadrans Vernier à 10 tours assurent une résolution supérieure à une partie dans 10^3. Le contrôle d'amplitude à lecture directe accélère le réglage et permet des contrôles répétés. Les circuits constitués de corps solides et à compensation de température sont d'une grande fiabilité et la forme d'onde d'impulsion est exceptionnellement claire. Le Datapulse 107 permet le fonctionnement à impulsions simples ou doubles, des taux de répétition de 0,5 Hz à 1 MHz, une sortie de 50 V dans des temps de montée de 50 Ω inférieurs à 10 nsec, avec commande de dégradation, des retards d'impulsion allant jusqu'à 1 sec et des durées d'impulsion de 50 nsec à 10 msec.

La lecture directe, les commandes à haute résolution et le fonctionnement sans dérive font du générateur d'impulsions 107 un appareil particulièrement indiqué pour le contrôle de production et de la qualité. Un oscilloscope n'est pas nécessaire pour le réglage de contrôle ou de stabilité.

EE 90 767 *pour plus amples renseignements*

RELAIS À LAME VIBRANTE

B. & R. Relays Ltd, Temple Fields, Harlow, Essex

(Illustration à la page 124)

La société B. & R. Relays Ltd a mis au point deux nouvelles gammes de relais à lame vibrante, les modèles RO8 et RO9, fournissant ainsi une gamme complète et à bas prix de dispositifs de commutation pour une variété étendue d'applications. Les deux modèles consistent essentiellement en une bobine enroulée sur un rouleau phénolique dans lequel les lames vibrantes sont insérées. Les commutateurs à lame vibrante sont maintenus à l'intérieur du rouleau et cette méthode de construction permet de les retirer facilement.

La gamme RO8 comporte quatre dimensions de relais à lame vibrante étudiés pour pouvoir y incorporer 1, 2, 3 ou 4 commutateurs à lame vibrante miniature.

Le modèles MR 300 et MR 306, qui sont des commutateurs normalement ouverts, et le commutateur à permutation MR 203, peuvent être inclus dans la gamme RO8.

La gamme RO9 est également prévue en quatre versions pour un, deux, trois ou quatre commutateurs à lame vibrante de dimensions normales. Le MR 706, normalement ouvert, la lame vibrante de contact au molybdène à charge élevée MR 840 et le commutateur à permutation MR 201 sont recommandés pour l'emploi avec le RO9.

Il existe des bobines standard des deux types pour tensions continues de 3, 6, 12, 18 et 24 V.

EE 90 768 *pour plus amples renseignements*

DISPOSITIF DE NETTOYAGE D'ÉLECTRODES À ULTRASONS

Electronic Instruments Ltd, Lower Mortlake Road, Richmond, Surrey

(Illustration à la page 124)

La société Electronic Instruments Ltd vient de réaliser un nouvel appareil basé sur les plus récentes techniques de nettoyage par ultrasons. Cet appareil constitue un accessoire facultatif des systèmes à électrodes pH industriels E.I.L. Il existe des trousses d'adaptateurs pour ajuster le dispositif de nettoyage à divers systèmes d'électrodes pH et "redox".

Le principal avantage de ce système est qu'il permet de résoudre les problèmes les plus ardus de nettoyage de revêtements d'électrodes et qu'il fournit un rendement parfait moyennant une augmentation négligeable des frais d'entretien. Une installation dans une fabrique de papier par exemple a permis d'étendre l'intervalle d'entretien des électrodes de deux heures à huit semaines.

EE 90 769 *pour plus amples renseignements*

AMPLIFICATEUR À COURANT CONTINU

Airmec Ltd, High Wycombe, Buckinghamshire

(Illustration à la page 124)

Compact, entierement transistorisé et entraîné par deux piles au mercure intérieures d'une durée de vie normale de 250 heures, cet amplificateur peut recevoir une entrée dans la gamme de fréquence allant du courant continu à 200 kHz et il assure une amplification variable sans plots de ×10 à ×100.

Il est particulièrement utile comme préamplificateur pour travaux oscilloscopiques et il est pleinement compatible avec l'oscilloscope d'affichage Airmec type 279.

Des commandes de zéro préréglées sont prévues pour empêcher le retour à la source de courants indésirables. Ces commandes empêchent également le zéro de sortie d'être affecté par des changements de résistance de source.

La résistance d'entrée à 1 kHz est supérieure à 400 kΩ. La résistance de sortie est inférieure à 1 kΩ à la même fréquence et le niveau de sortie maximum est de ±1 V de pointe au minimum à toutes les fréquences dans des charges supérieures à 50 kΩ. L'amplificateur peut résister à des surcharges d'entrée allant jusqu'à ±30 V de pointe sans risque d'endommagement.

EE 90 770 *pour plus amples renseignements*

TÊTE ENREGISTREUSE EN FERRITE

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

(Illustration à la page 124)

La société Mullard Ltd fournit maintenant une gamme de têtes magnétiques de ferrite pour le matériel d'enregistrement aux fréquences acoustiques.

Contrairement aux têtes en métal feuilleté, les propriétés électriques des têtes d'enregistrement et de reproduction en ferrite restent inchangées pendant leur longue durée de vie. Ainsi les circuits amplificateurs utilisés avec les têtes de ferrite ne nécessitent pas les fréquents réglages compensatoires qui doivent être effectués durant la durée de vie beaucoup plus courte d'une tête métallique. Les essais ont montré qu'après 300 heures de fonctionnement à une vitesse d'avance de la bande de 19 cm/sec, la sortie d'une tête de ferrite, mesurée à 8 kHz, n'était réduite que de 1,5 dB et demeurait substantiellement constante après 5000 heures supplémentaires.

Etant donné que les têtes de ferrite ont une perte de noyau haute fréquence et une perte de courant de Foucault toutes deux inférieures à celles des têtes métalliques, on peut utiliser des courants de polarisation beaucoup plus faibles. On peut ainsi utiliser un courant d'entraînement plus faible et simplifier les circuits enregistreurs et réduire le bruit de tête. On peut, en outre, réduire encore le courant d'entraînement en augmentant la fréquence de polarisation, une technique qui ne donne qu'un avantage marginal avec les têtes métalliques.

La réponse de fréquence d'une tête de ferrite est généralement beaucoup meilleure que celle d'une tête métallique car la longeur de l'espace électrique effectif correspond étroitement à la longeur de l'espace physique. Cela est dû à la nature exempte de contrainte du ferrite qui permet de l'usiner avec précision en cours de fabrication, sans modifier ses caractéristiques électriques, un avantage que les têtes métalliques ne comportent pas.

Des têtes d'enregistrement et de reproduction sont en cours de production. Il existe divers types de têtes jumelées ainsi que des têtes stéréo. Les têtes de reproduction ont des espaces de 3 μm. Les têtes enregistreuses sont fournies avec des espaces de 25 μm 12 μm ou 7 μm. Elles sont toutes à codage de couleurs.

Les têtes peuvent être fournies comme assemblages fixes ou réglables. Les têtes réglables sont logées dans un assemblage à écran de mumétal. On peut, en outre, régler d'avance la hauteur et l'inclinaison ainsi que l'azymuth avec l'écran en mumétal en position. Un écran extérieur magnétique ou non-magnétique peut également être fourni. Avec les versions standard, des cales minces sont prévues pour le réglage de la hauteur de la tête; les adapteurs de montage peuvent également être fournis. Il existe enfin des écrans simple et doubles en mumétal.

EE 90 771 *pour plus amples renseignements*

MATRICE À COUPER ET À ASSEMBLER LES FILS

Kingham Electronics Ltd, 17 Briary Wood Lane, Welwyn Heath, Welwyn, Hertfordshire

(Illustration à la page 125)

Cette nouvelle matrice transistorisée à couper et à assembler les fils obvie à la nécessité d'utiliser des tampons à transistors entre ces derniers et la plaquette de circuit imprimé. La matrice coupe trois ou quatre fils à des longeurs prédéterminées et les assemble en une seule opération. Le transistor peut être fixé dans des ouvertures et reste en place avec un entrefer de 0,31 à 0,39 cm, quel que soit la position de la plaquette de circuit imprimé. On peut ainsi nettoyer facilement la plaquette. Les fils de transistors peuvent être soudés à la main ou par ondes et il n'est pas nécessaire de les ébavurer après soudage. Il existe également des matrices pour l'ébavurage à la longeur voulue et pour le coupage et l'assemblage d'une seule extrémité pour pouvoir utiliser les tampons. La tête de matrice est interchangeable avec toutes les têtes de matrice de sertissage utilisées avec les blocs d'alimentation Pico des modèles 300, 330B ou 300BT.

