ELECTRONIC ENGINEERING

VOL. 31 No. 380

OCTOBER 1959

Commentary

ON 25 September, 1956, the first transatlantic telephone cable between the United Kingdom, the United States and Canada was opened for public service.

Its conception was an outstanding example of Anglo American co-operation and its design and manufacture was a tremendous engineering achievement.

There was a background of experience of submarine telegraph cables extending into the past for over a hundred years and although there were comparatively short lengths of submarine cables throughout the world nothing on such a grand scale had been done before.

It was, therefore, a risk—a very considerable risk—for in spite of the most elaborate precautions taken during the manufacture and laying, there was always the possibility of breakdown of the cables due to a faulty component or valve.

But such fears have been dispelled and the first three years of operation of the cable have proved without doubt the practicability of transmitting speech over long distances through a deep sea submarine cable using submerged repeaters or amplifiers.

It was anticipated at the time of course, that this would not be the only telephone cable spanning the Atlantic and plans were in preparation for other cables to link the Old World with the New.

Now, just three years after the first venture, a new submarine cable system between the United States and France was formally opened for public service on 22 September of this year.

Owned jointly by the American Telephone and Telegraph Company, the French Ministry of Post and the German Federal Ministry of Posts and Telecommunications, it is of the same design as the first transatlantic cable and has a capacity of 36 channels with terminals at New York, Paris and Frankfurt.

On the American side, the Bell Telephone Network connects with the transatlantic circuit at Portland, Maine, where it is conveyed by a radio link to Sydney Mines, Nova Scotia, a distance of some 575 miles.

A single cable laid across the Cabot Strait to Terranceville on the west coast of Newfoundland and across the mainland, carries the circuits to Clarenville, the point where the main transatlantic cable starts.

This cable is some 330 nautical miles in length and is equipped with 16 two-way repeater amplifiers manufactured by Standard Telephones and Cables of this country.

The link across the Atlantic consists of twin cables, laid 20 miles apart, between Clarenville and Penmarch on the Brittany coast of France, where connexion is made to the telephone networks of France, Germany, Italy, Switzerland, Belgium and the Netherlands.

The total length of the main transatlantic cable is some 2 100 nautical miles and its manufacture was shared by Britain, France, Germany and the United States.

Such then is a brief outline of this second venture which was completed at a cost of about £14M and except for the disastrous fire on the cable ship 'Ocean Layer', which broke out when she had almost completed laying her 900 nautical mile section and which subsequently caused her to be written off, the whole operation has become largely a routine matter.

An even more ambitious project has been occupying the attention of the telecommunication experts of the commonwealth countries for some time past. It is no less than a round-the-world telephone cable linking the principle countries of the commonwealth and it is estimated that the total length of a cable of this nature would be about 27 000 nautical miles or nearly 30 000 nautical miles if the transatlantic crossing is included and the cost would be about £88M.

The technical and financial issues arising from this proposal were examined at a conference held in London early last year from which there emerged a blue print to which approval was given in the autumn at the Commonwealth Trade and Economic Conference at Montreal.

The first stage of this commonwealth link is the construction of a cable of some 1 600 nautical miles between Scotland and Newfoundland and although the rest of the route has not yet been decided in detail, it is anticipated that the west coast of Canada will be linked with New Zealand and Australia, and from there the cable will run through the Indian Ocean connecting Malaya, India, Pakistan and Ceylon, and thence to South Africa by way of Kenya.

The last link of this girdle round the world would then be a cable between South Africa and the United Kingdom.

The present transatlantic telephone links are twin cables with one way repeater amplifiers, one cable carrying traffic in the westerly direction and the other for traffic in the easterly direction, but development by the United Kingdom Post Office Research Laboratories, in co-operation with the communications industry of this country now makes it possible to provide two way traffic for telephone and telex services over a single cable, thus following British practice in the shallow water sections of the existing transatlantic cables and elsewhere.

The stage has now been reached where these new techniques have been accepted as practicable and they will be used for the first time in the cable between Canada and the United Kingdom, which is due for completion in 1961.

The order for this section of some 1 600 nautical miles has now been placed with Submarine Cables Ltd who have also received orders for some 550 nautical miles of armoured cable for the shore ends and for a 400 nautical mile extension from Newfoundland up the St. Lawrence to the Province of Quebec.

Designed to carry 60 telephone channels compared with 36 in the present cables it will be a special lightweight non-twisting coaxial cable developed by the British Post Office.

Instead of the conventional outer armouring its mechanical strength will be provided by a central stranded steel wire which is itself enclosed in a copper tube to provide the central conductor, and the whole is insulated with polythene to a diameter of one inch.

Aluminium tapes form the outer conductor and the final diameter is about one and a quarter inches.

An Amplitude Distribution Meter

By M. Drayson*, B.Sc.Tech., A.M.I.E.E.

The significance of signal amplitude distribution measurements is described, and a brief outline of the mathematical expression of amplitude distribution is given, with particular reference to the approximations involved in any practical measuring system. A novel kind of waveform sampling device has been produced, which permits measurements to a resolution of at least 1 per cent, and an account of this is followed by a detailed description of the circuits associated with it. These provide an instrument suitable for measuring amplitude distributions between the ranges of 0·3 to 30V amplitude and 0·1c/s to 10Mc/s frequency, which will display the amplitude distribution plot of waveforms above 1kc/s minimum frequency. Photographs of distribution plots on a c.r.t. are shown.

(Voir page 636 pour le résumé en français: Zusammenfassung in deutscher Sprache Seite 643)

THE design of communication and control systems is dependent upon the nature of the signals it is required to transmit, and it has become more important with the increasing use of information theory, that both wanted signals and unwanted interference or noise should be completely defined.

Complete definition of a waveform can be provided in terms of its power versus frequency characteristics, its phase versus frequency characteristics and its amplitude distribution density. The last is a statistical property of the waveform, and is therefore particularly useful in providing information regarding the amplitude characteristics of non-repetitive signals such as occur in the normal course of information transmission.

Amplitude Distribution Density

The probability distribution or amplitude distribution density of a time-dependent voltage waveform is evaluated by determining the relative lengths of time the waveform spends at each voltage level throughout its amplitude range. By plotting the relative time durations at each voltage level against the level itself, the complete amplitude distribution density of the waveform is provided.

Mathematically, the amplitude distribution or probability distribution of a time-dependent signal voltage V_t can be expressed as $P(V_t)$ where $P(V_t).\Delta V$ is defined as the probability that the signal V_t will lie within an infinitely narrow range of voltages of width ΔV, between the levels V and $V + \Delta V$.

In the limit, as ΔV, tends to zero, the Δt during which any signal will be within these extremes will also approach zero. The way in which Δt changes with ΔV as the latter tends to zero is a measure of the probability distribution of the signal.

Hence

$$P(V_t) = 1/T\ (\Delta t/\Delta V)_{V\to 0}$$

$$= 1/T \frac{dt}{d(V_t)} \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (1)$$

where T is the maximum time possible for the signal to occupy voltage strip dV, and will in practice be the length of the period for which the probability distribution is examined.

To demonstrate equation (1), consider the voltage sine function $V_t = V_{max} \sin \omega t$ of period $T = 2\pi/\omega$ as shown in Fig. 1(a).

The probability distribution $P(V_t)$ is:

$$P(V_t) = 1/T \frac{dt}{d(V_t)}$$

$$= 1/T\ .\ T/2\pi\ (1 - (V_t/V_{max})^2)^{-\frac{1}{2}}$$

* *The Plessey Co. Ltd.*

Since the waveform occupies all voltage levels twice in one period the correct probability distribution is:

$$P(V_t) = 1/\pi\ (1 - (V_t/V_{max})^2)^{-\frac{1}{2}}$$

This distribution is shown in Fig. 1(b).

In any practical system for evaluating the amplitude distributions of waveforms there will be a finite limit to the minimum width of ΔV, and the signal will lie between V and $V + \Delta V$ for a finite time Δt.

The duration of Δt will be a direct measure of the probability of the signal lying within these limits, and the amplitude distribution density will be the sum of the individual times Δt_1, Δt_2, etc., during which the signal lies within these

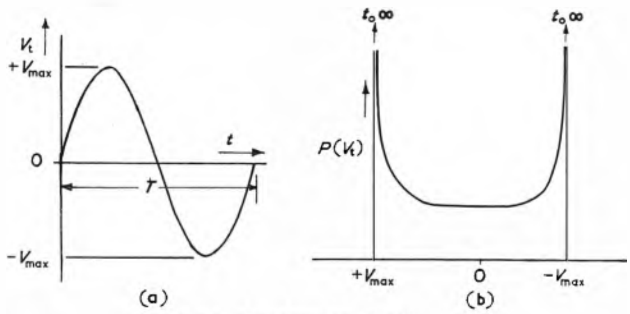

Fig. 1 (a). Sine function $V_t = V_{max} \sin\omega t$
(b). Probability of V_t

limits, divided by the period of observation T, or amplitude distribution density:

$$\frac{\Delta t_1 + \Delta t_2 + \Delta t_3 \ldots \Delta t_n}{T} = \int_{V}^{V+\Delta V} P(V_t)dV_t$$

As ΔV becomes smaller the nearer the amplitude distribution density approaches the true probability distribution of the waveform, and it is evident that for good resolution any practical means of measurement must use a width of sampled voltage ΔV which is very much less than the peak-to-peak amplitude of the measured waveform V_t.

Several specialized instruments which plot the amplitude distribution densities of noise and signal waveforms have already been described[1,2,3]. These are, however, all confined to measurements at relatively low frequencies, and their outputs are either displayed on meters or pen recorders. The time taken to plot the distribution density of a waveform is usually several seconds, since in all cases the observation time at a particular voltage level must be at least the period of the lowest frequency component of the waveform to ensure accuracy.

The input levels required to operate these instruments vary from $50\mu V$ for the study of atmospheric noise at a medium wave intermediate frequency, to 12V for the study of signals from d.c. to 1kc/s.

The requirement, in the present case, was for an instrument which would provide an accurate amplitude distribution density plot of signals and noise waveforms in the range 0·1c/s to 10Mc/s over a wide amplitude range, and which would moreover display quickly and continuously the amplitude distribution plots of all waveforms wherever possible.

A Practical Amplitude Sampling Device[1]

In order to facilitate the measurement of amplitude distribution densities a specially constructed cathode-ray tube was devised, which was manufactured by 20th Century Electronics Ltd. Fig. 2 shows the tube assembly. The electron gun system is of the usual type. Electrons emitted by the heated cathode C are directed and controlled by the grid G and the subsequent anodes A_1, A_2 and A_3 to form a beam which passes between the Y deflexion system of the tube. There is in addition, however, a thin metal tape which is stretched across the diameter of the tube, a small distance behind the tube face and in a plane at right-angles to the

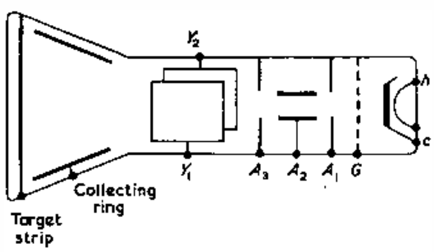

Fig. 2. Sampling c.r.t. assembly

Y deflexion axis. This is connected to a cap on the outside of the tube. The tape emits secondary electrons when the main electron beam of the tube strikes it. The secondary electrons are collected by a conducting layer forming a ring round the inside surface of the tube, which is held at a voltage slightly positive with respect to the target strip.

Provided the beam is falling on the target strip, a current will flow in the circuit caused by connecting the target strip to the collector ring. The magnitude of this current will vary with the beam intensity and velocity.

If the beam is deflected by applying a changing voltage to one of the Y plates, a current pulse will flow in the target strip circuit during the time for which it interrupts the beam. The duration of the current pulse will be dependent upon four factors, namely; the diameter of the electron beam; the width of the target strip; the deflexion sensitivity of the Y system; and the rate of change of the deflecting voltage with time. The first three of these factors are decided by the tube structure, and hence the duration of the tube output will vary inversely with the time rate of change of the deflecting voltage only.

To determine the amplitude distribution density of a particular voltage waveform, it is only necessary to apply it to one of the Y plates, say Y_1, and a known variable voltage to Y_2. Current will flow in the target strip circuit during the time when the voltages on the two deflecting plates are equal, and this period will be inversely proportional to the rate of change of the voltage on Y_1.

The voltage on Y_2 is altered, and by plotting the duration of the current pulse at each new level of this sampling voltage, the current amplitude distribution density plot of the waveform is obtained.

Ideally one would wish to sample the waveform over an infinitely small voltage region to obtain maximum resolution. The resolution in practice is limited by the finite widths of the electron beam and the target strip. If the minimum desirable sampling voltage width is taken to be that voltage change necessary to move the diameter of the electron beam completely across the face of the target wire, and is given the value v volts, and the sampled waveform has a peak-to-peak amplitude of V volts, then the percentage resolution is $v/V \times 100$ per cent.

For the above tube, the diameter of the beam when correctly focused was 0·02in; the width of the target strip was 0·02in; the deflexion sensitivity of the Y plates was 33V/in, and hence the voltage V necessary for the electron beam to traverse the target strip is $33 \times 0{\cdot}02 = 0{\cdot}66$V, and, in order to achieve 1 per cent resolution it was necessary to provide an input waveform to the Y_1 plate of at least 66V peak-to-peak.

The voltage on the Y_2 plate of the sampling tube was made a sawtooth waveform of amplitude at least as great as that of the sampled waveform. This waveform was also used as the time-base for a display cathode-ray tube. The current pulses due to the electron beam striking the target strip at each instant of voltage equality between the two Y plates, were passed through an integrating network to produce a voltage proportional to the sum of the pulse durations over the sampling period. This voltage, after suitable amplification was displayed on the cathode-ray tube, thereby providing a visual plot of the amplitude distribution density of the waveform in question.

Equipment Performance Requirements

The resolution of the sampling tube had been determined as 1 per cent for a sampled waveform of 66V peak-to-peak. This was considered adequate, it being necessary that no circuits subsequent to the sampling tube should degrade the resolution further.

The minimum input waveform amplitude it was considered necessary to examine was of the order of 0·3V peak-to-peak. To achieve the required 1 per cent resolution at this input level it was necessary to provide an amplifier prior to the Y_1 plate of the sampling tube with a gain of at least $66 \div 0{\cdot}3 = 220$ times, and it was decided to design this with an amplitude/frequency response, flat to within -1dB from 0·1c/s to 10Mc/s, with an amplitude roll-off above this range gradual enough to ensure inappreciable phase distortion within the working frequency range.

The phase linearity of the amplifier is important, since in general waveforms have amplitude distribution densities which alter when passed through a non-linear phase shift network.

The accuracy of measurement was required to be such that the plotted ordinate of the amplitude distribution curve at any point should be within 5 per cent of its correct value, ignoring those errors introduced by the finite resolution width of the equipment

To achieve this it was necessary to ensure that the integrated current from the sampling tube output, did not contain more than 5 per cent amplitude variation due to the frequency components of the sampled waveform. It was considered necessary, therefore, to attenuate sufficiently all unwanted frequency components by passing the tube output through a suitably designed low-pass filter. At the same time, it was necessary that the filter should not have a response such that it distorted the true envelope of the amplitude distribution waveform. It was decided to provide different speeds of sampling sawtooth waveforms, and to provide low-pass filters of different bandwidths. The slower the speed of the sampling sawtooth the lower the frequency of the components of the amplitude distribution envelope, and hence the easier it is to select a suitable low-pass filter, and remove the unwanted modulation components without distorting the envelope.