EE 90 772 *pour plus amples renseignements*

VOLTMÈTRE À GAMME ÉTENDUE

Linstead Electronics Ltd, 35c Newington Green, London, N.16

(Illustration à la page 125)

Cet instrument transistorisé mesure les signaux de courant alternatif sur quinze gammes avec des déviations sur la totalité de l'échelle de 10 mV à 500 V. L'impédance d'entrée est de 10 MΩ, sauf sur les trois gammes les plus sensibles où elle varie entre 6 et 10 MΩ. La gamme de fréquence s'étend de 10 Hz à 100 kHz. Des tensions continues peuvent être mesurées sur trois gammes de 0 à 4 V, de 0 à 40 V et de 0 à 400 V à une impédance de 20 kΩ/V. L'instrument peut être utilisé également comme amplificateur avec une faible impédance de sortie et un gain d'environ 80 sur la gamme la plus sensible.

Il est actionné par une batterie intérieure de 9 V d'une grande capacité composée avec la consommation de courant. L'indicateur est du type à face "nette" de 10,16 cm.

EE 90 773 *pour plus amples renseignements*

INSTRUMENT DE MESURE DU POINT DE ROSÉE

Shaw Moisture Meters, Rawson Road, Westgate, Bradford

(Illustration à la page 125)

Le "Lamplighter", qui vient de faire l'objet d'un brevet provisoire, est un instrument automatique et de réalisation simple servant à indiquer les températures de point de rosée dans la gamme de 0° C à 100° C.

Une lampe électrique spéciale, munie d'une ouverture à travers laquelle s'effectue la conduction lorsque se forme la condensation, constitue la base du nouvel instrument. Un thermistor indique sur un cadran la température de la lampe qui s'allume en cas de condensation et se chauffe jusqu'à ce que la condensation disparaisse. La lecture du point de rosée ne prend que quelques secondes et aucun réglage manuel ou une habileté particulière ne sont nécessaires. De même que plusieurs instruments récents pour la mesure de l'humidité que vient de réaliser cette société, le Lamplighter ne comporte pas de commandes.

Cette méthode assure des indications absolues; l'instrument de mesure du point de rosée est portatif et ne comprend pas de tubes à cathodes chaudes.

EE 90 774 *pour plus amples renseignements*

MASER À ONDES SUB-MILLIMÉTRIQUES

G. & E. Bradley Ltd, Electrical House, Neasden Lane, London, N.W.10

Le "Teratron" est une source de puissance cohérente et à impulsions d'une longeur d'onde de 0,337 mm réalisée par la société G. & E. Bradley Ltd en collaboration avec le National Physical Laboratory. Il fournit des puissances de sortie de pointe allant jusqu'à plusieurs Watts dans cette partie éloignée du spectre. Cette possibilité s'avère déjà de grande utilité dans la spectroscopie à rayon infrarouges, la mesure de la transmission et de la dispersion ainsi que pour les démonstrations pratiques de l'optique des ondes dans l'enseignement et pour la mesure à distance et les communications.

Le système se compose d'un tube à vide de 121 cm de long avec miroirs confocaux ou plans montés à son extrémité, l'un des miroirs étant ajustable pour permettre le changement de la longeur de cavité.

Le matériau actif peut être un des gaz organiques ou une des vapeurs contenant le radical CN, tel que le cyanure de méthyle ou le cyanure de propyle. On fait passer la vapeur dans une des extrémités du tube à vide où une décharge d'environ 10 kV sépare les molécules. Il se produit ainsi une inversion des éléments dans la transition CN nécessaire à l'émission stimulée de radiations. L'effet Maser se produit alors le long du tube de la manière habituelle, cependant que les produits de déchets sont rejetés par pompage continu du tube pour pouvoir être éliminés en toute sécurité.

La sortie s'obtient par une fenêtre en Melinex à l'une des extrémités du tube, soit par diffraction au-delà d'un des miroirs soit par division du faisceau à l'intérieur de la cavité.

La ligne d'émission CN se trouve dans l'un des "miroirs" connus dans la courbe d'absorption de la vapeur d'eau dans l'atmosphère, de sorte qu'il y a de nombreuses possibilités d'utiliser le Teratron pour la transmission à distance, même à travers le brouillard. On surmonte ainsi l'un des principaux obstacles à l'emploi des Masers optiques pour les communications.

EE 90 775 *pour plus amples renseignements*

Résumés des Principaux Articles

Suspension magnétique pour soufflerie aérodynamique par A. Wilson et B. L. Luff

Résumé de l'article aux pages 72 à 76

Cet article traite d'un système à suspension magnétique pour maquettes de souffleries aérodynamiques. Ce système a été conçu pour maintenir des modèles axi-symétriques dans une veine d'air de Mach. 8. Les modèles, qu'on peut faire voler à de faibles degrés d'incidence, sont contrôlés à 5 degrés de liberté, et les forces aérodynamiques peuvent être mesurées dans des conditions se rapprochant d'avantage du vol libre.

L'étude d'amplificateurs de courant continu différentiels à rejet élevé de mode commun par G. Meyer-Brotz et A. Kley

Résumé de l'article aux pages 77 à 81

L'amplification de faibles tensions continues, telles que les signaux des systèmes de traitement de données, exige des amplificateurs différentiels pouvant amplifier le signal mais rejetant les tensions perturbatrices sur les câbles de terre. L'article analyse, en premier lieu, deux amplificateurs de base (un amplificateur à trois bornes et un amplificateur à quatre bornes), puis examine le rejet de mode commun obtenu par quatre amplificateurs différentiels.

L'étude d'antennes de pylone moyenne-fréquence en cage par P. Knight

Résumé de l'article aux pages 82 à 85

Lorsque une antenne de pylone moyenne-fréquence est entourée d'une cage de fils reliés à l'antenne à un point se trouvant à environ un quart de longueur d'onde au-dessus du niveau du sol, la base du pylone peut être mise à la terre et le fond de la cage peut être actionné à sa place. On évite ainsi la nécessité d'utiliser des isolateurs de base portant une charge et le pylone peut être utilisé facilement pour porter des antennes à hyperfréquences, des transformateurs d'isolement n'étant plus exigés. Le rendement du pylone, en tant qu'antenne moyenne fréquence, se trouve, cependant légèrement modifié si la cage de fils ne forme pas un écran parfait.

Etude d'une source de courant et d'un transformateur d'isolement pour des expériences électrophysiologiques
par A. van Oosterom

Résumé de l'article aux pages 86 à 89

Cet article décrit un étage de sortie produisant des impulsions de courant dans une gamme de $10\mu A$ à $25mA$. Il traite également de l'étude d'un transformateur mis au point pour cette source et pouvant recevoir des impulsions carrées allant jusqu'à $12mA$ et d'une durée de $12msec$ dans des charges pouvant atteindre jusqu'à $10k\Omega$.

Un circuit de neutralisation de faisceaux par A. M. Andrew

Résumé de l'article aux pages 90 à 91

Le circuit dont il est question dans cet article fournit des signaux à variations continues se caractérisant par une grande précision entre les points à grande différence de potentiel stable. La réponse de fréquence est linéaire du zéro à une valeur très élevée. Ce circuit ne comporte pas d'oscillateur haute fréquence et il est facile à construire.

Un thermomètre électronique de précision pour la calorimétrie par R. K. Quigg

Résumé de l'article aux pages 92 à 95

Il s'agit ici d'un thermomètre électronique semi-automatique de précision basé sur un pont de Wheatstone à courant alternatif. Les changements de température sont suivis à l'aide d'un amplificateur à bande étroite, à faible bruit et à grain élevé et d'un indicateur de zéro sensible aux phases et amorcé à 75Hz qui entraîne un compteur-minuterie automatique pour l'enregistrement des moments d'équilibre de pont. Des changements de température de 0,0001°C sont enregistrés avec précision.

Lorsque le thermomètre électronique est utilisé avec un calorimètre approprié, il peut mesurer des sorties de chaleur de 10 cal. avec une précision de $\pm 0{,}1$ pour cent.

Considérations sur la forme d'onde complexe de Fourier à z modifié par Y. Azar

Résumé de l'article aux pages 96 à 99

Les tables actuelles de la forme d'onde complexe à z modifié se sont avérées peu satisfaisantes car elles donnent des résultats erronés qui sont toujours retardés par une période d'échantillonnage T et, en outre, elles n'indiquent pas la forme d'onde réelle. La nouvelle méthode élimine le retard artificiel du parcours avant et le place dans l'échantillonneur de sortie fictif. L'auteur montre que cette amélioration produit les formes d'onde voulues. Il développe une formule pour convertir les vieilles tables aux nouvelles et donne quatre exemple de l'application de cette formule. La nouvelle méthode alliée au théorème de la valeur initiale modifiée dote l'ingénieur d'un outil puissant qui lui permet de trouver une valeur unique à n'importe quel moment sans recourir à l'ensemble de la série.

Générateur de signaux aléatoires basse fréquence par P. N. Nikiforuk, R. Pronovost et G. Squires

Résumé de l'article aux pages 100 à 101

Les auteurs décrivent un générateur de signaux aléatoires basse fréquence. La sortie de ce générateur est une tension stationnaire dont la fonction de densité d'amplitude gaussienne, la valeur moyenne nulle et la densité spectrale de puissance sont constantes de zéro à 50 cps. Une méthode de démodulation est utilisée pour produire le signal. Le générateur comprend un schéma de circuit et des caractéristiques de sortie-type.

Un générateur de fonctions SIN/COS à haute résolution par S. Harigovindan

Résume de l'article aux pages 102 à 106

L'auteur décrit un nouveau générateur de fonctions pouvant produire un θ sin/cos comme tension directe d'une précision supérieure à 0,1% à fond d'échelle, lorsque θ est inséré comme chiffre décimal codé binaire avec une résolution de 0,1% dans la gamme de 0 à 2π. Le circuit se compose d'étages diviseurs de potentiel de haute précision commutés par des relais. Deux variantes sont également décrites.

Les équations logiques et d'entrée de circuits flip-flop par N. N. Biswas

Résumé de l'article aux pages 107 à 111

Après avoir donné un bref aperçu des propriétés logiques de cinq circuits flip-flop de base, l'auteur analyse la méthode cartographique pour obtenir les équations d'entrée des circuits flip-flop. Il indique diverses règles sous forme tabulaire, qui peuvent s'avérer à la fois utiles et instructives pour les réalisateurs de circuits logiques séquentiels.

Un tachymètre redresseur de valeurs moyennes et de pointes combinées par A. Kislovsky

Résumé de l'article aux pages 112 à 113

L'auteur décrit un tachymètre pour moteurs à combustion interne à allumage par étincelle. Ce tachymètre comprend un circuit à diode-pompe, comme dans les modèles précedents, mais en raison du choix de constantes de temps, il est d'une insensibilité inhérente aux phénomènes transitoires causés par les rebondissements de contact, etc.

NEUE AUSRÜSTUNGEN

Übersetzung der Seiten 120 bis 125

Beschreibung neuer Bauelemente, Zubehörteile und Prüfgeräte auf Grund der von Herstellern gemachten Angaben.

Funktionsverstärker

Tektronix UK Ltd. Beaverton House, Station Approach, Harpenden, Hertfordshire

(Abbildung Seite 120)

Der Breitband-Funktionsverstärkereinschub 3A8 wurde für Verwendung in jedem Tektronix-Oszillografen der Baureihe 560, der Einschübe der Serie 3 aufnehmen kann, entwickelt.

In einem Oszillografen Baureihe 560 oder—wenn keine visuelle Darstellung nötig ist—mit einer Stromversorgung 129 verrichtet der Einschub 3A8 genau die Operationen Integration, Differenzierung, Funktionserzeugung, lineare und nichtlineare Verstärkung.

Jeder Einschub 3A8 besteht aus zwei vielseitigen Funktionsverstärkern und einem Oszillograf-Ablenkverstärker. Das Produkt Verstärkung × Bandbreite ist für jeden Funktionsverstärker 10 MHz oder höher, und die Leerlaufverstärkung bei Gleichstrom ist höher als 15 000. Frontplatten-Bedienelemente erlauben eine breite Auswahl von Eingangs- und Rückkopplungselementen für jeden Funktionsverstärker.

Das Durchlassband des Vertikalverstärkers ist 0...3,5 MHz, der Ablenkfaktor 20 mV/Teil...10 V/Teil. Schalteinrichtungen gestatten unabhängigen Einsatz des Vertikalverstärkers oder seine Verwendung zur Überwachung des Ausgangs des einen oder anderen Funktionsverstärkers.

Optimale Ausnutzung der Fähigkeiten des 3A8 ergibt sich bei Einsatz mit einem Speicheroszillografen wie z.B. dem Tektronix 564. Typische Anwendungsbeispiele sind Prüfen des Halbleiterreststroms, wo die hohe Sperrschichtkapazität manchmal eine Zeitablenkung von 2 s/Teil erfordert, um den Einfluss des Verschiebungsstroms zu vermeiden, sowie in der Untersuchung einmaliger Vorgänge, für die früher Fotografie angewandt wurde.

Der 3A8 ist auch in Oszillografen sehr nützlich, die für X-Y-Betrieb sowohl Vertikal- wie Horizontaleinschübe aufnehmen. Einen 3A8 kann man zur Erzeugung und zum Auftragen der unabhängigen Variablen einsetzen, während der andere 3A8 zur Darstellung einer Funktion auf der abhängigen Variablen Verwendung finden kann.

EE 90 751 *für weitere Einzelheiten*

Schutzgaskontakte

Hivac Ltd. Stonefield Way, South Ruislip, Middlesex

(Abbildung Seite 120)

Schutzgaskontakte einer neuen Baureihe, die als Umschalter ausgeführt sind, können jetzt von Hivac Ltd bezogen werden. Die neuen Umschaltkontakte entsprechen abmessungsmässig den anderen Hivac-Schutzgaskontakten mit einem Glasdurchmesser von 5,1 mm und 35 mm Länge.

Umschaltkontakte sind in zwei Kontaktplattierungen erhältlich, und zwar als XS11 mit teilweise diffundiertem Gold und als XS13 mit völlig diffundiertem Gold. Ausserdem ist jeder dieser Typen in zwei Ansprechempfindlichkeiten lieferbar: als XS11/1 und XS11/2 sowie XS13/1 und XS13/2.

Typische Ansprechzeiten sind unter 0,5 ms für Ruhekontakte und unter 2,5 ms für das Arbeitskontaktpaar. Geschlossen haben beide Kontaktpaare weniger als 90 mΩ Widerstand; die Durchschlagspannung für die Kontakte ist höher als 300 V—. Für lange Lebensdauer wird als Höchstbelastung der Kontakte 200 mA bei 50 V— empfohlen.

EE 90 752 *für weitere Einzelheiten*

Digitalchronometer

Vertrieb: Britec Ltd. 17 Charing Cross Road, London, W.C.2

(Abbildung Seite 120)

'SERCEL Chronoseis' ist ein volltransistorisiertes Chronometer mit Digitalanzeige in zehntausendstel Sekunden. Das hauptsächlich für den Aussendienst gedachte Instrument ist in einem tragbaren Duralgehäuse von 406 × 267 × 197 mm Abmessungen mit abnehmbarem Deckel untergebracht und wird aus einem internen Nickel-Kadmium-Akku gespeist, der für acht Stunden Dauerbetrieb ausreicht und durch direkten Anschluss an das Wechselstromnetz oder einen externen 6-V-Akku geladen werden kann.

Das Modell Chronoseis misst Zeitintervalle zwischen zwei Steuerimpulsen im Bereich 0,1...199,9 ms und bei Temperaturen von −10°C...+60°C mit ±0,1 ms Genauigkeit. Durch den eingebauten hochempfindlichen Verstärker mit regelbarer Verstärkung können Impulse mit Amplituden von wenigen Mikrovolt bis zu mehreren Volt akzeptiert werden. Eine in 10-ms-Stufen geeichte, eingebaute Verzögerungseinrichtung verhindert vorzeitiges Triggern der Steuerschaltung durch Streustörungen oder ungewünschte Signale. Koaxiale Buchsen im Bedienfeld gestatten, die Steuerimpulse einer 10-ms-Marke in einen Oszillografen zu speisen.

Die Hauptanwendung des Gerätes ist in seismischen Messungen. Bei Einsatz in seismischen Aufschlussarbeiten misst das Chronoseis die zwischen dem Aufschlag eines Hammers oder der Explosion im zu erforschenden Boden und der Ankunft des Schwingungssignals, das durch ein in einiger Entfernung vom Aufschlagpunkt installiertes Geophon nachgewiesen wird, verflossene Zeit. Mittels einer einfachen Formel lassen sich dann leicht Tiefe und Eigenschaften der verschiedenen Straten (Sand, Lehm, Kalk, Granit usw.) bis zu etwa 50 m unter der Vermessungslinie bestimmen.

Das Chronoseis-Modell kann auch zur Bestimmung des Aufschlagpunktes in einem Gebiet mit bekannten Parametern, z.B. eines Artillerieeinschlages im Zielgelände auf Land oder See, oder des Aufsetzens eines Flugzeuges innerhalb oder ausserhalb der Sicherheitszone der Piste eingesetzt werden.

EE 90 753 *für weitere Einzelheiten*

Galvanometer

Airmec Ltd, High Wycombe, Buckinghamshire

(Abbildung Seite 121)

Das Airmec-Galvamp 391 ist ein neues, hochempfindliches Festkörper-Galvanometer, dessen elektrische und mechanische Konstruktion grobe Behandlung aushält.

Es besteht im wesentlichen aus einem transistorisierten Gleichspannungs-Chopperverstärker, der ein grosses Messgerät mit beiderseitigem Ausschlag speist. Durch Gegenkopplung wird ausgezeichnete Stabilität erzielt, und eine Spezialschaltung macht den Nullpunkt von der Quellenimpedanz des Eingangssignals unabhängig. Der Verstärker wird aus einer internen 9-V-Batterie oder einer externen Stromversorgung für 7 ... 9 V— gespeist.