To examine very low frequency waveforms it was considered necessary to replace the sawtooth sampling voltage by a manually controllable d.c. sampling voltage. The low-pass filter associated with this means of measurement could have a time-constant long enough to remove all the unwanted modulation frequencies down to the minimum input frequency of 0·1c/s.

The Equipment

A block diagram of the equipment is shown in Fig. 3.

The sawtooth generator providing the voltage to the Y_2 plate of the sampling tube is a transitron connected Miller integrator. The run-down rate can be switched to give scan durations of 20msec, 40msec or 100msec approximately, or for low frequency measurements the Y_2 plate can be connected to a calibrated variable d.c. supply.

Square waves from the screen grid of the sawtooth generator are applied as brightening pulses to the grid of the sampling tube during the sampling period. The tube is cut-off during flyback, thereby providing a zero output datum level for the display tube. The display tube time-base is also provided by the sawtooth generator.

The input to the equipment is a.c. coupled to a cathode-follower designed to pass all frequencies to 10Mc/s at a maximum amplitude of 30V peak-to-peak. A switchable attenuator following this stage reduces the input to the video amplifier to a level not greater than its limiting sensitivity of 0·3V peak-to-peak.

The video amplifier is a d.c. connected negative feedback amplifier with a closed loop gain of approximately 220 and amplitude response flat to within $\pm 0{\cdot}25$dB to 8Mc/s and not falling by more than 1dB at 10Mc/s relative to d.c. The amplifier output is d.c. connected to the Y_1 plate of the sampling tube.

The target strip current is passed through a high resistance to develop a voltage across the input of a cathode-follower. This is followed by five low-pass filters of different bandwidths of 4kc/s, 2kc/s, 0·8kc/s, 200c/s and 0·05c/s, any of which can be selected to suit the conditions of measurement. The filter output is then amplified by a.c. coupled triode stages, and the output is applied to the Y deflexion system of the display tube.

A 'chopper' using the square pulse from the sawtooth generator can be connected to the input of the low frequency amplifier when low frequency waveforms are to be examined. The d.c. voltage to the amplifier, resulting from sampling at discreet d.c. levels is interrupted by the square wave, thereby providing an a.c. waveform which can be amplified readily by the l.f. amplifier, and then displayed.

At the output of the l.f. amplifier there is a 'windowing' circuit. This can be switched in to allow the examination of any one-tenth part of the output waveform, which is amplified ten times and then displayed as before.

The e.h.t. supply for the two cathode-ray tubes is generated from an r.f. oscillator, transformer coupled to a voltage doubler rectifying circuit which produces a voltage of 2kV negative to chassis. The 2kV is applied to the display tube, while the sampling tube requires a voltage of -1kV. This is stabilized by feeding back via a d.c. amplifier to the screen grid of the r.f. oscillator.

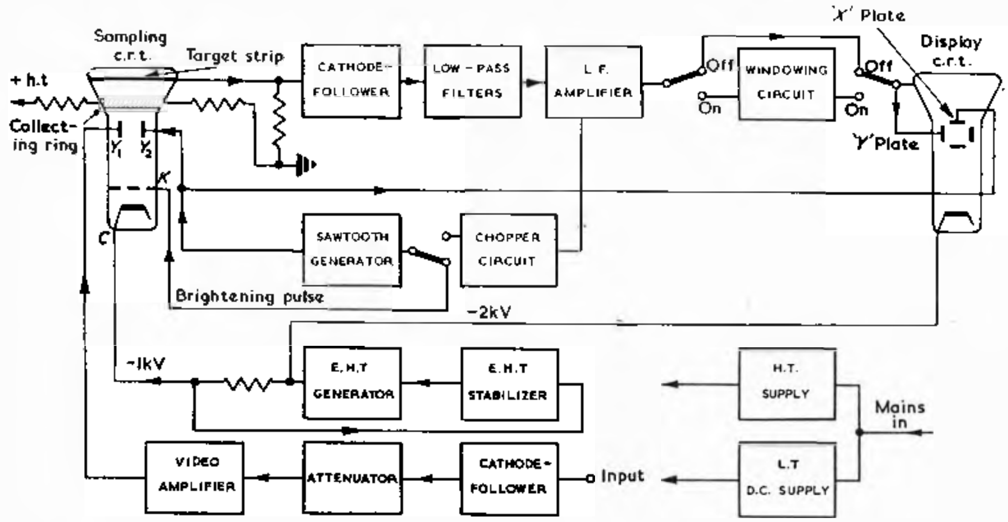

Fig. 3. Arrangement of the equipment

The h.t. supplies are +300V from a series stabilizer delivering 200mA d.c., and $-$ 150V at 10mA from a neon stabilizer. The ripple on both h.t. supplies was kept to less than 10mV peak-to-peak.

The heaters of the video amplifier and l.f. amplifier input stages were supplied from a stabilized d.c. source employing germanium diode rectifiers and a pnp junction transistor series regulating element, with a Zener diode providing a voltage reference.

The circuit diagram of the whole equipment, excluding the h.t. and l.t. supplies is shown in Fig. 4. Further technical description is confined to the low-pass filter and the video amplifier designs.

Low-Pass Filters

The function of the low-pass filters following the output from the sampling tube, is to remove from the envelope of the amplitude distribution waveform, the unwanted frequencies due to the spectrum of the sampled waveform. This must be achieved with minimum distortion to the true envelope shape. Since this waveform will, in general, have fast rates of change of amplitude relative to the sampling speed, it is necessary that the transient responses of the filters should be such that negligible distortion occurs to the changing edges of the waveform. This demands a good filter phase response.

The speed of the transient components is governed by the rate of sampling and it was considered that a maximum

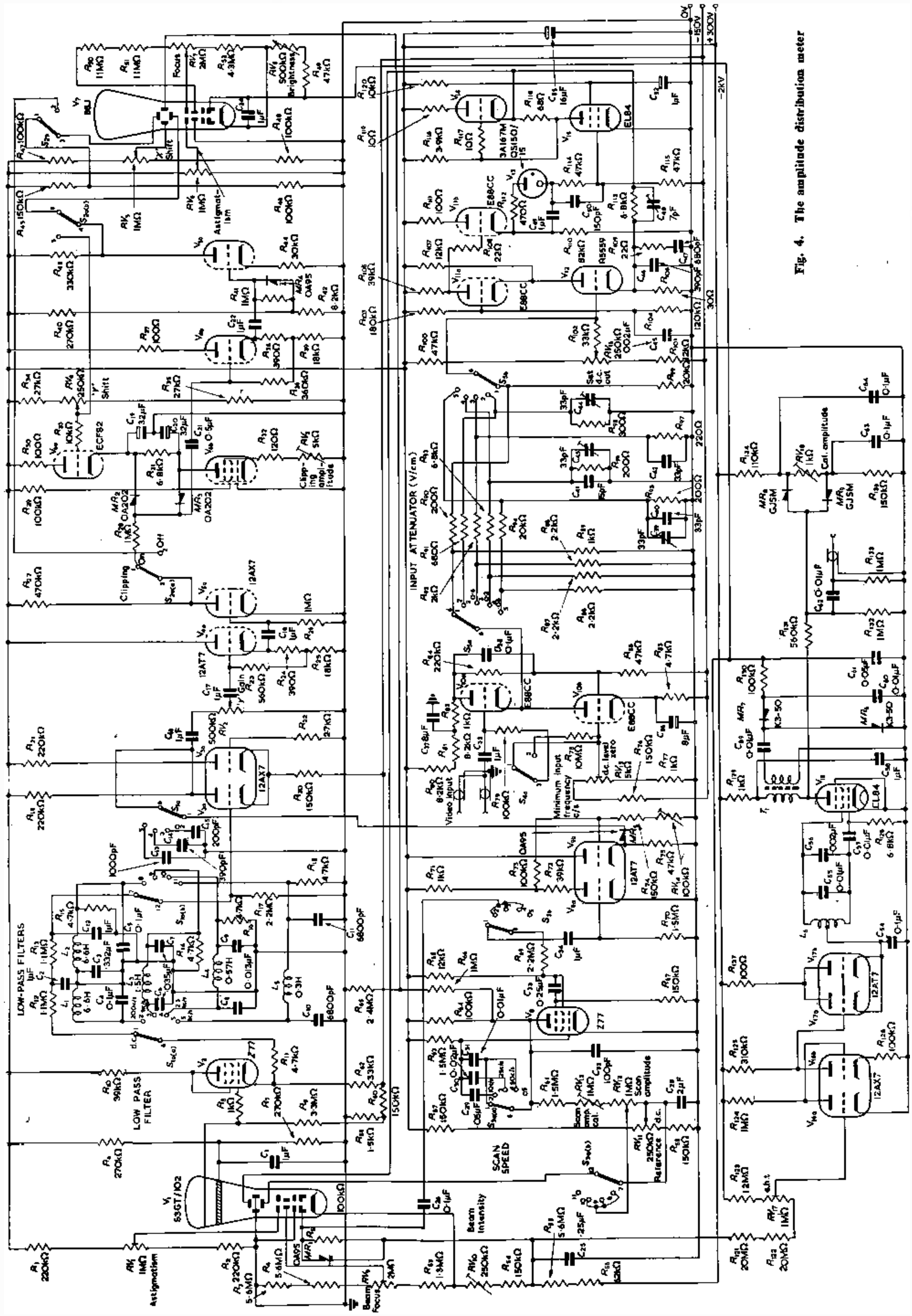

Fig. 4. The amplitude distribution meter

sampling rate of 50c/s would provide a good display, without imposing too severe restrictions on the filter response.

The filter has to respond completely to an input step function voltage in 1 per cent of the scan period of 20msec approximately and this demands a filter response time of 200μsec. The relationship between the frequency response and time-constant of a minimum phase-shift network becomes more complex as the number of elements comprising the network increases. For the simplest form of frequency discriminating network as shown in Fig. 5(a), the time-constant and 3dB bandwidth are related by:

$$f_{(3dB)} = 1/2\pi T \text{ where time constant } T = CR$$

Assuming complete response to an input step-function in five time-constants or a time of 200μsec the necessary bandwidth becomes:

$$f_{(3dB)} = \frac{5 \times 10^6}{2\pi 200} = 4\text{kc/s}.$$

Increased filter complexity will result in a decreased response time for the same bandwidth, and also increased attenuation of frequencies beyond cut-off. There will, however, be a deterioration in phase linearity, causing an inferior transient response.

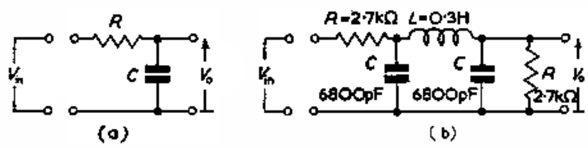

Fig. 5 (a). Single branch low-pass filter

$$\left|\frac{V_o}{V_{in}}\right| = \frac{1}{(1 + (\omega/\omega_o)^2)^{\frac{1}{2}}}$$

$$\Phi = -\text{Tan}^{-1}\,\omega/\omega_o$$

$$\omega_o = 1/CR$$

(b). Three branch low-pass maximally flat matched filter

$$\left|\frac{V_o}{V_{in}}\right| = \frac{1}{(1 + (\omega/\omega_o)^6)^{\frac{1}{2}}}$$

$$\Phi = \text{Tan}^{-1}\,\omega/\omega_o\,\frac{2-(\omega/\omega_o)^2}{1-2(\omega/\omega_o)^2}$$

$$L = 2CR^2 \qquad \omega_o = 1/CR$$

It was finally decided that 4kc/s would be an acceptable filter bandwidth, and attempts should be made to provide as flat an amplitude characteristic to as high a frequency as was compatible with good phase linearity over this region.

This demand automatically rejected the use of m-derived types of filters, since the use of all-pass phase correcting networks was not envisaged. The most suitable network was finally chosen to be that shown in Fig. 5(b).

This is a three branch maximally flat filter, with a matched load. The actual values of components and their relationships are shown on the diagram. The phase and amplitude response of the filter are shown in Fig. 6, curve 2.

The phase response is given, for convenience, in terms of relative time delay T where

$$T = (d\Phi/d\omega)_{\omega=0} - (\Phi)_\omega/\omega$$

Considerable improvement in phase response is achieved if this filter is followed by the simple CR network shown in Fig. 5(a), this having the same bandwidth as the former. The resultant amplitude and relative time delay responses of these two filters together are shown in Fig. 6 curve 3.

In practice each filter has a 3dB bandwidth of 5kc/s, and the resultant overall bandwidth is the required 4kc/s.

Tests using an input waveform rise-time at least as fast as that which could be applied in operation showed negligible overshoot or 'ringing' of the output. The rise-time was no greater than the maximum of 200μsec permissible.

For use with the reduced sampling speeds of 25c/s and 10c/s there were designed similar filters of overall bandwidths of 2kc/s and 800c/s. Two further filters for use at measurements at low frequencies, when the sampling sawtooth is replaced by a variable d.c. voltage, are available. They have bandwidths of 200c/s and 0·05c/s respectively.

Fig. 7 shows the circuit diagram of the video amplifier.

A low capacitance input is ensured by using a cathode-follower as an isolating stage between the input, and the input attenuator and amplifier proper. The time-constant of the input coupling network can be increased for inputs

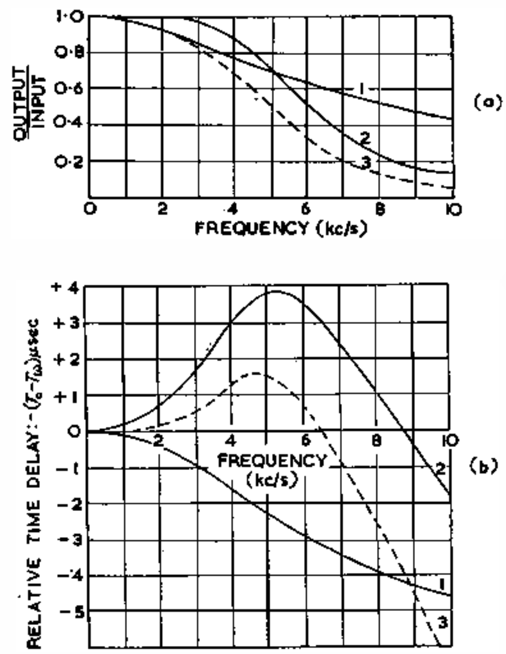

Fig. 6 (a). Amplitude responses of filters
(b) Relative time delay due to filters
1. Single branch filter (see Fig. 5(a))
2. Three branch filter (see Fig. 5(b))
3. Composite filter due to 1 and 2

down to 0·1c/s by switching the value of grid leak from 100kΩ to 10MΩ. Continuous use with a high value of leak resistor was not considered advisable. The input stage is a d.c. coupled shunt-regulated cathode-follower which will pass frequencies to 10Mc/s at 30V peak-to-peak amplitude. Apart from the advantage of a very low input capacitance, the stage, due to its regulating action will also easily provide the current to supply the resistive and capacitive load due to the following attenuator and amplifier stage. A preset control, RV_{15}, adjusts the d.c. level at the grid to give an output balanced about earth.