Die Empfindlichkeit des Galvamp ist 2 μV/mm; trotzdem kann es ohne Beschädigung Überlastungsspannungen bis zu 20 V vertragen und mechanischen Schwingungen und Stössen ausgesetzt werden, unter denen konventionelle Galvanometer nicht arbeiten könnten.

Das Gerät hat drei überlappende Spannungsbereiche, deren höchster eine logarithmische Skala für 1 V—0—1 V hat und in einer Brückenschaltung einen schnellen, näherungsweisen Abgleich ermöglicht. Den sehr genauen Endabgleich kann man dann durch aufeinanderfolgendes Umschalten auf die beiden linearen Bereiche (2,5 mV und 100 μV Vollausschlag) erreichen. Durch diese Merkmale und das schnelle Ansprechen des Messwerkes lassen sich Messungen sehr schnell ausführen.

Das Instrument ist in einem kleinen Gussgehäuse unterbracht, in dem die Skala im bequemsten Winkel für Tischbeobachtung angebracht ist. Nach Entfernen des Gehäuses kann das Galvamp auf Wunsch in den Frontplattenausschnitt anderer Geräte montiert werden.

EE 90 754 *für weitere Einzelheiten*

Phasenempfindlicher Gleichrichter und Phasenschieber

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

(Abbildung Seite 121)

Das Brookdeal-Modell PP 313 A enthält in einem Gerät einen phasenempfindlichen Gleichrichter und Phasenschieber, die mit HF-Silizium-Planar-Transistoren bestückt sind. Neue Konstruktionsmethoden geben beiden Teilen einen Frequenzgang bis zu 300 kHz und haben ausserdem die Einstellung so vereinfacht, dass nur noch Frequenzbereichwahl und Grobeinregeln der Amplitude der Bezugsspannung erforderlich sind.

Eigenschaften des phasenempfindlichen Gleichrichters sind äusserst niedrige Gleichstromdrift und ein hoher Linearitätsgrad. Die niedrige Gleichstromdrift gestattet Einsatz bei sehr niedrigen Signalpegeln, wodurch das Risiko der Übersteuerung durch Rauschsignale reduziert wird; es ist daher in den meisten Anwendungen unnötig, als Signalverstärker mit dem PP 313 A einen abgestimmten Verstärker einzusetzen. Übersteuerung wird durch Neonlicht auf der Frontplatte angezeigt; ein ähnlicher Indikator zeigt die korrekte Einstellung der Bezugsspannungsamplitude.

Der Ausgang des phasenempfindlichen Detektors kann einen Schreiber oder ein externes Messgerät treiben. Die Ausgangsklemmen gestatten Anschluss von Kondensatoren für die gewünschte Zeitkonstante.

Der Phasenschieber ist für die Phasenkorrektur der in den Bezugskanal des phasenempfindlichen Gleichrichters gespeisten Signale ausgelegt und gestattet Phasenverschiebung bis zu 180°. Über den gesamten Phasenbereich weicht die Eingangsspannung des Phasenschiebers von seiner Ausgangsspannung um nicht mehr als 1 dB ab, wodurch—in Kombination mit dem Begrenzer des Bezugskanals—die Überprüfung der Nullung des Phasendetektorausgangs nach Änderung der Phase überflüssig wird.

Das Gerät wird entweder für Einbau in ein 19″-Gestell oder im Tischgehäuse geliefert und lässt sich auch zusammen mit jedem beliebigen Brookdeal-Verstärker in ein Doppelgerät-Tischgehäuse einbauen.

EE 90 755 *für weitere Einzelheiten*

Programmierbarer Impulsgeber

Tektronix UK Ltd, Beaverton House, Station Approach, Harpenden, Hertfordshire

(Abbildung Seite 121)

Der programmierbare 10-MHz- und 10-V-Impulsgeber R116 ist in erster Linie für Anwendungen gedacht, in denen verschiedene Impulsamplituden, -polaritäten und -formen sowie andere Parameter in schneller Folge benötigt werden, z.B. beim Testen von Systemen oder in der Fertigung. Alle Funktionen sind programmierbar, der R116 lässt sich aber ausserdem von geeichten Bedienelementen auf der Frontplatte steuern, wenn die Anwendung kein externes Programmieren erfordert.

Geeicht und programmierbar sind u.a. folgende Parameter: Triggerquelle, Periode, Verzögerungs- oder Ballungsdauer, Breite, Amplitude, Betriebsart, Anstiegs- und Abfallzeit sowie Polarität. Schaltbereich sowie regelbare Bedienelemente sind für alle Funktionen programmierbar.

Anstiegs- und Abfallzeit sind zwischen 10 ns und 110 μs, die Amplitude zwischen 0,4 V und 10 V, Dauer zwischen 50 ns und 550 μs, Verzögerungs- oder Ballungsdauer zwischen 50 ns und 550 μs, Periode zwischen 100 ns und 11 ms und Gleichspannungsversetzung zwischen +5 V und −5 V stufenlos regelbar. Anstiegs- und Abfallzeit sind—wenn im selben Bereich—unabhängig voneinander regelbar.

Programmierbare Sonderbetriebsarten sind u.a. ein verzögerter Einzelimpuls, d.h. ein normaler Ausgangsimpuls nach vorgewählter Verzögerung; Doppelausgangsimpulse, d.h. zwei Impulse je Vortriggerausgang mit regelbarer Verzögerung eines Impulses gegenüber dem anderen; Ausgangsstösse mit Ballung von Ausgangsimpulsen für jeden Triggerimpuls; Torschaltungen, die für die Dauer des Einblendens Ausgangsimpulse durchlassen.

Vollständiges Programmieren erfordert 21 Bits und 7 Analogleitungen. Durch seine Kenndaten eignet sich das Modell R116 für Halbleiter- und Universalanwendung.

Modell R116 ist für 19″-Gestelleinbau ausgelegt, 108 mm hoch und 206 mm tief.

EE 90 756 *für weitere Einzelheiten*

Transistor-Tester

Standard Telephones & Cables Ltd, STC House, 190 Strand, London, W.C.2

(Abbildung Seite 121)

Der vor kurzem von Standard Telephones & Cables Ltd eingeführte Transistor-Tester 74163-B ist tragbar, robust konstruiert und einfach zu bedienen. Er misst die dynamische Stromverstärkung von Silizium- und Germanium-Transistoren der pnp- wie auch der npn-Typen. Mit dem Zweibereich-Messinstrument können bei etwa 1 mA Emitterstrom und etwa 5 V Kollektorspannung Verstärkungen bis zu 200 gemessen werden. Das Messgerät ist direkt in Stromverstärkung geeicht.

In dem Leichtmetallgehäuse ohne Deckel—aber mit Seitenflanschen zum Schutz des Messinstrumentes—sind zwei Trockenbatterien untergebracht. Drucktastenschalter vereinfachen das Herstellen der entsprechenden Schaltungsanordnungen; die Spannungen der beiden Trockenbatterien können am Messinstrument kontrolliert werden. Eine dreipolige Fassung und drei Anschlussklemmen sind vorhanden, so dass der Transistor-Prüfling in der bequemsten Weise angeschlossen werden kann.

Die Aussenabmessungen einschliesslich Traggriff sind 197 × 133 × 92 mm und das Gewicht 1,5 kg.

EE 90 757 *für weitere Einzelheiten*

Hochohmtester

K.S.M. Electronics Ltd, 139-149 Fonthill Road, Finsbury Park, London, N.4

(Abbildung Seite 122)

Der Hochohmmesser HR1 ist ein tragbares Instrument für das Messen hoher Widerstände im Bereich 50 kΩ ...

200 GΩ und eignet sich vor allem zum Messen von Isolationswiderständen. Spannungen von 100 V und 500 V stehen zur Verfügung. Der 100-V-Bereich hat einen Messumfang von 50 kΩ ...20 GΩ, der 500-V-Bereich einen von 500 kΩ ... 200 GΩ. Das Gerät ist auch für Leitwertmessungen von 20 μS ... 5 pS geeicht.

Die beiden Speisespannungen sind zur Erzielung konstanter Widerstandsmessungen stabilisiert. Das Gerät kann auch den Widerstand hochkapazitiver Bauelemente wie z.B. Kondensatoren und grosser Kabellängen messen.

EE 90 758 *für weitere Einzelheiten*

Datenregistrierung

Deakin Phillips Electronics Ltd, Tilly's Lane, High Street, Staines, Middlesex

(*Abbildung Seite 122*)

Das im Auftrag des britischen Bauforschungsamtes entwickelte Digital-Registriergerät wird nunmehr von den Konstrukteuren, der Firma Deakin Phillips Electronics Ltd, für die wirtschaftliche Registrierung von Vorgängen in einer Anzahl von Quellen vertrieben. Das Informationsprotokoll jeder Quelle erkennt den Vorgang durch die zugeteilte Kanalnummer und die seit dem Einschalten verflossene Zeit, die die Echtzeit ergibt, in der der Vorgang stattfindet, und sein Auftreten in bezug auf die Reihenfolge der anderen registrierten Quellen.

Die maximale Kanalkapazität der beschriebenen Einrichtung erlaubt Anschluss von 127 Quellen; Anfang und Ende eines Vorganges wird durch zwei erkennbare Signale auf Lochstreifen registriert. Ferner wird eine Zeitmarke in den Papierstreifen gestanzt, so dass sich bei Lesen des Streifens in einem geeigneten Messgerät die Dauer des Vorgangs und Kennzeichnung des Kanals bestimmen lassen.

Die Vorgangsquelle ist in einfachen Binärzahlen kodiert, und zur Unterscheidung zwischen Anfang und Ende hat der entsprechende Binärcode zur Kennzeichnung des Endes des Vorgangs ein in einzigartiger Position gestanztes zusätzliches Loch. Ein weiterer eindeutiger Code erlaubt eine Zeitregistrierung, die in dem für das Bauforschungsamt entwickelten Gerät in internen Intervallen von 15 Sekunden stattfindet; andere Zeitintervalle sind auch möglich. Obwohl das beschriebene System für die Aufzeichnung registrierter Vorgänge in bezug auf Zeit ausgelegt ist, kann offensichtlich jeder Parameter in bezug auf einen anderen registriert werden, was Korrelationsprotokolle ermöglicht.

In dem System werden verriegelte Pufferspeicher verwendet, um Verlust des Protokolls eines Vorganges zu vermeiden, dessen Beginn oder Ende in das Intervall fällt, in dem die Zeitmarke gestanzt wird. Diese Speicher sorgen auch dafür, dass keine komplementären Verluste des Zeitprotokolls auftreten, während Anfang oder Ende des Vorgangs in den Streifen gestanzt wird.

EE 90 759 *für weitere Einzelheiten*

Verseuchungs-Monitor

EMI Electronics Ltd, Hayes, Middlesex

(*Abbildung Seite 122*)

Ein von EMI Electronics Ltd angekündigter neuer Strahlenverseuchungs-Monitor PCM3 ist kompakt und vielseitig.

Der neue Monitor wiegt nur 2,7 kg und kann mit geeigneten Nachweissonden zur Überwachung von Strahlungen aller Art eingesetzt werden. Er ist volltransistorisiert und wird aus einer internen Batterie gespeist.

Der PCM3 ist in erster Linie für Verwendung mit den EMI-Doppelphosphorsonden DP2 sowie DP2R ausgelegt und gestattet gleichzeitiges Überwachen von Alpha- und Beta-Gamma-Verseuchung. Er kann auch mit anderen kerntechnischen Strahlensonden der EMI oder Nachweissonden anderer Hersteller zur Überwachung von Alpha-, Beta-, Gamma- oder Röntgenstrahlungen eingesetzt werden.

Der Monitor zeigt die Strahlungshöhe visuell und akustisch an. Die visuelle Anzeige erfolgt auf einem Messgerät mit 127 mm langer Skala, die akustische Warnung wird durch einen internen Lautsprecher emittiert, und zwar mit klar unterscheidbaren Tönen für Alpha- und Beta-Gamma-Strahlen. Ausserdem ist Anschluss eines externen Relais für Alarm bei einer vorgewählten Strahlungshöhe möglich.

Eine eigensichere Schutzschaltung verursacht Abstrahlen eines hochfrequenten Dauertons über den Lautsprecher, wenn die Sonde beschädigt oder übermässig schweren Strahlungen ausgesetzt wird.

Die Transistorschaltungen des Gerätes sind zur Erzielung höchster Zuverlässigkeit auf gedruckten Schaltungen aufgebaut. Messinstrument und alle Bedienelemente sind vor dem Handgriff gruppiert.

EE 90 760 *für weitere Einzelheiten*

Polyesterkondensatoren

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

(*Abbildung Seite 122*)

Die für Anwendung in Rundfunk und Fernsehen bestimmten Mullard-Polyesterfolien-Kondensatoren der Reihe C280 sind jetzt in 15 bevorzugten Kapazitätswerten zwischen 0,01 μF und 2,2 μF bei einer Betriebsspannung von 250 V— lieferbar. Bisher war die obere Kapazitätsgrenze 0,22 μF und die Betriebsspannung 30 V—.

Diese Erhöhung der Betriebsspannung erlaubt Verwendung dieser Kondensatoren in Röhrenschaltungen mit Anodenspannungen bis zu 250 V. Wenn nötig, kann diese Betriebsspannung um die während der Anheizperiode der Röhre übliche Erhöhung der Anodenspannung überschritten werden.

Die kleinen Abmessungen, rechteckige Bauform und Schutzbeschichtung der mit C280AE bezeichneten neuen Kondensatorenserie ermöglichen—vor allem in Verbindung mit gedruckten Schaltungen—maximale Raumausnutzung.

EE 90 761 *für weitere Einzelheiten*

Steckleisten für Wickelverbindung

Ultra Electronics (Components) Ltd, Western Avenue, London, W.3

(*Abbildung Seite 122*)

Das umfangreiche Fertigungsprogramm der Ultra Electronics (Components) Ltd für Miniatur-, Subminiatur- und Mikrominiatur-Steckverbindungen wurde durch zwei neue Steckleisten mit Wickelverbindung (wire wrap) ergänzt.

Die 12 + 12teiligen und 18 + 18 teiligen Steckleisten haben einen Kontaktabstand von 5,08 mm und Anschlüsse von 1,168 mm². Die Kontaktform gibt eine Balgaktion, und ausserdem ist der Kontakt-Anschluss-Zusammenbau auswechselbar.

Die Steckleisten haben offene Enden, wobei jedoch eine mit Reibsitz passende Führungsfeder für perfektes Fluchten der Druckschaltung sorgt.

Die Steckleisten haben den weiteren Vorteil, dass sie—nach dem Bausteinprinzip gestapelt—jede gewünschte Polzahl, die ein Mehrfaches von 12 oder 18 ist, ergeben können.

EE 90 762 *für weitere Einzelheiten*

Steuereinrichtung für statistisches Rauschen

Derritron Ltd, 24 Upper Brook Street, London, W.1

(*Abbildung Seite 123*)

Die automatischen Derritron-Steuerpulte für statistisches Rauschen wurden zur Befriedigung des Bedarfes an britischen Ausrüstungen dieser Art und zur Befriedigung der wachsenden Nachfrage in Grossbritannien, Kontinentaleuropa und anderen Ländern entwickelt.

Die fortschrittlichen Schaltungen sind alle in Festkörpertechnik auf Einsteck-Druckkarten ausgeführt, so dass der Temperaturanstieg im Gehäuse sehr niedrig ist, wodurch hohe Zuverlässigkeit und Konstanz erreicht werden.

Es gibt drei Grundausführungen der Derritron-Steuerpulte:

Typ ARN. 1 40 Kanäle
Frequenzbereich 5... 2000 Hz
Typ ARN. 2 80 Kanäle
Frequenzbereich 5...20 000 Hz
Typ ARN. 3 160 Kanäle
Frequenzbereich 5... 5000 Hz

Ein Bedienfeldschalter stellt die Bereiche der ARN. 1 und ARN. 2 sofort auf 2 kHz...4 kHz, 4 kHz...6 kHz, 6...8 kHz oder 8...10 kHz und des ARN. 3 auf 5 kHz...10 kHz um.

Untenstehend folgen technische Kurzdaten des ARN. 2 Die anderen Tische stimmen im allgemeinen mit dieser Spezifikation überein.

Anzahl der Analysator-Entzerrerkanäle: 80.
Betriebsfrequenzbereich: 5..8 kHz, 8..10 kHz, durch Frontplattenschalter einstellbar.
Dynamischer Bereich (automatisch oder manuell): 50 dB.
Betriebsarten: Automatisch oder manuell, durch Frontplattenschalter wählbar.
Filtertyp: Magnetostriktiv.
Brummen und Rauschen: 60 dB unter voller Ausgangsleistung.
Programmieren: Mittels Schieber auf der Frontplatte—ein Schieber je Kanal, insgesamt 80.
Programmumfang: 50 dB.
Restwelligkeit: $\pm 1,5$ dB.
Filterbandbreite: siehe untenstehend.

Filterbandbreite	*Frequenzbereich*	*Filterzahl*
12,5 Hz	0-400 Hz	32
25 Hz	400-1200 Hz	32
50 Hz	1200-2000 Hz	16

Jedes Steuerpult wird komplett mit Rauschgenerator, Oszillograf, Röhrenvoltmeter für echte Effektivwerte/Messer für spektrale Dichte, Begrenzer für Weg, Geschwindigkeit und Beschleunigung, Emitter- Folgeverstärker, Beschleunigungsmessereingang und alle für den Betrieb der Ausrüstung notwendigen Bedienelemente geliefert. Als Zubehör können ein X-Y-Schreiber und ein logarithmischer Umsetzer geliefert werden.