The switched attenuator following this stage has insertion losses of 0, 10, 20, 30 or 40dB, and is designed to have an input impedance of not less than 1kΩ at low signal levels, and an output impedance constant at 200Ω. This ensures that the d.c. level of the input to the amplifier proper does not alter with attenuator setting. This level is adjusted by RV_{16} to correct for changes of valve characteristics with ageing and ensures a d.c. level of 150V at the amplifier output.

The amplifier is d.c. coupled and has overall negative feedback from output to input via the network R_{113}, R_{106}, C_{16}, C_{18}, C_{10} and C_{17} and consists basically of a cascode stage coupled through a cathode-follower to the shunt regulated output amplifier.

The cascode comprises a triode R.5559 as the bottom valve and an E88C section as the top valve. The gain of a cascode is given by:

$$G = \frac{\mu_1(1 + \mu_2)R_2}{r_{a1}(1 + \mu_2) + r_{a2} + R_2}$$

where $\mu_1 r_{a1}$ and $\mu_2 r_{a2}$ apply to the characteristics of the bottom and top valves respectively.

$\mu_1 = 40$; $r_{a1} = 1{\cdot}6\text{k}\Omega$; $\mu_2 = 33$; $r_{a2} = 5\text{k}\Omega$; $R_2 = 39\text{k}\Omega$

and substituting these values in the above equation gives a gain of approximately 550.

It is apparent that this gain is well in excess of any obtainable from a single pentode with an anode load of identical value working with an anode current of under 4mA as is the top value in this case. The use of the extra envelope to achieve this high figure of merit was considered well worthwhile.

The shunt regulated amplifier has the useful feature of

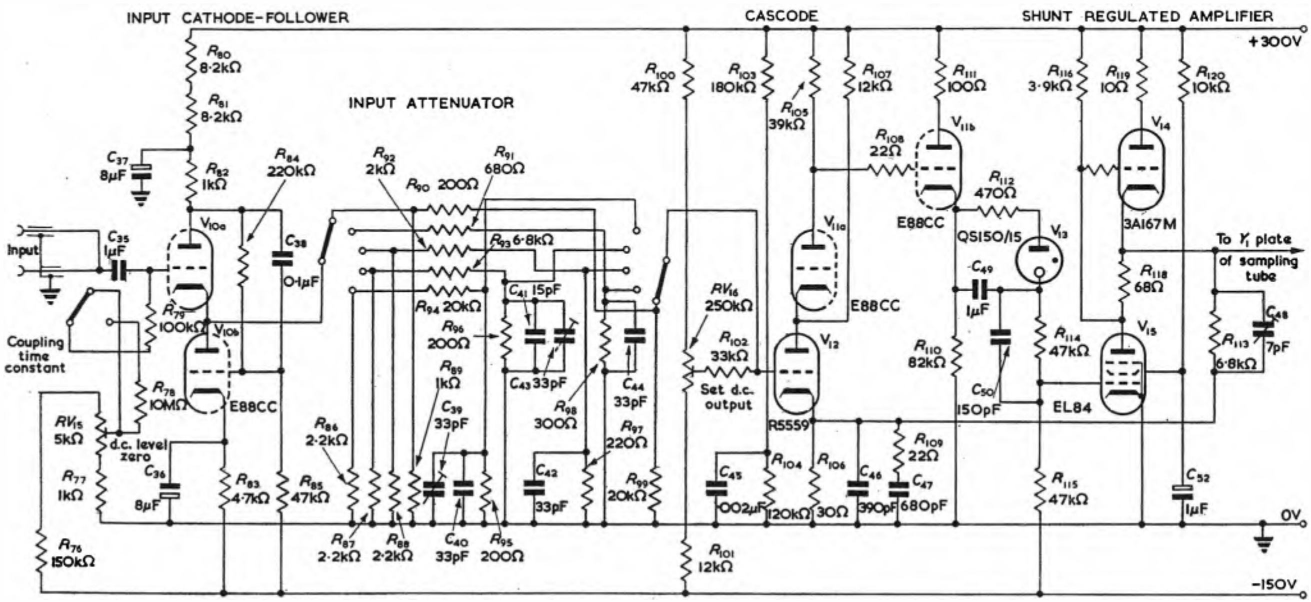

Fig. 7. The video amplifier circuit

being able to provide a large load current, while having a relatively small d.c. current consumption.

Assuming a capacitance of 20pF loading the amplifier output, the required reactive current for 100V peak-to-peak output swing at 10Mc/s is 126mA. For a single valve to provide this current without distortion it would be necessary for it to have a standing direct current of approximately 80mA. With the shunt regulated circuit only 50mA standing current is necessary. By virtue of the d.c. negative feedback there is an increase in d.c. consumption due to the steady drain through R_{113}, but since this would also apply when using a single ended output amplifier, the advantage still remains.

The value of R_{118} is chosen carefully so that the load currents provided by V_{15} and V_{14} are apportioned in the optimum ratio for minimum overloading of both valves at maximum output level.

If one assumes a load at d.c. of R_L, across the output, then at d.c.:

$$\text{Stage gain} = g_{m1}R_L(1 + \alpha)$$

Assuming $R_L \ll$ pentode r_a

Where g_{m1} = pentode mutual conductance,

$$\alpha = \frac{\text{triode a.c. current swing}}{\text{pentode a.c. current swing}}$$

A full derivation of the above expression for stage gain is given in the Appendix.

Using triode 3A/167M for V_{14} and pentode EL84 for V_{15}, the optimum value of α was taken as $\alpha = 0{\cdot}7$, while $g_{m1} = 10\text{mA/V}$.

The value of R_L, due to R_{116}, and R_{113} in parallel is approximately $2{\cdot}5\text{k}\Omega$ and the gain is:—

$$\text{Gain} = 10^{-2} \times 2{\cdot}5 \times 10^3 (1 + 0{\cdot}7)$$
$$= 42.$$

The effect of the negative feedback due to the screen dropper R_{120} for V_{15}, has been neglected, but it is not expected that the overall gain will be reduced appreciably by it.

The output impedance of the stage can be shown to be equal to the triode V_{14} anode impedance, and is therefore in this case $1\text{k}\Omega$.

The open loop gain A, of the complete amplifier, is equal to the product of the individual stage gains and is:

$$A = 550 \times 42 = 23\,000.$$

The required closed loop gain G is at least 220 and hence:

$$\text{Feedback factor } F = A/G \simeq 100.$$

The closed loop output impedance is now $\dfrac{Z_{o(\text{open loop})}}{F}$

$= 10^3/10^2 = 10\Omega.$

The improvement in amplifier bandwidth, stability, distortion level, and insensitivity to component value changes, due to a large feedback factor are well known.

The negative feedback network is shown to consist of R_{113}, R_{106}, C_{48}, C_{46}, R_{109} and C_{47}, the latter four components being selected to give the optimum amplifier frequency response to 10Mc/s.

The heaters of V_{12} are connected to a stabilized low voltage d.c. supply. This is necessary to reduce the drift in output d.c. level due to mains voltage changes.

The amplifier stability is such that after an initial five minutes warming up period, the output level remains constant to within 2V of its design level of 150V d.c. and changes of this magnitude are not sufficiently fast to cause a deterioration in the accuracy of measurement.

The overall frequency response of the amplifier at full output level is flat to 8Mc/s and does not fall by more than 2dB at 10Mc/s. At half full output level the response does not fall by more than 1dB at 10Mc/s.

Performance and Facilities

The equipment is fitted with provision for photographing its display tube waveforms, and in Fig. 8 are shown the amplitude distribution density plots obtained from various input waveforms.

A series of interchangeable graticules for insertion before the display tube face, was prepared. These included the amplitude distribution densities of random functions, i.e.,

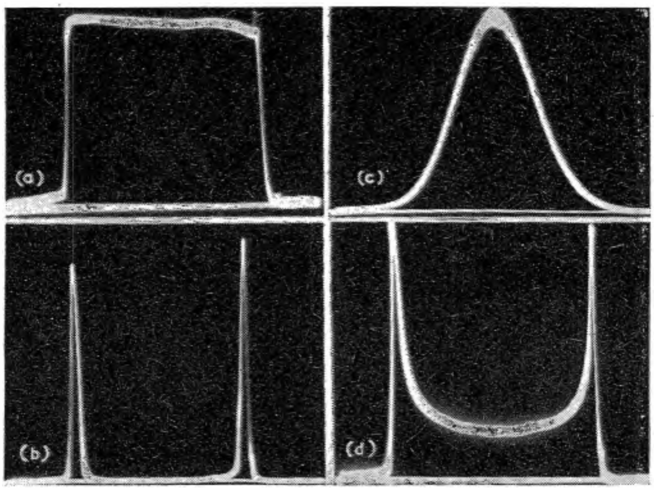

Fig. 8. Traces of amplitude distributions of
(a) *Sawtooth*
(b) *Square wave*
(c) *Gaussian noise*
(d) *5Mc/s sine wave*

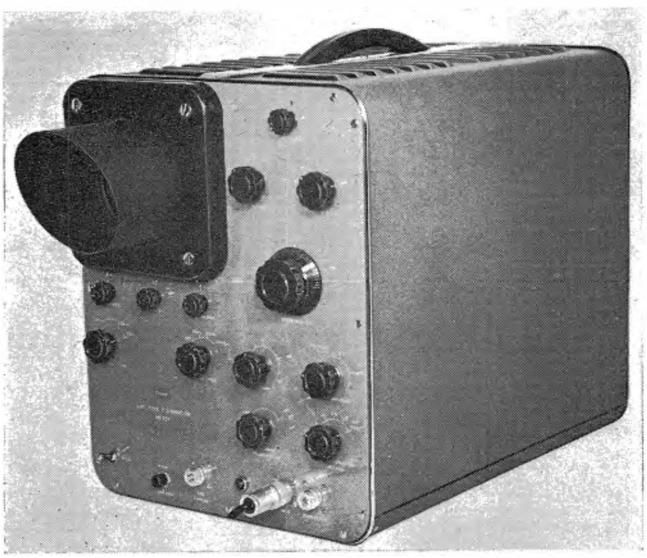

Fig. 9. The amplitude distribution meter

the Gaussian distribution, exponential distributions, and the sinc-function distribution as shown in Fig. 1(b). There was also a graticule marked in 1cm squares which allowed reasonably accurate measurement of input signal amplitudes and also provided an easy means of evaluating the relative probability waveforms of low frequency inputs when using the d.c. sampling method. In these circumstances the display tube output is a horizontal trace displaced vertically by an amount proportional to the amplitude distribution density.

A photograph of the equipment is shown in Fig. 9.

Acknowledgments

The author would like to express his thanks to Mr. L. C. Walters and others for their help and advice, and also to the Directors of The Plessey Co. Ltd for permission to publish this article.

Appendix

Analysis of Shunt-regulated Amplifier

The amplifier circuit is shown in Fig. 10(a) and its equivalent circuit in Fig. 10(b). From the latter:—

$$i_1 = g_{m1}\ V_{in} \quad \ldots\ldots\ldots\ldots \quad (1)$$

$$i_2 = \frac{\mu_2 e_2 - V_o}{r_{a2}} \quad \ldots\ldots\ldots\ldots \quad (2)$$

$$e_2 = i_1 R \quad \ldots\ldots\ldots\ldots \quad (3)$$

$$V_o = (i_1 + i_2) Z_L \quad \ldots\ldots\ldots\ldots \quad (4)$$

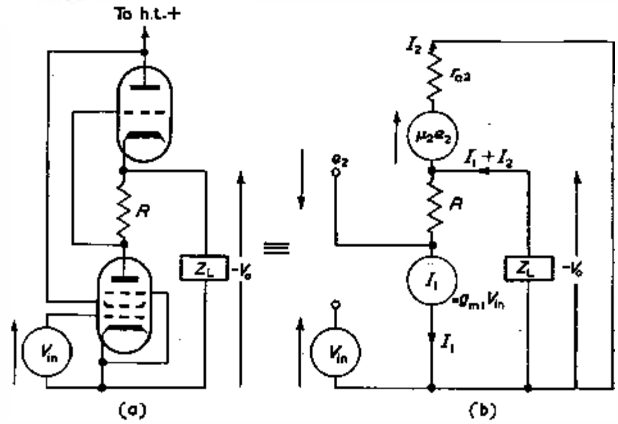

Fig. 10. Shunt regulated amplifier

Stage Gain

Substituting in equation (2) for e_2 and V_o, from equations (3) and (4) respectively:—

$$i_2 = g_{m2}\ i_1\ R - Z_L/r_{a2}\ (i_1 + i_2)$$

Therefore $i_2(1 + Z_L/r_{a2}) = i_1(g_{m2} R - Z_L/r_{a2})$

Let $i_2/i_1 = \alpha$ = ratio of triode a.c. current to pentode a.c. current.

$$\therefore \quad \alpha = \frac{\mu_2 R - Z_L}{r_{a2} + Z_L} \quad \ldots\ldots\ldots\ldots \quad (5)$$

Now $i_2 = \alpha\ i_1$ and $V_o = (i_1 + i_2)Z_L$.

Therefore:—$V_o = i_1 Z_L(\alpha + 1)$.

Since from equation (1) $i_1 = g_{m1}\ V_{in}$

$$\text{Gain} = V_o/V_{in} = g_{m1}\ Z_L(\alpha + 1) \quad \ldots\ldots\ldots \quad (6)$$

Stage Output Impedance

Substituting for α from equation (5) in equation (6).

$$\text{Gain} = g_{m1}\ Z_L \frac{r_{a2} + \mu_2 R}{r_{a2} + Z_L}$$

If $Z_L = \infty$ the open-circuit gain is

$$G_{o.c.} = g_{m1}\ (r_{a2} + \mu_2 R).$$

$$\therefore\ G = G_{o.c.} \frac{Z_L}{r_{a2} + Z_L}$$

From this it is obvious that the output impedance Z_o is equal to r_{a2}, the triode anode impedance.

$$Z_o = \text{Triode anode impedance};\ r_{a2} \quad \ldots\ldots\ldots \quad (7)$$

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The Design of Transistor Push-Pull D.C. Convertors

By W. L. Stephenson*, B.Sc., L. P. Morgan*, B.A., and T. H. Brown*

There are many methods by which d.c. may be converted from low to high voltage using some oscillating device, but one of the most efficient methods uses a transistor square wave oscillator controlled by a saturating transformer. For such a system design formulae are derived in terms of operating parameters.

Most of the practical limitations and difficulties are outlined, but are not discussed in detail as individual solutions are usually required.

(Voir page 636 pour le résumé en français: Zusammenfassung in deutscher Sprache Seite 643)

THE conversion of low to high voltage d.c. requires some oscillating device to convert low voltage d.c. to a.c. which can then be stepped up in a transformer and rectified at a high voltage.