Die Abbildung zeigt das 80-Kanal-Derritron-Steuerpult ARN. 2, das an eine öffentliche britische Forschungsstelle geliefert wurde.

Pulte sind auch komplett mit allen für die Fernsteuerung geschlossener Vibratorsysteme benötigten Steuerorgane und Messinstrumente lieferbar.

EE 90 763 *für weitere Einzelheiten*

Mechanische Miniaturfilter

Telefunken A.G., Ernst Reuter Platz 7, 1 Berlin 10, Germany

Mechanische Miniaturfilter für 455 kHz Mittenfrequenz und Abmessungen von 22,4 × 11,7 × 4 mm wurden besonders für ESB-Geräte entwickelt. Durch Verwendung von Torsionsresonatoren werden Kurven mit besonders hoher Flankensteilheit erreicht, so dass die 60-dB-Werte der für Bandbreiten (3 dB) von 1,5 kHz, 3 kHz und 6 kHz gebauten Filter nur 4 kHz, 6,5 kHz bzw. 40 kHz sind. Die Welligkeit liegt in allen Fällen unter 3 dB und beträgt für die 6-kHz-Filter nur 1,5...2,5 dB. Dieses Filter ist daher vor allem für UKW-Funktelefonie geeignet. Die Kapazitäten der externen Resonanzkreise sind im Filter enthalten, die Spulen der Resonanzkreise müssen extern hinzugefügt werden. Das ermöglicht besonders für Transistorschaltungen genaue Anpassung.

EE 90 764 *für weitere Einzelheiten*

Lastdosen

Westland Aircraft Ltd, Saunders-Roe Division, Osborne Works, East Cowes, Isle of Wight

(Abbildung Seite 123)

Die Abteilung Saunders-Roe der Westland Aircraft Ltd gibt die Komplettierung eines neuen Fertigungsprogramms für Lastdosen bekannt.

In allen Dosen finden die bekannten aufgeklebten Dehnmesstreifen der Saunders-Roe in voller Brückenschaltung Anwendung, was hohe Messgenauigkeit und Zuverlässigkeit in einem kompakten, robusten und preiswerten Zusammenbau gewährleistet.

Drei Typen überstreichen die Bereiche von 0...11,3 kg bis zu 0...61 t sowohl für Zug- als auch Drucklasten. Ein mit 10 V erregter Brückenwiderstand von 250 Ω gibt bei voller Last zwischen 20 und 30 mV ab.

Für Wiegeanlagen, in denen die Last in mehr als einem Punkt aufliegt, sind Dosen in Halbbrückenanordnung lieferbar, die Zusammenschalten von zwei, vier oder mehr Messdosen in einer Verbundschaltung gestatten, um genauere Anzeigen für Gesamt- oder Mittelwertlasten zu erreichen.

EE 90 765 *für weitere Einzelheiten*

Ultraschall-Reinigungsgeräte

Radyne Ltd, Wokingham, Berkshire

(Abbildung Seite 123)

Ultraschall-Reinigungsgeräte, in denen fortschrittliche Schaltungen mit neuen Entwicklungen in der Schwingertechnik zusammenwirken, sind jetzt lieferbar.

Der Generator beruht auf einem steuerbaren Siliziumgleichrichter, der als Wechselrichter angewendet wird; Nachstimmen wird durch die automatische Frequenzregelung überflüssig. Durch Vereinfachung der Schaltung werden weniger als 50% der in herkömmlichen Generatoren erforderlichen Bauelemente benötigt.

In sich geschlossene Geräte sind für CW-Ausgangsleistungen von 300/600 W, 600/1200 W und 1 kW lieferbar.

Die aus rostfreiem Stahl gefertigte Reinigungswanne ist ummantelt und wärmeisoliert. Ein magnetostriktiver Schwinger neuer Bauart ist eine mechanisch robuste Ultraschallenergiequelle, die den gesamten Wannenboden überdeckt. Vollisolierte Erregerspulen arbeiten bei niedriger Spannung, um durch Feuchtigkeit verursachtes elektrisches Versagen zu verhindern. Wannen sind in drei Standardvolumen von 16 Liter, 33 Liter und 90 Liter lieferbar. Es ist ein Merkmal des Programms, dass jede Standardwanne mit allen Generatoren bei unterschiedlicher Energiedichte Verwendung finden kann.

EE 90 766 *für weitere Einzelheiten*

Präzisions-Impulsgeber

Dynamco Instruments Ltd, Salisbury Grove, Mytchett, Aldershot, Hampshire

(Abbildung Seite 124)

Unsicherheiten unter 1 Prozent für Impulsverzögerung, Impulsdauer und Folgefrequenz bei Umgebungstemperaturen von 0° bis 50° C, direktanzeigende Amplitudenregler, Festkörperschaltungen und umfangreiche Mehrzweckeinsatzmöglichkeiten kennzeichnen den Präzisions-Impulsgeber "Datapulse 107."

Dekadenschalter und 10gängige Feineinstellskalen gestatten ein Auflösungsvermögen von besser als 1×10^{-5}. Die direktanzeigende Amplitudenregelung beschleunigt das Einstellen und gewährleistet wiederholbare Tests. Die temperaturkompensierten Festkörperschaltungen arbeiten sehr zuverlässig, und die Impulswellenform ist klar gezeichnet. Der 107 bietet sowohl Einzel- wie Doppelimpulsbetrieb, Folgefrequenzen von 0,5 Hz...1 MHz, 50 V Ausgangsspannung an 50 Ω, Anstiegszeiten von unter 10 ns (mit Degradierungskontrolle), Impulsverzögerung bis zu 1 s und Impulsdauern von 50 ns bis zu 10 ms.

Durch direktanzeigende Regler mit hohem Auflösungsvermögen und driftfreiem Betrieb ist der 107 ein für Fertigungsprüfungen und Qualitätskontrolle geeigneter Impulsgeber. Für Testrüsten und Überwachung der Konstanz ist kein Oszillograf erforderlich.

EE 90 767 *für weitere Einzelheiten*

Schutzgaskontaktrelais

B. & R. Relays Ltd, Temple Fields, Harlow, Essex

(Abbildung Seite 124)

Zwei von B. & R. Relays Ltd eingeführte neue Schutzgaskontaktrelais-Baureihen werden mit RO8 und RO9 bezeichnet; sie bilden ein preisgünstiges, umfassendes Schaltelementprogramm mit breiten Anwendungsmöglichkeiten. Die Grundkonstruktion ist für beide dieselbe und besteht aus einer auf einen Phenoplastkörper gewickelten Spule, in die die Schutzgasschalter mit Schiebepassung eingebaut werden. Diese Konstruktion gestattet leichtes Entfernen der Schutzgasschalter.

Baureihe RO8 besteht aus Schutzgaskontaktrelais in vier Grössen, die für Einbau von ein, zwei, drei oder vier Miniatur-Schutzgasschaltern ausgelegt sind. Baureihe RO8 kann mit den Arbeitskontaktschaltern MR300 und MR306 und dem Umschalter MR203 bestückt werden.

Baureihe RO9 ist auch in vier Ausführungen lieferbar, die für ein, zwei, drei oder vier Schutzgasschalter in Standardabmessungen ausgelegt sind. Für diese Baureihe werden Modell MR706 mit Arbeitskontakt, Modell MR840 mit hochbelastbaren Molybdän-Kontaktzungen und MR201 mit Umschalter empfohlen.

Für beide Baureihen sind Standardspulen für 3, 6, 12, 18 und 24 V— lieferbar.

EE 90 768 *für weitere Einzelheiten*

Ultraschall-Elektrodenreiniger

Electronic Instruments Ltd, Lower Mortlake Road, Richmond, Surrey

(Abbildung Seite 124)

Electronic Instruments Ltd hat jetzt als wahlweisen Zubehör zu ihrem industriellen pH-Elektrodensystem ein Gerät nach dem neusten Stand der Ultraschall-Reinigungstechnik herausgebracht. Mit lieferbaren Adaptor-Bausätzen kann das Reinigungsgerät für die verschiedenen pH- und Redox-Elektrodensysteme nachgerüstet werden.

Der Hauptvorteil dieses Systems liegt in der Bewältigung der schwierigen und hartnäckigen Elektrodenverschmutzungsprobleme und ungestörtem Betrieb bei vernachlässigbarer Erhöhung der Betriebskosten. Das Wartungsintervall für die Elektroden in der Anlage einer Papierfabrik wurde z.B. von zwei Stunden auf acht Wochen erhöht.