The efficiency of the oscillatory system is of prime importance where large powers are required to be handled. The transistor when used as a switch is an extremely efficient device, passing high current at low voltage in the 'on' condition and passing very low currents in the 'off' condition, and is therefore very suitable for this application. Such an oscillatory system can be arranged using two transistors in push-pull[1]. This has the advantage over the single-ended system[2] that the voltage across the transformer and hence the peak voltage applied to the transistors is limited to twice the supply voltage, and that the power supplied to the load is never stored in the transformer, but passes from the conducting transistor through the transformer into the load. In this way more power can be handled at greater efficiency, and with better regulation than with the single-ended circuit.

For the push-pull circuit to work at maximum efficiency, the circuit should work into the load on both halves of the operating cycle, i.e. the output circuit should be either a full-wave or a voltage doubler rectifier system. Where a half-wave rectified output is required, only one of the transistors sees the load, and so the other may be a low power type which is only required to pass inductive current to reset the transformer core.

Methods of Push-Pull Operation

SINUSOIDAL: CLASS-C

This method of operation is the most efficient way of providing a.c. power in sinusoidal form, but in cases where the output is to be rectified, neither the d.c.-a.c. inversion nor the final rectification is as efficient as with square wave operation.

SQUARE WAVE: DRIVEN OUTPUT STAGE

At one time it was thought that for high powers it would be difficult to arrange for the output stage to oscillate and still be efficient, but this was subsequently shown to be untrue. The main disadvantages of a driven system are the low overall efficiency due to the drive stage, and the necessity for either matched output transistors or a large output transformer to carry the difference current without saturating. On the other hand such a system has the advantage that the frequency of operation is completely independent both of the applied load and of the supply voltage to the output stage.

SQUARE WAVE: SELF OSCILLATING

With this system of operation, feedback is applied to the output transistors from the output transformer (Fig. 1). The voltage across the transformer is limited by the bottoming of the transistors, so that the feedback voltage is determined by the supply voltage and the feedback turns ratio. The mode of operation of the device depends very much on whether the transformer is designed to saturate or not.

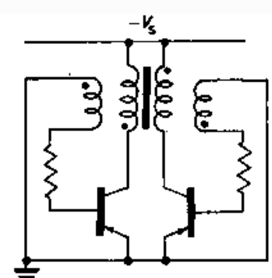

Fig. 1. Basic self-oscillating circuit

Non-saturating Transformer

During one half of the operating cycle, one transistor is bottomed and a constant voltage is developed across the transformer primary winding. A certain amount of drive voltage is therefore applied to the transistor from the output transformer.

The bottomed transistor has a low output resistance so that its collector current is determined by the load, which consists of a reflected load resistance in parallel with the transformer shunt inductance. The current through the resistive portion (I_R) is constant and the current through the inductance (I_L) increases linearly with time (Fig. 2(a)). When the total collector current reaches its maximum value (I') determined by the drive, the voltage across the transformer falls and feedback causes the transistor to switch off; the time for the half cycle (T) is therefore given by:

$$I' = I_R + (V_s T / L) \quad \ldots\ldots\ldots\ldots\ldots \quad (1)$$

where V_s is the supply voltage and L the transformer shunt inductance. Due to the self balancing action of the transformer, it can be shown that the appropriate value of I' is the average of the peak currents available from the two transistors.

It can be seen that the half-cycle time T will depend on I_R which means that there will be considerable variation of operating frequency with load.

Saturating Transformer

In the case where the transformer is designed to saturate using a square loop material, the resistive current I_R is

* *Mullard Research Laboratories.*

constant as before, but the inductive current is initially very small, and increases rapidly as saturation is reached. In this way the half-cycle time is made relatively independent of the load, and also of the peak current of the transistor.

The Self-Oscillating Circuit with a Saturating Transformer

PRINCIPLES OF OPERATION

The basic circuit is shown in Fig. 1. The circuit is a simple push-pull oscillator, with the transistors operated in grounded emitter.

With one transistor on and bottomed the collector current has a resistive component (I_R) through the reflected load, and an inductive component (I_L) through the transformer shunt inductance.

Since the voltage across the transformer is constant and approximately equal to the supply voltage, the current I_R is constant and the current I_L is initially small (high-inductance) and increases rapidly as the transformer nears saturation (Fig. 2(b)). When the transistor can no longer supply the rising current needed to maintain the voltage across the transformer, this voltage falls and feedback causes the conducting transistor to be cut off. The rapid reduction of current in the shunt inductance causes an overshoot voltage to appear across the transformer, and this is applied to the other transistor (which was previously cut off) causing it to conduct. The result is rapid regeneration as the circuit switches over. It will be apparent that the switching operation is affected by the value of the shunt inductance in the saturated region, so that too few turns for the area/length ratio of the transformer core will cause poor regeneration resulting in an output waveform as shown in Fig. 3(b). This design parameter is normally covered by other requirements, except at the higher operating frequencies (> 1 kc/s).

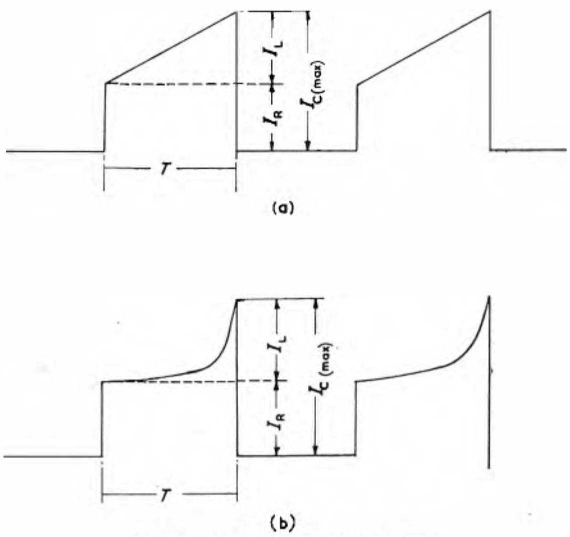

Fig. 2. Collector current waveforms
(a) Non-saturating transformer
(b) Saturating transformer

The time (T) for one half cycle of oscillation may be calculated[3] from the formula:

$$V_s = -N\,(d\phi/dt)\,10^{-8} \quad \ldots\ldots\ldots\ldots (2)$$

where V_s is the supply voltage
N is the number of turns
ϕ is the flux in the transformer core.

$$\int_0^T V_s dt = \int_{-\phi_s}^{+\phi_s} N d\phi \;.\; 10^{-8}$$

$$V_s T = 2N\phi_s \;.\; 10^{-8} \quad \ldots\ldots\ldots\ldots (3)$$

But $\phi_s = A \;.\; B_s$, where B_s is the saturation flux density in gauss and A is the cross-sectional area in square centimetres so that:

$$T = \frac{2NAB_s 10^{-8}}{V_s}$$

and therefore the frequency of operation is:

$$f = 1/2T = \frac{V_s \;.\; 10^8}{4 \;.\; NAB_s} \text{c/s} \quad \ldots\ldots\ldots\ldots (4)$$

Another condition that must be satisfied for proper operation is that there should be sufficient current available to saturate the core, having regard to the number of turns on the transformer and the length of the magnetic path.

Thus

$$H = \frac{1{\cdot}25 N I_L}{l} > H_o \quad \ldots\ldots\ldots\ldots (5)$$

where H_o has an arbitrary value obtained from the characteristics of the material used, I_L is the inductive current, and l is the magnetic path length in centimetres.

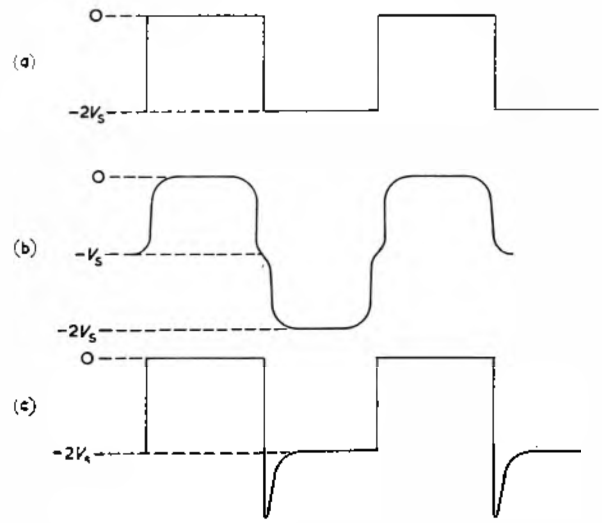

Fig. 3. Collector voltage waveforms
(a) Normal
(b) Insufficient residual inductance
(c) Primary not bifilar wound (insufficient coupling)

EFFICIENCY: POWER LOSSES

The main losses that occur in the circuit will be described since these must be minimized for high efficiency.

(1) *Transistor Loss*

This has three components, occurring during the 'on' and 'off' half cycles and during the transients. The loss during the 'on' period depends on the collector current, the bottoming voltage and the base voltage and current. That during the 'off' half cycle depends on the value of I_{co} at twice the supply voltage; at high temperatures this may be considerable. In addition, transient losses may be significant particuarly if the switching time is more than a very small fraction (1 per cent) of the cycle time.

(2) *Transformer Loss*

This has two components, copper loss and hysteresis loss. These are determined mainly by the size of the transformer, and are not entirely independent since a large transformer designed for low copper loss will have a high hysteresis loss and vice versa. There is therefore an optimum size of transformer for a given power handling capacity, although it is not generally very critical.

Supply Voltage Limitations

The peak voltage at the collector of each transistor is twice the supply voltage and occurs when the transistor is cut off. This would imply that the maximum supply voltage that may be used is one half of the collector diode breakdown voltage. In practice, however, the maximum supply voltage is limited to less than this. During the switching period a condition may occur in which the transistor is still conducting due to hole storage, but the voltage at the collector is approaching its maximum value due to the inductive load impedance; in such a case avalanche breakdown with high dissipation may cause transistor burn-out. It is important to note that this effect is most marked when the collector load impedance is almost purely inductive, i.e. under no-load conditions. It is advisable therefore to limit the peak voltage to the turnover voltage of the collector characteristics measured at high current with the base open-circuit.

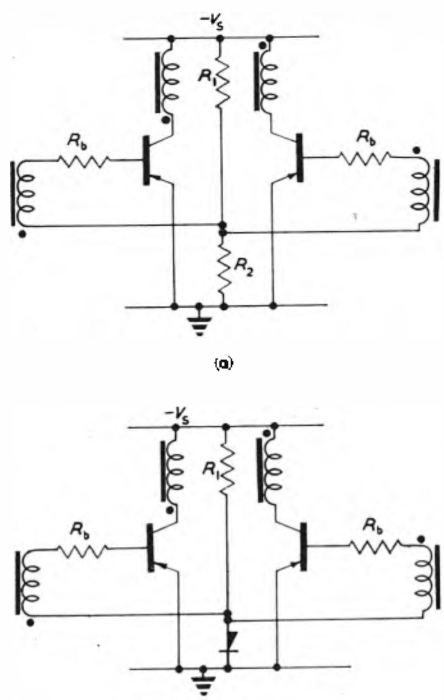

Fig. 4. Starting circuits

Where it is necessary to have a supply voltage in excess of that permitted for the normal push-pull circuit, a bridge circuit may be used[4].

Starting Circuits

The basic circuit of Fig. 1 will not necessarily oscillate when the supply voltage is applied, since both transistors are almost cut-off and the loop gain is less than one. It is essential therefore for the transistors to be biased into conduction initially. This can be arranged by means of a potentiometer chain to the bases as in Fig. 4(a). Initially the transistors are both biased on, but as soon as oscillations build up, the base current flowing through R_2 causes the mean voltage at the bases to go positive, and the circuit is self-biased into class-C operation. This method of starting is effective, but requires either a large amount of feedback giving high drive losses or a low value of starting resistor R_2 involving a large current drain through the potentiometer chain. Replacing the resistor R_2 by a diode (Fig. 4(b)) gives a high resistance when starting and a low resistance, with an almost constant voltage bias, when running, and thereby eliminates the compromise required for the value of R_2. The disadvantage of the diode circuit with respect to the resistance circuit is that in the event of failure to oscillate under overload conditions, the high resistance diode in the base circuits may cause thermal instability at high ambient temperatures. A resistor may be used to shunt the diode, thus gaining the best of both systems.

The values of the resistors required for starting may be calculated by considering the loop gain of the circuit with a given load.

With a total load R_L appearing across the primary of the transformer, the loop gain of the circuit is $g_t R_L/n$ where g_t is the transfer conductance for each transistor stage and n is the feedback turns ratio from the primary winding.

The g_t of each transistor stage is given by:

$$1/g_t = r_e + R_b/\alpha_0'$$
$$= \frac{0{\cdot}025}{I_e} + R_b/\alpha_0'$$

where I_e is the d.c. emitter current and R_b is the resistance in series with each base. For the loop gain to be greater than one:

$$1/g_t < R_L/n$$
$$\therefore \frac{0{\cdot}025}{I_e} < R_L/n - R_b/\alpha_0'$$
$$I_e > \frac{0{\cdot}025}{R_L/n - R_b/\alpha_0'} \quad \ldots\ldots\ldots\ldots (6)$$

But $I_e \simeq \alpha_0' I_b$

$$I_b > \frac{n}{40(\alpha_0' R_L - nR_b)} \quad \ldots\ldots\ldots\ldots (7)$$

In the case of the diode circuit the value of R_1 can be calculated approximately from

$$R_1 \simeq \tfrac{1}{2} \,.\, V_s/I_b \quad \ldots\ldots\ldots\ldots (8)$$

Where there is a resistor (R_2) between emitter and base, the minimum value of base emitter voltage V_{be} for the required I_b must be determined, then:

$$R_1 \simeq \frac{V_s}{2I_b + \dfrac{V_{be} + I_b R_b}{R_2}} \quad \ldots\ldots\ldots\ldots (9)$$

The value of R_2 is not critical, but should be of the order of tens of ohms.

When the load consists of a rectifier system with a reservoir capacitor, more bias may be required, and this will depend on the value of capacitance used and the leakage inductance of the transformer.

Overload Protection

If the circuit is subjected to overload, it may continue to oscillate out of bottoming, or cease oscillating altogether, but remain with full starting bias applied. Under these conditions, excessive dissipation may damage the transistors, and it may be considered advisable to incorporate some form of protection circuit to ensure that the unit switches off when subjected to overload.

This can most conveniently be done by monitoring the minimum collector voltage with two diodes as in Fig. 5. When the transistors come out of bottoming for any reason, the voltage at the control point c goes more negative and this change in potential may be amplified and used to operate a relay and disconnect the supply. (Fig. 5(b)). Where electromechanical relays are undesirable, transistors may be used to short-circuit the drive to the output transistors (Fig. 5(c)) and so render the oscillator inoperative. The third transistor in the circuit shown is used to hold the protection circuit inoperative for a short time when switching on.

When a protection circuit is used, a manual reset system must be incorporated.

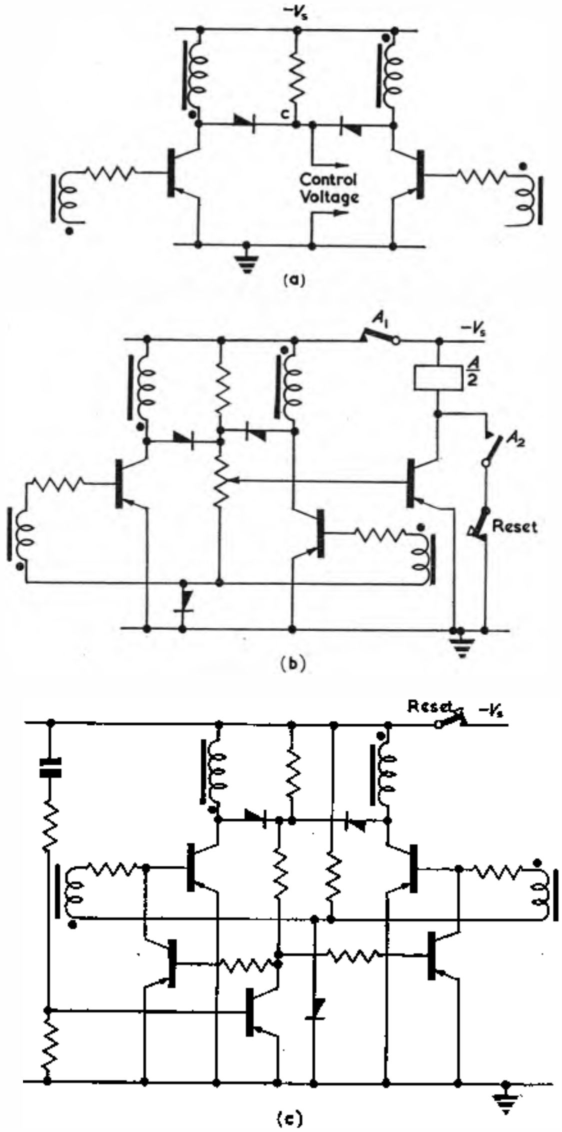

Fig. 5. Protection systems

Practical Design

Choice of Transistor Type

The type of transistor required for the circuit is determined primarily by the peak current required in the transistor. It must also be borne in mind that the maximum supply voltage may be limited by avalanche effects as well as by the collector diode breakdown voltage.

The maximum power dissipation of the transistor is usually of secondary consideration since the transistor is operating as a switch, i.e., alternately at low voltage, high current, and high voltage, low current.

Operating Frequency

There are several factors governing the choice of operating frequency, but none is very critical. The choice is primarily one between size and efficiency.

In general, transformer core losses and transistor transient losses increase with the frequency of operation, so the frequency should be kept as low as possible. On the other hand, the physical size of the transformer decreases as the operating frequency is increased.

A frequency of 400c/s has been found to be a good practical compromise between these two requirements.

Transformer Design

Since it is desirable for the switching action to be as fast as possible close coupling between primary and feedback windings is required. In addition, it is essential that the two halves of the primary be bifilar wound to achieve maximum coupling between them. If this is not done, high peak voltages will appear across the cut-off transistor (Fig. 3(c)) with the consequent danger of avalanche breakdown.

Since the saturation characteristics of interleaved E and I laminations are unpredictable due to the effect of the air-gaps, it is advisable to use transductor laminations or a toroidal core.

Primary Windings

Basically the transformer is designed to satisfy equations (4) and (5). The exact design procedure depends to a large extent on the particular application, since this will determine some of the parameters involved. In any case a certain amount of trial and error is involved to determine the optimum for any given application.

Assuming a required output power and supply voltage, the load current referred to the primary may be calculated, allowing 0·3V drop across the transistors when bottomed.

Having decided on the type of core material (e.g. H.C.R. alloy with $B_s \simeq 15$kG, $H_o \simeq 2$ Oc) and the frequency of operation, the product NA may be calculated from equation (4).

$$NA = \frac{(V_s - 0{\cdot}3)\,.\,10^8}{4\,.\,f\,.\,B_s} \quad \ldots\ldots\ldots\ldots \quad (10)$$

A possible lamination size should now be chosen. If t is the lamination tongue width and s is the thickness of the stack (allowing for the stacking factor), equation (10) becomes:

$$N\,.\,s = \frac{(V_s - 0{\cdot}3)\,.\,10^8}{4\,.\,f\;B_s\,.\,t} \quad \ldots\ldots\ldots\ldots \quad (11)$$

and from equation (5):

$$N\,.\,I_L > \frac{H_o \cdot l}{1{\cdot}25} \quad \ldots\ldots\ldots\ldots \quad (12)$$

From equations (11) and (12) a set of values of N and I_L for differing s may be obtained. Trial winding designs can then be made for practical values of s and I_L, and if necessary the process repeated using a different lamination size.

Feedback Turns

Having decided on the number of primary turns, the feedback windings must be designed to ensure that adequate current is available from the transistor at the minimum supply voltage. In addition, it must be ensured that with the supply voltage at its maximum the peak current in the transistor does not exceed the maximum rated current.

This is essentially a matter of spreads in the transistor characteristics, and the effects of these can be minimized by suitable choice of external base resistance R_b.

In addition, allowance must be made for the voltage dropped across the starting diode ($\simeq$ 0·4V) or the resistor R_2 whichever is applicable.

Starting Circuit

Having determined the required feedback turns ratio and the series base resistance values, there remain to be evaluated the resistors required to start the circuit oscillating under maximum load. These can be found from equations (7), (8) and (9).

It will be noticed that in the case of the resistor starting circuit, the required values of R_1 and R_2 depend on the feedback turns ratio (equations (7) and (9)) and that the feedback turns ratio depends on the values R_1 and R_2. Since the applied starting bias should have an adequate safety factor, it should not be difficult to achieve a design to satisfy both these conditions.

The values so obtained refer to a resistive a.c. load, and if a rectifier sytem with a reservoir capacitor is used, the resistance values required may be less than those calculated, and will depend on the leakage inductance in the transformer and the value of capacitor used. The use of *LC* smoothing filters also presents problems on starting, due to resonance in the filter.

Efficiency: Power Loss Calculation

The losses in the whole system can be itemized under various headings.

Transformer Loss

Hysteresis loss is proportional to the volume of the core and also increases with frequency of operation. The value may be determined from the characteristics of the material used. Copper loss can be determined directly from the resistance of the windings and the appropriate currents flowing.

Transistor Loss

In the 'on' condition, the power dissipated at the collector may be calculated from the product of the bottoming voltage and the r.m.s. collector current over the half cycle.

In the 'off' condition the power dissipated at the collector is the product $2V_s I_{co}$, where I_{co} is meaured at the peak voltage applied. This is not necessarily insignificant at high temperatures.

Transient loss occurs during both the switch-on and switch-off periods. Since the collector load is inductive, the current lags the voltage by an amount depending on the resistive load, so that the power lost in each transient will be load-dependent. The total transient loss is, however, a peak power of the order of $V_s . I_{pk}$ during the transient periods.

Some of the drive power is also dissipated in the transistor. From the transistor point of view this is not negligible in comparison with that at the collector and may be determined from characteristic curves, but from the point of view of overall efficiency, it is simpler to consider the drive power as a whole, including that in the series base resistance.

Feedback Loss

This can be calculated from the drive voltage from the transformer and the base current of the transistors.

Starting Circuit Loss

This is due to the presence of R_1 and is approximately V_s^2/R_1.

Overall Efficiency

In general the overall efficiency of a convertor working at a frequency of about 400c/s is approximately 80 per cent or even higher. A transformer core which is either too big or too small will considerably reduce this figure, and since most of the losses are independent of the load, the efficiency will be at its maximum under maximum load conditions.

Conclusions

Of the types of d.c. convertor which can be made, the one least affected by external influences is the square wave oscillator controlled by a saturating transformer.

For this type of operation, design formulae have been derived, giving relationships between supply voltage, transformer size and material and frequency of oscillation. The limitations on the power handling capacity of the circuit are determined by the maximum peak voltage and current ratings of the transistors used, and the spread in the transistor characteristics.

Acknowledgments

The authors are indebted to Messrs. E. Wolfendale and P. C. Dearlove and Dr. A. G. Milnes for their assistance in the investigation.

They also wish to thank the Director of Mullard Research Laboratories and the Directors of Mullard Ltd for permission to publish this article.

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New Varieties of Magnetic Materials

A Celebrity Lecture was given to the Permanent Magnet Association on 8 October by Dr. R. M. Bozorth of the Bell Telephone Laboratories.

In his lecture Dr. Bozorth discussed the properties of various types of magnetic material from several points of view.

He explained that the different kinds of magnetism are determined by the different kinds of interactions between the atoms of which the materials are composed. Materials are thus paramagnetic, ferromagnetic, antiferromagnetic, or ferrimagnetic. Three different kinds of interactions giving rise to ferromagnetism and ferrimagnetism were specially considered. Examples were given showing the characteristic bulk properties of each and it was pointed out that borderline cases, and materials in which several kinds of interaction can occur at the same time, make classification sometimes difficult or even pointless.

He stated that the relatively new non-metallic materials such as the ferrites have found many applications and are better understood than most metals, and that after the spinels (Ferroxcube), we know barium ferrite (Ferroxdur), and the garnets. The latter have the simplest structure, and substitutions of a great variety of different elements into the prototype yttrium iron garnet, have given rise to many new compounds of this type. So far, these have provided a further valuable insight into the nature of interatomic forces, but no commercial applications have yet been made. New kinds of ferromagnetic ferrites are continually being discovered: gadolinium ferrite types have been investigated, and galium ferrite has recently been discovered and is being investigated.

A distinctly new type of material is the ferromagnetic superconductor which exists at very low temperatures. These materials have negative initial permeabilities, and hysteresis loops of an entirely new shape. It is supposed that they are composed of domains of ferromagnetism alternating with domains of super-conduction, the latter having a magnetization anti-parallel to the applied field.

Dr. Bozorth concluded with a few remarks about the quality of permanent magnets, metallic and non-metallic, now being manufactured in America.

Some Aspects of the Logical and Circuit Design of a Digital Field Computer

By I. F. Brown*, and B. Meltzer†, Ph.D., F.Inst.P.

The principles of a new type of digital computer for the solution of field problems are described. By making the calculation at all the lattice points of the field simultaneous, this computer shortens computation time very greatly. A unitary (abacus) rather than binary or decimal arithmetic is used. The computation is done by lattice units, one associated with each lattice point of the field. The synthesis of lattice units from simpler basic units is described. The experimental design of a basic unit suitable for potential and other problems is presented. Some of the many outstanding problems of this computer are briefly referred to.

(Voir page 637 pour le résumé en français: Zusammenfassung in deutscher Sprache Seite 644)

A NEW type of digital computer for the solution of partial differential equations has recently been proposed[1]. This involves dividing the field of the independent variables into a lattice of points, and satisfying the finite-difference form of the equation at every lattice point of the field simultaneously, in contrast with conventional methods —e.g., the relaxation method—which deal with lattice points in serial order. By this means the time of the computation is reduced to the time taken to represent in pulse form the largest number handled, and so will be very short. Since, after restricted storage capacity lack of sufficient speed is the most serious limitation of modern digital machines when used on field problems, this proposal may provide an avenue of development for the future.

Unitary Arithmetic

For this type of computer, the use of decimal or binary arithmetic is practically out of the question since—as will be seen below—this would involve a vast number of carry circuits—in fact, one for each lattice point of the field—which would add enormously to the already considerable expense of the scheme. Furthermore, the binary system would be technically inconvenient for the arithmetical operations envisaged and discussed below.

A unitary arithmetic is therefore used, namely one in which the integral number N is represented by N pulses in serial order in time. This is a return to the most primitive, the abacus form of counting and—just as with the abacus—it is only the total number of pulses that matters, not their disposition in time. That is to say, a pulse followed by a gap followed by a pulse represents the number 2 equally well as does a pulse followed by a pulse.

Of course, unitary arithmetic would be out of the question in most conventional digital computers, since the duration in time of a single number would be far too long. But, as mentioned above, the computation in a field computer lasts effectively only as long as a single number, so this arithmetic is quite feasible.

Lattice and Basic Computer Units

To illustrate the principle of the proposed computer, consider the solution of Laplace's equation in two dimensions:

$$\partial^2 V/\partial x^2 + \partial^2 V/\partial y^2 = 0.$$

Its finite-difference form for a square two-dimensional lattice of points is

$$V_0 = \tfrac{1}{4}(V_1 + V_2 + V_3 + V_4),$$

which has to be satisfied at every lattice point, V_0 being the value of the dependent variable at the point and V_1, V_2, V_3 and V_4 the values at the four nearest lattice points.

One needs therefore to design a computer unit which, when fed with four pulse trains of respectively V_1, V_2, V_3 and V_4 pulses, produces continuously an output train of V_0 pulses,

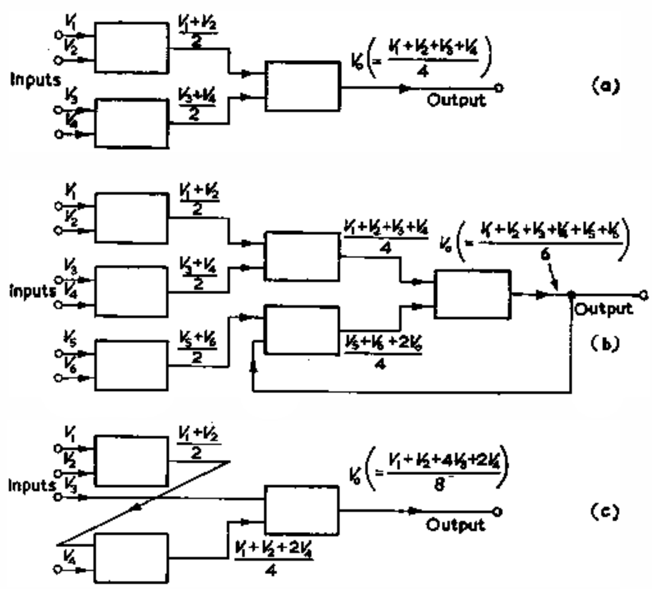

Fig. 1. The synthesis of basic units into lattice units
(a) Lattice unit for two-dimensional Laplace equation $\partial^2 V/\partial x^2 + \partial^2 V/\partial y^2 = 0$
(b) Lattice unit for full Laplace equation $\partial^2 V/\partial x^2 + \partial^2 V/\partial y^2 + \partial^2 V/\partial z^2 = 0$
(c) Lattice unit for $5(\partial^2 V/\partial x^2) + 3(\partial^2 V/\partial y^2) - 6(\partial V/\partial x) - 2(\partial V/\partial y) = 0$

V_0 being equal to the average of V_1, V_2, V_3 and V_4, to the nearest integer. Such a unit is designated a lattice unit. An interconnected square array of such identical lattice units, with output terminal of each connected to one input terminal of each of its four nearest neighbours, will solve the problem, if trains of pulses of the correct total number equal to the given boundary values of V are fed into the boundary units. All pulses throughout the computer would be synchronized from a clock pulse generator.

Now lattice units can be synthesized by a combination of simpler units, to be designated basic units. For example, consider a basic unit which has two inputs one consisting of a train of N_1 pulses the other of a synchronized train of N_2 pulses, and produces continuously a train of

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$\frac{1}{2}(N_1+N_2)$ synchronized pulses. Such a unit could be used for building up lattice units for a large class of field problems.