EE 90 769 *für weitere Einzelheiten*

Gleichstromverstärker

Airmec Ltd, High Wycombe, Buckinghamshire

(Abbildung Seite 124)

Dieser kleine, volltransistorisierte Verstärker wird aus zwei internen Quecksilberelementen mit einer normalen Lebensdauer von 250 Stunden gespeist. Er akzeptiert Eingangssignale im Bereich 0...200 kHz und gibt eine zwischen $\times 10$ und $\times 100$ stufenlos regelbare Verstärkung.

Der Verstärker ist besonders als Vorverstärker für Oszillografen nützlich und ist mit dem Airmec-Oszillografen 279 vollkompatibel.

Voreingestellte Nullregler sind vorhanden, um Rückführung ungewünschter Ströme zur Quelle zu verhindern und zu gewährleisten, dass der Ausgangsnullpunkt nicht durch Änderung des Quellenwiderstands beeinflusst wird.

Der Eingangswiderstand ist bei 1 kHz höher als 400 kΩ, der Ausgangswiderstand bei derselben Frequenz unter 1kΩ; der maximale Ausgangspegel an einer Last von über 50 kΩ ist bei allen Frequenzen mindestens ± 1 V_S. Der Eingang des Verstärkers kann ohne Beschädigung mit Spannungen bis zu ± 30 V_S überlastet werden.

EE 90 770 *für weitere Einzelheiten*

Ferrit-Magnetköpfe

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

(Abbildung Seite 124)

Ein Fertigungsprogramm von Ferrit-Magnetköpfen für professionelle Tonbandgeräte wurde von Mullard Ltd bekanntgegeben.

Die elektrischen Eigenschaften von Ferrit-Magnetköpfen für Aufnahme und Wiedergabe bleiben—ungleich denen von lamellierten Ausführungen—während der vollen langen Lebensdauer unverändert. Es ist daher nicht nötig, die Kompensation der zugeordneten Verstärkerschaltungen nachzuregeln, was während der viel kürzeren Lebensdauer der lamellierten Köpfe sehr häufig geschehen muss. So haben Tests bei 8 kHz und 19 cm/s Bandgeschwindigkeit gezeigt, dass Ausgangssignale nach 300 Stunden nur 1,5 dB niedriger und nach weiteren 5000 Stunden im wesentlichen unverändert waren.

Ferritkerne haben viel niedrigere HF-Verluste und niedrigere Wirbelstromverluste als lamellierte Köpfe, so dass die Vormagnetisierungsströme niedriger gehalten werden können. Daraus ergeben sich auch niedrigere Steuerströme, wodurch die Schaltung des Bandgerätes vereinfacht und das Kopfrauschen reduziert wird. Durch Erhöhung der Vormagnetisierungsfrequenz kann man eine weitere nützliche Herabsetzung des Steuerstroms erreichen—eine Methode, die bei lamellierten Köpfen kaum rentabel ist.

Im allgemeinen ist der Frequenzgang eines Ferritkopfes viel besser als der eines lamellierten Kopfes, da die effektive elektrische Spaltbreite enger mit der mechanischen übereinstimmt. Das ist auf die spannungsfreie Beschaffenheit des Ferrits zurückzuführen, die in der Fertigung genaue Bearbeitung ohne Beeinflussung der elektrischen Eigenschaften gestattet—ein Vorteil, den die lamellierten Köpfe nicht haben.

Es werden sowohl Aufnahme- wie Wiedergabeköpfe angefertigt. Sie sind für Halb- und Vollspur, Stereo-Doppelspur und in zwei Viertelspur-Doppelkopfausführungen lieferbar. Wiedergabeköpfe haben 3 µm breite Spalte. Aufnahmeköpfe sind mit 25 µm Breite (nur für Vollspur), 12 µm und 7 µm Breite lieferbar. Sie sind alle farbkodiert.

Die Köpfe werden entweder starr oder als einstellbarer Zusammenbau geliefert, im letzteren Fall in MU-Metallabschirmung mit einstellbarer "Höhe" und "Neigung" und abschliessender Azimuth-Justierung des ganzen Kopfzusammenbaus. Es gibt auch eine äussere magnetische oder nichtmagnetische Abschirmung. Mit der starren Ausführung kommen Ausgleichscheiben für das Einstellen der Kopfhöhe. Einzel- und Doppelabschirmbecher aus MU-Metall sind ebenfalls zu haben.

EE 90 771 *für weitere Einzelheiten*

Schneid- und Verschränkungswerkzeug

Kingham Electronics Ltd, 17 Briary Wood Lane, Welwyn Heath, Welwyn, Hertfordshire

(Abbildung Seite 125)

Dieses neue Werkzeug zum Schneiden und Verschränken von Transistor-Anschlussdrähten macht das Transistorpolster zwischen Transistor und Druckschaltungskarte überflüssig. Das Werkzeug schneidet drei oder vier Anschlussdrähte auf eine vorgewählte Länge zu und verschränkt sie in einem Arbeitsgang. Der Transistor kann dann in die Standard-Lochanordnung geschnappt werden und behält unabhängig von der Lage der gedruckten Schaltung mit 3 bis 4 mm Abstand von der Leiterplatte seine Position bei. Man kann dann die Platte ohne Schwierigkeiten reinigen, ohne dabei mit den normalerweise bei Polstern auftretenden Festsetzproblemen rechnen zu müssen. Transistor-Anschlussdrähte lassen sich entweder manuell oder im Wellenbad löten, ohne dass die Drähte danach gestutzt werden müssen. Die Werkzeuge sind auch entweder nur für Zuschneiden auf Länge oder nur für Stutzen und Verschränken der zugeschnittenen Drähte lieferbar, falls Polster verwendet werden sollen. Der Werkzeugkopf ist gegen alle Quetschköpfe in den Pico-Kraftwerkzeugen Modell 300, 300B und 300BT austauschbar.

EE 90 772 *für weitere Einzelheiten*

Grossbereich-Voltmeter

Linstead Electronics Ltd, 35c Newington Green, London, N.16

(Abbildung Seite 125)

Dieses transistorisierte Instrument kann Wechselstromsignale in 15 Teilbereichen mit Skalenendwerten von 10 mV bis zu 500 V messen. Mit Ausnahme der drei empfindlichsten Teilbereiche, in denen sie zwischen 6 und 8 MΩ liegt, ist die Eingangsimpedanz 10 MΩ. Der Frequenzbereich ist 10 Hz...100 kHz. Gleichspannungen können bei einer Impedanz von 20 kΩ/V in drei Bereichen 0...4 V, 0...40 V und 0...400 V gemessen werden. Das Gerät kann auch als Verstärker mit niedriger Ausgangsimpedanz und einer Verstärkung von etwa 80 im empfindlichsten Bereich Verwendung finden.

Das Instrument wird aus einer internen 9-V-Batterie gespeist, die eine der Stromentnahme entsprechende grosse Kapazität hat. Das Anzeigegerät hat eine 102 mm lange, deutlich ablesbare Skala.

EE 90 773 *für weitere Einzelheiten*

Taupunktmesser

Shaw Moisture Meters, Rawson Road, Westgate, Bradford

(Abbildung Seite 125)

Der "Lamplighter," für den ein vorläufiges Patent angemeldet wurde, ist ein einfaches automatisches Instrument zur Anzeige von Taupunkttemperaturen im Bereich 0°...100° C.

Eine elektrische Speziallampe mit einem Spalt, der leitend wird, wenn Kondensation auftritt, bildet die Grundlage für das neue Gerät. Die Lampe, die in Gegenwart von Kondensation aufleuchtet, erwärmt den Spalt, bis die Kondensation verschwindet; ihre Temperatur wird mittels eines Thermistors auf einer Skala angezeigt. Die Taupunktbestimmung, für die weder Fachkenntnisse noch manuelle Regelung erforderlich sind, nimmt nur ein paar Sekunden in Anspruch. Genauso wie die vor kurzem von der Firma angekündigten Feuchtigkeitsmesser hat der "Lamplighter" keinerlei Reglerelemente.

Das Verfahren gibt Absolutanzeigen; der Taupunktmesser ist tragbar und enthält keine Röhren.

EE 90 774 *für weitere Einzelheiten*

Submillimeterwellen-Maser

G. & E. Bradley Ltd, Electrical House, Neasden Lane, London, N.W.10

Der "Teratron" ist eine von G. & E. Bradley Ltd in Zusammenarbeit mit dem National Physical Laboratory entwickelte Quelle impulsförmiger, kohärenter Energie bei 0,377 mm Wellenlänge. Sie erzeugt in diesem entfernten Gebiet des Spektrums bis zu mehreren Watt Ausgangsspitzenleistung. Das hat sich bereits in der infraroten Spektroskopie, bei Durchlässigkeits- und Dispersionsmessungen als sehr nützlich erwiesen; ausserdem bestehen Anwendungsmöglichkeiten für die einfache Demonstration der Wellenoptik in Schulen, sowie für Entfernungsmessungen und den Nachrichtenverkehr.

Das System besteht aus einer 1,22 m langen Vakuumröhre, deren Enden mit Plan- oder konfokalen Spiegeln ausgerüstet sind. Ein Spiegel ist verstellbar, so dass man die Hohlraumlänge ändern kann.

Als aktiven Stoff kann man organische Gase oder Dämpfe benutzen, die das CN-Radikal enthalten, wie z.B. Acetonitril oder Propylzyanid. Der Dampf wird in ein Ende der Röhre gesickert, wo eine Entladung von etwa 10 kV die Moleküle dissoziiert. Dadurch wird die für die erzwungene Strahlungsemission erforderliche Besetzungsumkehr im beteiligten CN-Übergang erzeugt. Die Maser-Aktion tritt dann längs der Röhre in üblicher Weise auf, und die Abfallprodukte werden laufend ausgepumpt und gefahrlos weggeschafft.

Die Ausgangsleistung wird durch ein Melinex-Fenster an einem Ende der Röhre abgegeben, und zwar entweder durch Beugung um den Spiegel oder Strahlspaltung im Hohlraum.

Die CN-Emissionslinie liegt in einem der bekannten "Fenster" in der Absorptionskurve von Wasserdampf in der Atmosphäre, so dass der Teratron für Übertragungen über grosse Entfernungen und selbst durch starken Nebel zahlreiche Anwendungsmöglichkeiten hat. Damit würde einer der Haupteinwände gegen Anwendung des optischen Masers in der Nachrichtentechnik entfallen.

EE 90 775 *für weitere Einzelheiten*

Zusammenfassung der wichtigsten Beiträge

Magnetische Aufhängung für einen Windkanal des Royal Airforce Establishment von A. Wilson und B. L. Luff

Zusammenfassung des Beitrages auf Seite 72-76

Ein magnetisches Aufhängungssystem für aerodynamische Windkanalmodelle wird beschrieben. Das System wurde entwickelt, um achsensymmetrische Modelle in einer Luftströmung von Mach 8 festzuhalten. Die Modelle, die mit kleinem Anstellwinkel geflogen werden können, werden in fünf Freiheitsgraden gesteuert, und aerodynamische Modelle lassen sich unter Bedingungen messen, die denen des Freifluges viel näher kommen.

Der Entwurf von direkt gekoppelten Differentialverstärkern mit hoher Gleichtaktunterdrückung von G. Meyer-Brotz und A. Kley

Zusammenfassung des Beitrages auf Seite 77-81

Die Verstärkung niedriger Gleichspannungen die z.B. als Signale in Datenverarbeitungssystemen vorkommen, erfordert Differentialverstärker, die das Signal verstärken, Störspannungen in Erdleitungen jedoch unterdrücken. Mit zwei Grundverstärkern (einem dreipoligen und einem vierpoligen Verstärker) beginnend, bespricht der Beitrag dann die in vier Differentialverstärkern erreichbare Gleichtaktunterdrückung.

Der Entwurf von reusengespeisten KW-Maststrahlern von P. Knight

Zusammenfassung des Beitrages auf Seite 82-85

Wenn ein Kurzwellen-Maststrahler von einer aus Drähten bestehenden Reuse umgeben ist, die mit ihm in einem Punkt ungefähr eine viertel Wellenlänge über der Bodenfläche verbunden ist, kann der Mast geerdet und die Reuse statt dessen über ihr unteres Ende gespeist werden. Dadurch werden tragende Isoliersockel überflüssig, und der Mast kann—da keine Trenntransformatoren erforderlich sind—ohne weiteres als Träger für VHF-oder UHF-Antennen benutzt werden. Die Leistung des Mastes als KW-Antenne wird jedoch etwas beeinflusst, wenn die Reuse keine perfekte Abschirmung bildet.

Entwurf einer Stromquelle und eines Trenntransformators für elektrophysiologische Versuche von A. van Oosterom

Zusammenfassung des Beitrages auf Seite 86-89

Der Beitrag beschreibt eine Stromimpulse im Bereich 10 μA . . . 25 mA erzeugende Endstufe. Einige den Entwurf eines für diese Quelle entwickelten Transformators beeinflussende Überlegungen werden gegeben. Der Transformator akzeptiert Rechteckimpulse bis zu 12 mA und 12 ms Dauer bei bis zu 10 $k\Omega$ Abschluss.

Eine Strahlverdunklungsschaltung von A. M. Andrews

Zusammenfassung des Beitrages auf Seite 90-91

Eine beschriebene Schaltung überträgt mit grosser Genauigkeit stetig einstellbare Signale zwischen Punkten mit grosser gleichförmiger Spannungsdifferenz. Der Frequenzgang ist zwischen 0 und einem sehr hohen Wert geradlinig. Die Schaltung enthält keinen HF-Oszillator und bietet keine Fertigungsschwierigkeiten.

Ein elektronisches Präzisionsthermometer für Kalorimetrie von R. K. Quigg

Zusammenfassung des Beitrages auf Seite 92-95

Ein halbautomatisches elektronisches Präzisionsthermometer, das auf einer Wheatstoneschen Wechselstrombrücke beruht, wird beschrieben. Der Verfolgung der Temperaturänderungen dienen ein leistungsfähiger rauscharmer Schmalbandverstärker und ein bei 75 Hz erregter phasenempfindlicher Nulldetektor, der einen automatischen Zähler-Zeitmesser treibt, so dass die Zeitpunkte des Brückenabgleichs registriert werden. Temperaturänderungen von 0,0001°C werden genau registriert. Bei Einsatz mit einem geeigneten Kalorimeter lassen sich Wärmeabgaben von 10 Kalorien mit ±1 Prozent Genauigkeit messen.

Weitere Überlegungen betreffs der modifizierten z-Transformation von Y. Azar

Zusammenfassung des Beitrages auf Seite 96-99

Die bestehenden Tabellen der modifizierten z-Transformation haben sich als unzulänglich erwiesen und geben fehlerhafte Resultate, die immer um eine Sampling-Periode T verzögert sind und nicht die wahre Wellenform geben. Das neue Verfahren entfernt die künstliche Verzögerung vom Vorwärtsweg und bringt sie in den scheinbaren Ausgangsabtaster. Es wird gezeigt, dass dadurch in der erwarteten Wellenform Verbesserungen auftreten. Eine Formel für das Umwandeln der alten Tabellen in die neuen wird entwickelt und an Hand von vier Beispielen erläutert. Die neue Methode gibt dem Ingenieur in Verbindung mit dem modifizierten Anfangswerttheorem ein leistungsfähiges Werkzeug, das es ihm ermöglicht, jederzeit einen Einzelwert zu finden, ohne die ganze Serie heranzuziehen.

Ein niederfrequenter Rauschgenerator von P. N. Nikiforuk, R. Pronovost and G. Squires

Zusammenfassung des Beitrages auf Seite 100-101

Ein niederfrequenter Rauschgenerator wird beschrieben, dessen Ausgang eine unveränderliche Spannung mit einer Gaussschen Amplituden-Wahrscheinlichkeitsdichtefunktion, einem Mittelwert von Null und einer spektralen Energiedichte ist, die von Null bis zu 50 Hz konstant bleibt. Zur Erzeugung des Signals findet ein Demodulationsverfahren Anwendung. Ein Schaltungsschema und typische Ausgangskenndaten werden gegeben.

Ein sin/cos-Funktionsgeber mit hohem Auflösungsvermögen von S. Harigovindan

Zusammenfassung des Beitrages auf Seite 102-106

Ein neuartiger Funktionsgeber wird beschrieben, der sin/cos θ als Gleichspannung mit einer Genauigkeit von besser als 0,1 Prozent des Endwertes erzeugt, wenn θ als binärkodierte Dezimalzahl mit einer Auflösung von 0,1 Prozent im Bereich 0 . . . 2π eingegeben wird. Die Schaltung besteht aus Präzisions-Spannungsteilerstufen, die durch Relais geschaltet werden. Zwei Ausführungen werden beschrieben.

Die Logik und Eingangsgleichung der Flip-Flops von N. N. Biswas

Zusammenfassung des Beitrages auf Seite 107-111

Nach einer kurzen Besprechung der logischen Eigenschaften der fünf Grund-Flip-Flops analysiert der Beitrag die Planmethode zur Ableitung der Eingangsgleichungen von Flip-Flops. Verschiedene Regeln, die für diejenigen nützlich und lehrreich sein dürften, die sich mit dem logischen Aufbau von Folgeschaltungen befassen, sind in Tabellenform aufgeführt.

Ein Gleichrichtertachometer von A. Kislovsky

Zusammenfassung des Beitrages auf Seite 112-113

Ein Tachometer für Verbrennungskraftmaschinen mit Funkzündung wird beschrieben. Wie in vorhergehenden Entwürfen findet eine Diodenpumpschaltung Anwendung, jedoch ist die gewählte Zeitkonstante inhärent unempfindlich gegen Einschwingvorgänge, die durch Kontaktprellen usw. hervorgerufen werden.