Three basic units connected as in Fig. 1(a) obviously constitute the lattice unit for the two-dimensional Laplace equation. Six connected as in Fig. 1(b) provide a lattice unit of six inputs $V_1, V_2, \ldots V_6$ and an output

$$V_0 = 1/6(V_1+V_2+V_3+V_4+V_5+V_6),$$

which is the finite-difference form of the general Laplace equation

$$\partial^2 V/\partial x^2 + \partial^2 V/\partial y^2 + \partial^2 V/\partial z^2 = 0,$$

so that a three-dimensional interconnected array of such lattice units would solve the general potential problem.

Another example, produced at random, is shown in Fig. 1(c). This lattice unit's input-output equation is

$$8V_0 = V_1+V_2+4V_3+2V_4,$$

which is the finite-difference form of the two-dimensional equation

$$5(\partial^2 V/\partial x^2) + 3(\partial^2 V/\partial y^2) - 6(\partial V/\partial x) - 2(\partial V/\partial y) = 0$$

It is clear that other basic units could be designed, each basic unit defining a class of fields, for the lattice units of which it can serve as a building brick.

The experimental work so far done has been confined to the design and testing of the 'averager' basic unit referred to above.

Design of Basic Unit

In order to keep the power consumption low, the mark-to-space ratio of the pulse trains used is kept as low as possible consistent with fast computation. The requirement is that the unit should give out a train of pulses, the number of pulses up to any instant being the mean of the number of pulses received on the two inputs up till then. There are three possible input conditions; no pulses received, a pulse received on one input channel, or a pulse received on both input channels. In the first condition, no pulse should be sent out. In the second, whether a pulse is sent out or not depends on whether a single pulse has been received earlier or not. If one has, and no pulse was sent out, then one should be sent out this time. If two pulses are received together, then a pulse must be sent out. There are two possible ways of operation. One is for the output pulse to occur one pulse interval after the corresponding input pulse. In this method, it would be easy to maintain the size, shape and timing of pulses throughout the computer, but the computation would not be as fast as possible. The other method would be to have the output pulse coincident with the corresponding input pulse. This has the disadvantage that it is very difficult to maintain pulse shapes, and in a large computer it might be impracticable. However, this is the system which was used in the experimental circuit. There is no reason to believe that the circuit for the units using the first system would be any more complicated than the one actually developed. In order to simplify the practical problems, it was decided to

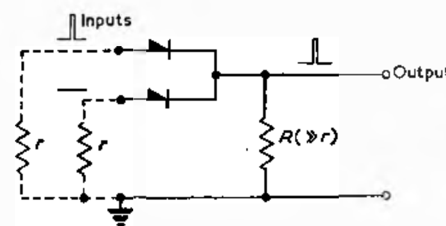


Fig. 2. 'Or' circuit

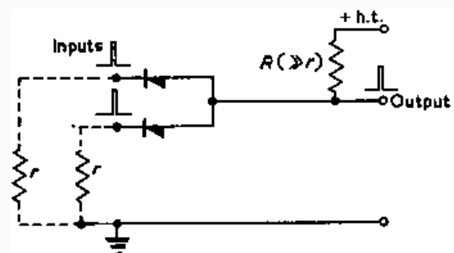
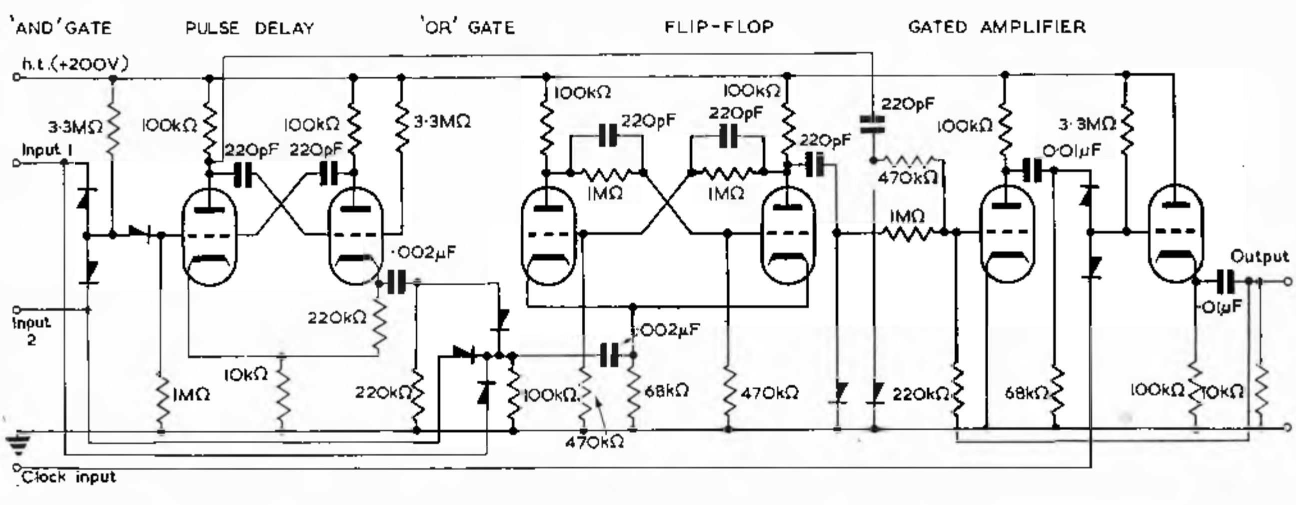


Fig. 3. 'And' circuit

use thermionic devices in the unit, although it was realized that a large computer would have a prohibitively high power consumption, and in practice semiconductor devices would have to be used throughout.

The circuit details can now be considered. The basis of the unit was a bistable circuit, or flip-flop, which gives out a pulse when changing from state A to state B but not when changing from B to A. It was followed by an output amplifier which is gated to inhibit any extraneous pulses. The rest of the unit is arranged to ensure that the three basic requirements are met. The two input lines are fed into an 'or' circuit, shown in Fig. 2, which gives out a pulse whenever a pulse is received on one or both inputs, without allowing a pulse received on one input to be fed out back to the other. The output pulse from the 'or' circuit operates the flip-flop, and each pulse or simultaneous pair of

Fig. 4. Complete circuit of basic unit



pulses received changes the state of the flip-flop, so giving a half-rate output. This would be correct if simultaneous pulses could not occur, but they can; the circuit must give an output pulse when they do, and the flip-flop, after being operated by them must be returned to its original state. To perform this, the two inputs are fed to an 'and' circuit, shown in Fig. 3, which gives an output pulse only when a pulse is fed in on each input. This is used to operate a monostable circuit, from which are derived two outputs; the first a pulse coincident with the input pulse and the second a pulse delayed by about half the standard pulse interval. The first is passed straight to the output amplifier, and provides the required output. The second is fed into the input 'or' circuit and so operates the flip-flop, if returning it to its original state. If the flip-flop was in state A when the two simultaneous pulses were received, then a pulse is produced coincident with the input, and this pulse as well as the one from the monostable circuit both get fed through to the output amplifier simultaneously; but this does not affect the amplitude of the output pulse at all as the output amplifier is normally driven into overload. If the flip-flop was in state B, it will initially be changed to state A, and the delayed pulse will then change it back to state B, so producing an output pulse. This pulse is spurious since the output has already been provided by the monostable circuit, but it is not accepted by the output amplifier because it is gated by the clock pulse generator. The diagram of the complete circuit is shown in Fig. 4 and does not require detailed description, the various component parts of it being fairly standard. The circuit components were chosen for a maximum pulse repetition rate of 3kc/s.

This circuit contains six triodes, and in one lattice unit six of these circuits are required, making a total of thirty-six valves. If the minimum sized lattice that would be useful is 10 by 10 by 20 this means that seventy-two thousand valves are necessary as a minimum. Probably one watt is required for each heater, so 72kW of heater power would be needed, with probably about 10kW of h.t. power. By the use of transistors, the total power consumption of the basic unit could probably be reduced to 50mW, giving a minimum total of 600W. As already mentioned, a large computer would have to have a basic unit which allowed one pulse interval of time for the computation, so lowering the overall speed, though only by a very small amount. The circuit required to do this would be of the same complexity as the one described here.

Conclusion

A computer working on this principle would be a practicable possibility, but the many problems concerned with programming the inputs and storing and reading out the answers have yet to be solved. They are considerable, as the number of inputs and outputs is so large. The crucial problem of reading and accuracy has been briefly discussed elsewhere[1]. Each boundary to the field will require a pulse train generator, and since there might be a considerable number of separate boundaries a simple pulse train generator should be developed; Dekatron tubes can very conveniently be used for generating trains of any length with absolute accuracy, and might find other applications if the pulse rates are kept quite low.

REFERENCE

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Charts for Deriving Transistor r-Parameters from h-Parameters

By G. W. E. Stark*, B.A., A.M.I.E.E.

Charts are described whereby the inference of the values of the r-parameters of a transistor from its measured grounded-base h-parameters is simplified.

(Voir page 637 pour le résumé en français: Zusammenfassung in deutscher Sprache Seite 644)

FOR most low-frequency circuit applications of transistors, the values of the r-parameters, r_e, r_b and r_c are of greatest use. Direct measurement of these, however, is not so straightforward as measurement of the hybrid set of so-called h-parameters.

The two are related by simple formulae, but it is convenient, particularly if the transformations have to be made frequently, to depict the relationships graphically.

Description of Charts and their Use

The h-parameters are the elements of the H matrix in:

$$\begin{bmatrix} V_1 \\ i_2 \end{bmatrix} = \begin{bmatrix} H \end{bmatrix} \begin{bmatrix} i_1 \\ V_2 \end{bmatrix}$$

where i_1, i_2, V_1 and V_2 are the currents and voltages at the input and output terminal pairs of the four-terminal network representing the transistor. They can be defined as:

		Grounded base nomenclature
h_{11}:	the input impedance, output short-circuited	h_{ib}
h_{12}:	the reverse voltage feedback ratio, input open-circuited	h_{rb}

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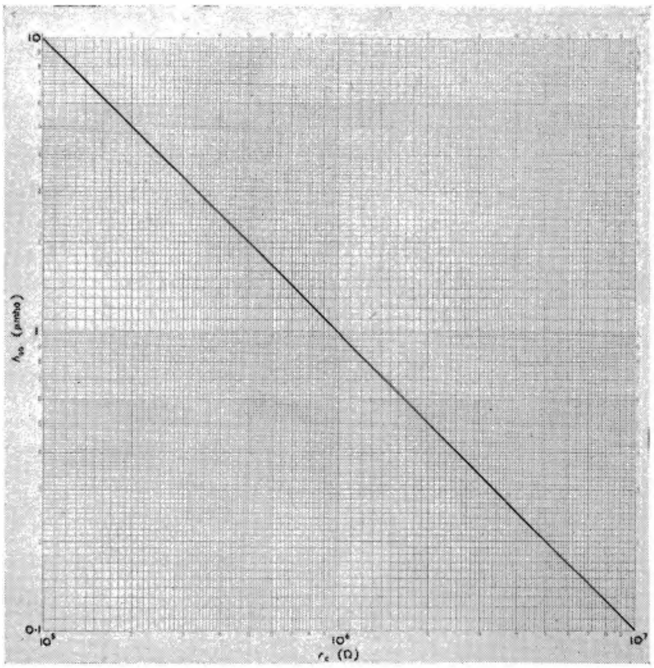

Fig. 1. Relationship of h_{ob} to r_c

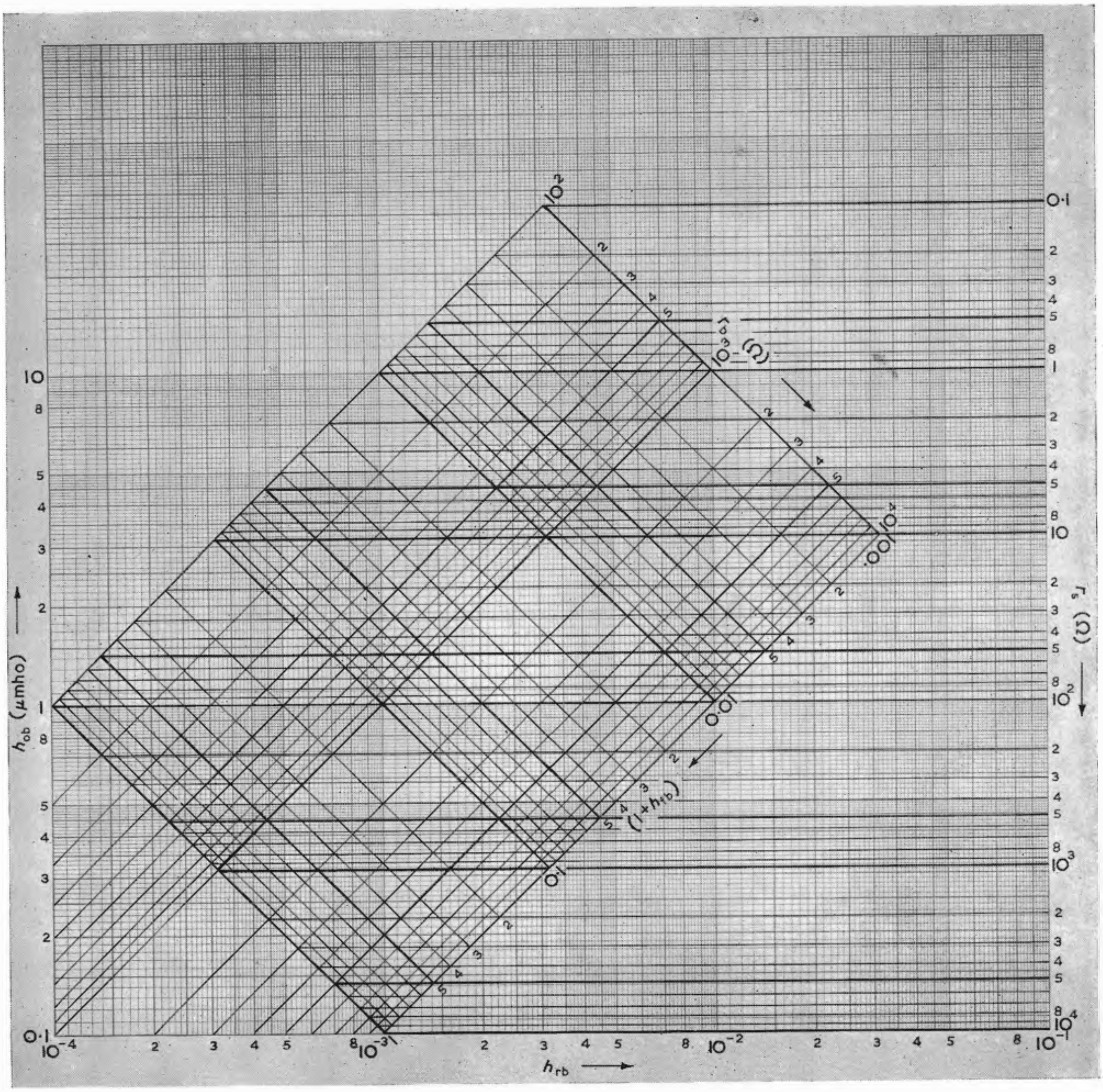

Fig. 2. *r* **parameter/*h* parameter chart for transistors**

h_{21}: the forward current gain ratio, output short-circuited .. .. h_{fb}

h_{22}: the output admittance, input open-circuited .. .. .. h_{ob}

Since these are related to circuit conditions, there is a set for each of the three circuit configurations. The grounded base set only are of concern here. They are approximately related to the *r*-parameters, the components of the low-frequency *T*-equivalent circuit, by the following expressions:

$$r_c = 1/h_{ob}$$

$$r_b = h_{rb}/h_{ob}$$

$$r_e = h_{ib} - \frac{(1 + h_{fb})h_{rb}}{h_{ob}}$$

The first of these relationships is depicted in Fig. 1, which is self-explanatory. Fig. 2 shows the derivation first of r_b and then, by way of a subsidiary parameter r_a, of r_e; it is used thus:

First, find the intersection of the appropriate h_{rb} and h_{ob} co-ordinates, reading the value of r_b on the scale which is at 45° to the h_{rb} and h_{ob} axes. Next, find the intersection of the r_b and $(1 + h_{fb})$ co-ordinates in the smaller section of the chart. Read off the value of r_a from the scale at the right-hand side of the main chart. Finally, subtract this value of r_a from the known value of r_b to give r_e.

Acknowledgment

Acknowledgment is made to the Controller, H.M.S.O. for permission to publish this article.

An Investigation into Some Aspects of Diode Quantizing Circuits

By H. V. Bell*, B.Sc., and W. Alexander*, M.Sc.(Eng.), Ph.D., M.I.E.E.

Quantization is defined and some past work on diode quantizers is reviewed. Three circuits are compared both theoretically and by measurement, and the results presented. A possible application of these circuits is then described.

(Voir page 637 pour le résumé en français: Zusammenfassung in deutscher Sprache Seite 644)

THIS article deals with an investigation into some diode quantizing circuits.

Quantization is the process in which the range of values of a wave is divided into a finite number of smaller sub-ranges, each of which is represented by an assigned value within the sub-range[1]. This value may be that corresponding to the mean of the sub-range, or some other, arbitrarily assigned, number.

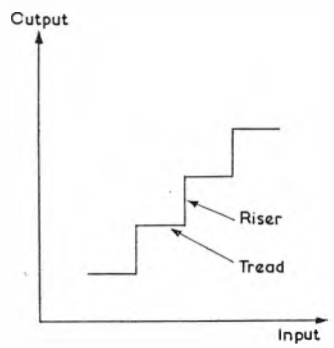

Fig. 1. Transfer characteristic of quantizing circuit

In practice, any device having a transfer characteristic of the form shown in Fig. 1 may be used. This 'staircase' characteristic allows the approximate representation of any continuously varying function by a set of discrete values, i.e., it 'quantizes' the input signal. This feature is employed as a means of displaying the characteristic on an oscilloscope. The analogy to a staircase is also useful when referring to the constituent parts of the characteristic, the horizontal portion being known as the 'tread' and the vertical portion as the 'riser'.

SYMBOLS USED

N	=	Number of quantization levels
n	=	Number of diode pairs
R_v	=	Equivalent vertical resistance
R_h	=	Equivalent horizontal resistance
R_f	=	Forward resistance of a diode
R_b	=	Back resistance of a diode
Z_s	=	Source impedance of a bias supply
R_1, etc.	=	Current bias supply series resistors
Z_o	=	Output impedance of cathode-follower
R_k	=	Cathode-follower load resistance
r	=	Riser under consideration
p	=	Number of diode sections 'switched' so that current flows in the top diode

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Past Work and Some Possible Circuits

Past work on the subject of quantizing networks using germanium diodes was carried out by Tootill[2,3] in England and by Henle[4] in America. Working without any knowledge of each other, they separately produced the circuit shown in Fig. 2(a) and used it as the basis of a multi-stable state device. Tootill also suggested another circuit, Fig. 2(b), but did not use it because of the number of floating voltage supplies required.

The operation of the circuit of Fig. 2(a) can be seen

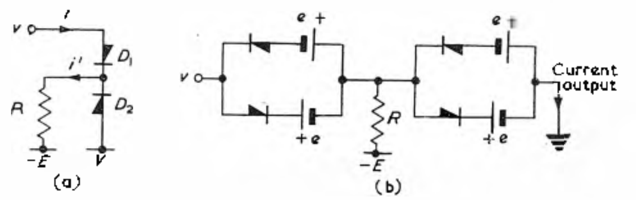

Fig. 2 (a). Basic element for connecting in parallel with further elements (b). Two elements of the basic series configuration

by considering various values of input voltage. At a value of input voltage, v, less than the battery voltage, V, D_1 is cut off and D_2 conducts, being provided with bias current from the effective current source consisting of the large negative supply voltage, E, and the high valued resistor, R. As v is increased above V, D_1 will conduct and D_2 will be cut off, resulting in a transference of current flow from one diode to the other.

In consequence, the current into the network, i, (originally zero), will increase rapidly to the value of the bias current, i'. Fig. 3(a) depicts the characteristic of such an element and shows the transition, or low resistance portion over which the transference of current flow occurs.

This forms the basis of the staircase characteristic: putting further elements in parallel with this one, and providing increasing bias voltages of V, $2V$, $3V$, etc., will complete the circuit and give the desired overall characteristic. It is seen, then, that the number of quantization levels, N, from a device of this configuration having n diode pairs, is $N = n + 1$.

With this circuit it is also possible to utilize the current flowing to earth from the network as the quantized version of the input, instead of the input current. This case is considered later, when the theory of the network is discussed.

Considering now the series type of circuit of Fig. 2(b), the mode of operation is slightly different. The characteristic of the single section is depicted in Fig. 3(b) and it is seen that there are two low resistance portions, where one of the diodes is conducting, separated by a high resistance section, where neither diode conducts.

To produce the full characteristic, these elements are connected in series and are supplied with bias current from the large negative voltage, E, and the high valued resistor, R, as in Fig. 2(b). Again, in this circuit, the current flowing to earth is used to provide the quantized version of the input. The number of quantization levels, N, possible with this configuration, is equal to the number of diode pairs, n.

Comparison on Theoretical Basis

In practice, it is found that those portions of the characteristics of these networks previously assumed horizontal and vertical are not, in fact, so. The departure from the desired form is termed the 'degradation' and as the characteristics are essentially current-voltage relationships, the degradations can be expressed as resistances.

In the paper referred to previously[2], the current was fed from a cathode-follower. The following expressions for the degradation were given for this circuit, R_v being the equivalent vertical resistance, and R_h the equivalent horizontal resistance.

Vertical:

$$R_v = Z_0 + 2R_f + (r - 1)Z_s \text{ for the } r^{th} \text{ riser} \quad \ldots\ldots (1)$$

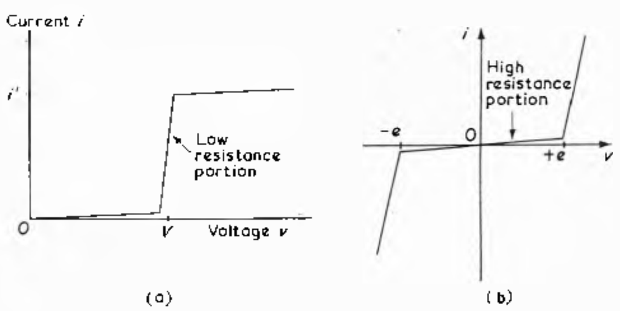

Fig. 3 (a). Current-voltage characteristic for parallel element of Fig. 2(a) (b). Current-voltage characteristic for series element of Fig. 2(b)

Horizontal:

$$(1/R_h) = (1/R_k) + ((1/R_1) + (1/R_2) + \ldots(1/R_n)) + (n/R_b) \quad \ldots\ldots\ldots\ldots (2)$$

In order that a comparison could be made, on a theoretical basis, between this circuit and the two others described above, the following expressions for the equivalent resistances were deduced.

For the modified original circuit, utilizing the current to earth as the quantized version of the input,

Vertical:

$$R_v = 2R_f + (r - 1)Z_s \quad \ldots\ldots\ldots\ldots\ldots (3)$$

Horizontal:

$$(1/R_h) = (n/R_b) \quad \ldots\ldots\ldots\ldots\ldots\ldots (4)$$

For the series circuit with the floating voltage bias supplies,

Vertical:

$$R_v = n(R_f + Z_s) \quad \ldots\ldots\ldots\ldots\ldots (5)$$

Horizontal:

$$(1/R_h) = \tfrac{1}{2}(R_b + 2(n - 1)R_f) \quad \ldots\ldots\ldots (6)$$

To make any real comparison between the circuits, it is necessary to study the operating conditions of the crystal diodes and so decide on the value of R_f which is to be taken. A curve was available showing the mean, measured characteristics of some 20 to 30 diodes of the CV448 (GEX54) type selected at random. It is not possible to reproduce this curve here, but those values of main interest are shown in Table 1. A value of 500kΩ was taken for the back resistance, R_b, this being the value with a reverse voltage of 50V. At lower reverse voltages, R_b is much higher in value, e.g. at −10V, R_b = 1MΩ.

The desired conditions are that $1/R_h$ and R_v shall be zero and so comparisons between the circuits can be made by considering how near the various values approach this desideratum. When inserting figures, it is generally more convenient to consider the horizontal degradation in terms of a resistance, rather than a conductance. In each case, a value of $n = 4$ is taken.

(a) *Original Circuit.* Equation (1) gives:

$R_v = (594 + 3Z_s)$ ohms, for the worst condition, i.e. the fourth riser.

Equation (2) gives:

$R_h = 24{\cdot}2$kΩ, again for the worst condition, i.e. the fifth tread.

(b) *Modified Original Circuit.* Equation (3) for the vertical resistance is similar to equation (1) but without the terms associated with the cathode-follower. This gives:

$R_v = (250 + 3Z_s)$ ohms for the worst case, i.e. the fourth riser.

Also $R_h = (R_b/n) = 125$kΩ.

(c) *The Series Circuit.* For this circuit, equation (5) gives the value:

$R_v = (500 + 4Z_s)$ ohms.

In equation (6), the value of $6R_f$ obtained is 750Ω, and if this is considered small, compared with the back resistance, then:

$$R_h \simeq 250\text{k}\Omega$$

It is difficult to estimate what values of forward resistance should be used owing to differing currents flowing through the various diodes. However, the variations from the above values should not be too great. In all these circuits, the resistors supplying the current bias will have to be graded in value to allow for the presence of the bias batteries and thus give a constant average slope to the staircase. Also, in order that the current bias sources may approximate to true current generators, then to achieve the requisite current, both the voltage and the series resistor must be high in value.

It is seen from this comparison that the second circuit gives the lowest value of equivalent vertical resistance, with a reasonably high value for the horizontal equivalent resistance. However, the series circuit, (c), with its very high value of equivalent horizontal resistance would be more suitable for any application where constant outputs at each of the quantizing levels is required.

Experimental Work and Results

Some measurements of the degradation had been carried out on the original circuit before the above comparison was made. This circuit had been built up, using CV448 germanium diodes, to have four sections, resulting in five levels of quantization. The battery potentials used were approximately 27, 18 and 9V and current bias was supplied from a 300V negative supply in conjunction with graded resistors of value approximately 150kΩ. The voltage

TABLE 1.
Diode Characteristics

CURRENT THROUGH DIODE (mA)	FORWARD RESISTANCE (Ω)
0·4	750
1·0	156
2·0	125
3·0	113
4·0	100
4 to 14	about 75

applied to the network and the current flowing into it were measured and the results plotted on a graph. This graph then displayed the transfer characteristic of the network and measurements of the degradation could be made. A comparison between the measured and calculated values for the horizontal equivalent resistance is given in Table 2. This shows some agreement between the two, the resistance decreasing with an increase in the number of levels, as the theory would suggest. The indication here is that the horizontal equivalent resistance for, say, ten quantization levels would probably be intolerably low.

A further disadvantage is that the vertical degradation, from equation (1), also varies. The graph does seem to confirm this, although it was not possible to take many readings on this portion of the characteristic.

It was then decided to build up the series circuit and make similar measurements. For the preliminary investigation, a small panel was made up, carrying terminal blocks

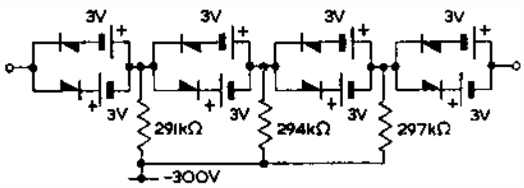

Fig. 4. Circuit of four-level quantizer

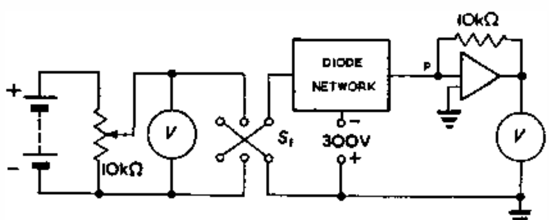

Fig. 5. Circuit used for all measurements

to which the germanium diodes could be attached. Four diode pairs were used, resulting in the device having four levels of quantization. To define the voltage intervals, 3-volt batteries were used, the tread width then being 6V.

A 300V negative supply was available and so series resistors were used in conjunction with this to give a current bias of 1mA. The series resistors were graded to take into account the battery voltage bias supplies, their values being marked on the circuit diagram, Fig. 4. The desired values were obtained by taking nominal 5 per cent tolerance high stability wire-wound resistors and measuring their resistance on a Wheatstone bridge. The deviation of the resistors from the desired values was made less than $\pm 0{\cdot}1$ per cent in all cases, it being necessary, for some, to make up the values from two smaller resistors in series.

The circuit with which all the subsequent measurements were taken, is shown in Fig. 5. Since the quantized version of the input to the network is the current flowing to earth, it was decided to use a 'chopper' stabilized d.c. amplifier to convert the current to a voltage. The use of the d.c.

TABLE 2

Comparison of Measured and Calculated Values of Resistance

TREAD NUMBER	MEASURED RESISTANCE (kΩ)	CALCULATED RESISTANCE (kΩ)
1	40	35·5
2	33	26
3	28	24·2
4	23	21

amplifier with a feedback resistor as shown, gives a virtual earth at point P, and the output voltage is proportional to the output current of the quantizer.

It was subsequently found desirable to be able to display the full characteristic of the network on an oscilloscope.

As mentioned in the first section, the network will, by a set of discrete values, represent approximately any applied voltage. If this voltage increases linearly with time, the effect will be a representation of the transfer characteristic of the quantizer, time being used as the parameter. Taking the time-base voltage from an oscilloscope proved to be the most convenient method of obtaining this voltage, the circuit being given in Fig. 6.

The amplitude and d.c. level of the input signal were adjusted by means of two potentiometers in conjunction with a conventional adder circuit. A sweep frequency of the order of 20c/s was used.

First measurements on the quantizer showed that two of the treads were distorted in a random manner. The transfer characteristic, when displayed on the oscilloscope, showed

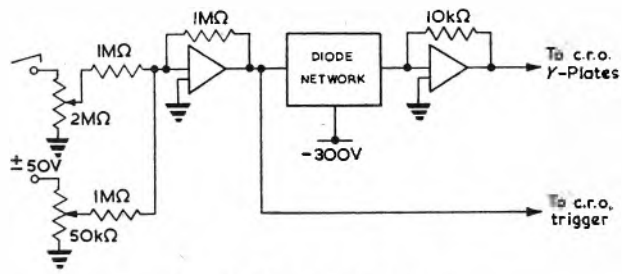

Fig. 6. Visual display circuit

the presence of high-frequency oscillations on these two treads. This instability was removed by placing a 15pF capacitor in parallel with the feedback resistor, thus modifying the amplifier transfer function for the higher frequencies.

Measurements were then continued and a graph drawn of the transfer characteristic from which it was possible to obtain values for the riser equivalent resistances. These showed a variation between about 550 and 1 000Ω, due to each diode carrying a different current and hence having a different effective resistance. If the variation of diode resistance is allowed for in the appropriate expression for the degradation, fairly close agreement is obtained between measured and calculated values.

No measurement of the horizontal equivalent resistance was practicable, since the change of output voltage on the treads was too small to be observed on the normal voltmeter connected at output. A conservative estimate for the horizontal resistance, based on the accuracy to which the output meter could be read, would be 100kΩ, although it seems likely that the value was higher than this.

It should, of course, be realized that on the risers, when the quantizer acts as a low resistance, there is an effective amplification of about 20, whereas on the treads, with the quantizer having a high resistance, there is an attenuation. In view of this, the change of output on the treads will be smaller than would be expected.

A photograph of the characteristic taken from the display oscilloscope, is shown in Fig. 7.

Some considerable saving in bulk could be made by dispensing with the batteries. Diodes of the CV2290 (GEX66) type have a characteristic such that at voltages below 0·3V, very little current flows, whereas above this voltage, there is a relatively high current flow. Putting two of these in parallel would produce the step characteristic required without the use of batteries, the width of the step being governed by the diodes themselves.

With this in mind, a circuit was built up using these diodes and having four levels of quantization. Once again,

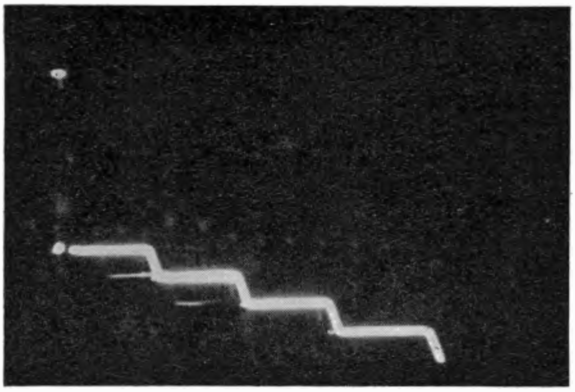

Fig. 7. Photograph taken from display oscilloscope of four-level quantizer characteristic

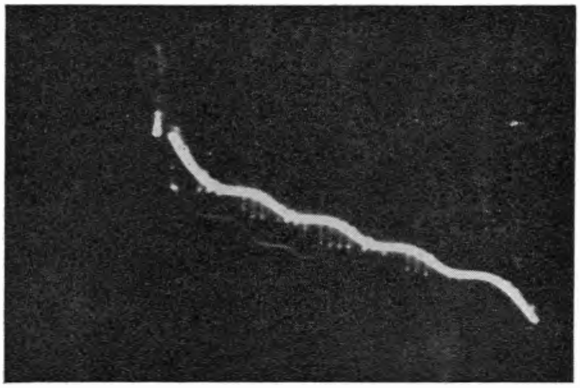

Fig. 8. Display obtained for circuit using CV2290 diodes

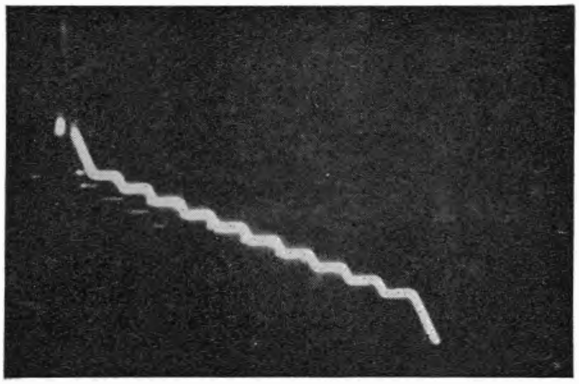

Fig. 9. Display of ten-level quantizer characteristic

1mA current steps were chosen, given by a 300V supply and 300kΩ series resistors. In this case, grading of the series resistors, again chosen to within ±0·1 per cent, was not necessary.

It was seen from the oscilloscope display that the characteristic was extremely degraded and this was borne out by the measurements and graph. From the graph, values of vertical equivalent resistance ranging between 250 and 500Ω were obtained. These values are rather less than for the previous circuit, but the degradation of the treads and the rounding of the step edges is very apparent. A photograph of the trace is shown in Fig. 8.

In the course of these measurements, two of the diodes were changed and the graph re-plotted. Due to this change, considerable increase of one of the step widths occurred indicating that the overall staircase characteristic was too dependent upon the individual diode characteristics. In order to obviate this difficulty, selection of the diodes would be necessary and would outweigh any advantage gained by dispensing with the batteries.

Ten Level Quantizer

Some further experiments were conducted on a multi-level quantizer. The series circuit was employed, some thought being given to reducing the size and weight of the quantizer and also to finding a more practical method of obtaining the required values for the series resistors. It was

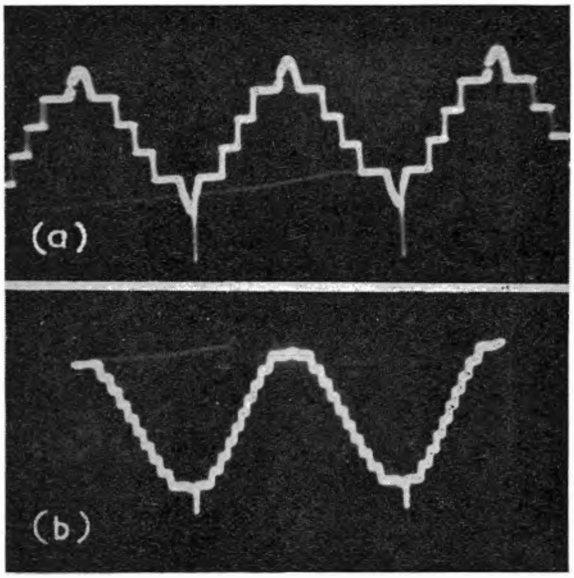

Fig. 10 (a). Sine wave quantized by four-level quantizer, and (b). by ten-level quantizer

decided to use ten levels which demanded the use of 20 germanium diodes, these being of the CV448 type, as before.

The smallest tread width that can be used is 3V, formed when 1·5V cells are used for the bias batteries. The manufacturers' lists were studied and a suitable miniature cell was found.

The series resistors were made up of a suitable value of high stability, close tolerance fixed resistor in series with a variable resistor. In this way, the total resistance could be adjusted to the desired value instead of having to select the resistors to ± 0·1 per cent as previously. At no time did the device appear sensitive to mechanical vibration. This procedure represents a more practical method of approach.

A further saving in space resulted from the use of a tag-board to which the germanium diodes could be soldered. Provided that no excessive heat reaches the diodes during the soldering process, their characteristics will not be altered.

Measurements made on this circuit showed it to have values of equivalent vertical resistance of from 1·6 to 2·0kΩ, as would be expected. The horizontal equivalent resistance was again unobtainable as there was no apparent change of output on the horizontal sections. Fig. 9 shows the charac-

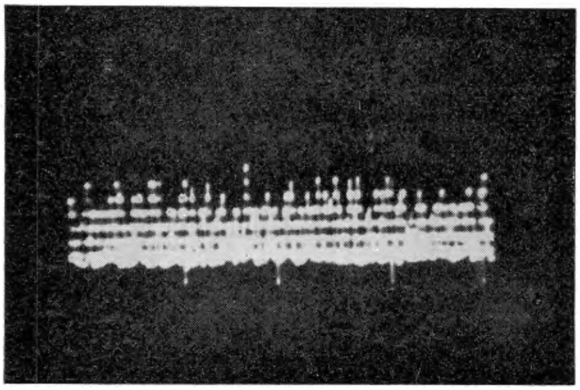

Fig. 11. Display obtained with speech applied to ten-level quantizer

teristic as it appeared on the display oscilloscope, and it is seen to be clean and free from any traces of instability.

Conclusions and Suggestions

The experimental results obtained in the course of the work have agreed very well with the theory that was developed. This theory is, of necessity, not applicable very rigidly in practice, mainly due to the non-linear characteristics of the germanium diodes themselves. It did, however, serve as a basis upon which a comparison of the circuits could be made.

The concept of quantization was first introduced in connexion with pulse modulation of a radio frequency wave. If the pulse spacing is kept constant and the amplitude is modulated, the result is a means of conveying information.

Increasing the number of quantization levels will increase the faithfulness with which the signal is reproduced. This is demonstrated in Fig. 10, where a 50c/s sine wave is shown quantized (a) by the four-level device and (b) by the ten-level device.

With a speech waveform, however, it has been found necessary to use only 64 or 128 levels, if these are graded logarithmically. This enables weak passages to be transmitted with nearly the same percentage precision as loud passages and so gives reproduction with quite a good degree of fidelity.

A photograph, (Fig. 11), is included showing the result of speech from a broadcast receiver being applied to the quantizer. Although only ten levels are used, the quantizing action can be seen clearly. The levels are not graded logarithmically and the photograph emphasizes the necessity for such grading. It does seem that the use of the circuit for this purpose is feasible; further investigation may show whether it has any advantage over present systems.

Considering the various equations for the present form of circuit, it seems that some real improvement could be effected only if a diode type was obtainable having rather different characteristics. The requirement in this respect is to have a lower forward resistance (say about 15Ω) while retaining the high back resistance necessary to give a constant output on the tread.

The circuits proved to be of interest, but it remains to be seen whether they will find wide practical application.

Acknowledgments

The experimental work described here was carried out by one of the authors (H. V. Bell) as part requirement of a first degree at the University of Nottingham. The other author supervised the work.

The authors wish to acknowledge their indebtedness to Mr. G. C. Tootill of R.A.E., Farnborough, for his help and advice during the early stages of the project.

Thanks are also extended to Professor J. E. Parton, Head of the Electrical Engineering Department of the University of Nottingham, for his interest in the work and for providing facilities.

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Sweden's Radar Defence System

Sweden is to have an air defence system embodying new and secret automatic electronic techniques which, it is claimed, will make it far more effective than any other in existence today. It provides for instantaneous action against enemy attack.

Although security forbids a detailed description, it can be said that the heart of the system is a very high speed computer which solves a large number of interception problems simultaneously and enables the defence weapons—fighters, guided missiles and other anti-aircraft devices—to be brought into action at precisely the right instant. Black-and-white and colour television and automatic information-dissemination techniques also play an important part in the system.

The Swedish Air Defence System has been evolved by scientists and engineers of Marconi's Wireless Telegraph Company of England and their opposite numbers in the Swedish Air Force, working in the closest collaboration. The initial contracts for the design and supply of the electronic equipment, which are to the approximate value of £1 500000, have been awarded to Marconi's via their Swedish associates Svenska Radioaktiebolaget.

The critical problem of modern air defence is that of initiating action instantaneously. In this age of ballistic missiles and supersonic aircraft the methods which served so well in World War II—the telephoning of information from the various radar sites to a Control Room which in turn informed the various defence authorities, also by telephone—are now not nearly fast enough. In the Swedish Air Defence System radar equipments of various types are used to detect and identify enemy raids, providing full positional and height information, while automatic tracking methods are employed to render the system proof against disorganization by saturation attacks. The relevant data is stored in an electronic 'memory' or 'information bank,' using digital techniques, and is continuously revised and brought up to date; the computer uses this information in its interception calculations and reduces the human element, with its fallibility in judgment, to a minimum.

Whereas, in the past, Executive Officers have been handicapped by delays in obtaining vital operational information and by having a far-from-complete picture of what was going on, the new system provides special tabular and synthetic displays supplemented by televised pictures of ancillary operational data (weather charts, etc.). Senior executives are able to make their assessment of the air situation from large-screen colour television projection displays.

Effective co-ordination of all types of defence weapons is achieved by vesting the overall responsibility in one Senior Control Executive who, via his specialist staff officers, can allocate targets for destruction by the most appropriate weapon. This welding of all elements of air defence is one of the primary advantages of the extremely comprehensive radar data handling system devised by Marconi's.

An Electronic Tachometer for Automobiles

By M. J. Wright*, B.Sc.

An electrical signal having a fundamental frequency proportional to engine speed is obtained from the automobile ignition system. A novel circuit, which combines filtering and limiting functions, removes undesired oscillations and feeds a fixed amplitude square wave to a diode pump frequency measuring circuit which is calibrated directly in terms of engine speed.

The complete circuit is simple, contains no active components and is powered by the automobile battery. The calibration is little affected by changes in either the battery voltage or ambient temperature.

(Voir page 637 pour le résumé en français: Zusammenfassung in deutscher Sprache Seite 644)

An electrical tachometer has the advantage of ease of installation over its mechanical counterpart. This is particularly desirable in a laboratory instrument in order that engine speed measurements may be made on any vehicle at short notice.

The tachometer to be described measures the fundamental frequency of an electrical waveform obtained from the automobile ignition circuit, this frequency being proportional to engine speed. The waveform is applied to a simple circuit which combines filtering and limiting functions and feeds a fixed amplitude square wave to a frequency measuring circuit. A milliammeter indicates engine speed directly.

Before describing the instrument itself the operation of the electrical ignition system is briefly explained, followed by a description of the relevant ignition waveform.

Electrical Ignition System

The wiring diagram of a typical ignition system is given in Fig. 1. A positive earth is shown since this is the standard practice with British car manufacturers.

The Engine Tachometer

With the contact-breaker closed the storage battery is applied to the ignition coil primary winding producing an exponential build-up of primary current. When the contacts open the primary circuit is broken and the flux associated with the iron core collapses. The high voltage induced in the multi-turn secondary winding is fed via the distributor to the appropriate sparking plug where electrical breakdown of the air-gap occurs.

Since the contact-breaker is driven by the engine, the ignition circuit provides an electrical signal whose fundamental frequency f is proportional to engine speed. For four-stroke engines, the relation is

$$f = (s/60) \times (N/2) \quad \ldots\ldots\ldots\ldots\ldots\ldots (1)$$

$f =$ contact-breaker frequency (c/s)
$s =$ Engine speed (rev/min)
$N =$ number of cylinders.

* *Joseph Lucas (Electrical) Ltd.*

Ignition Coil Primary Voltage Waveform

The voltage waveform across the ignition coil primary winding is shown in Fig. 2. With the contacts closed, the full battery voltage V_B is applied to the ignition coil primary winding. When the contacts open, the voltage across the winding reverses (back e.m.f. due to collapse of flux). It is the main function of capacitor C_P to restrict the rate of increase of voltage across the contacts as they are opening and so prevent arcing. After a short time delay (caused by the secondary winding stray capacitance) the induced secondary voltage causes breakdown of the sparking plug air-gap. A short duration high energy spark discharges the secondary capacitance and is followed by an arc discharge of longer duration. During this latter discharge period the gap voltage remains constant and an almost constant voltage is reflected into the primary circuit. Superimposed on this steady primary voltage component are rapidly decaying oscillations due to energy interchanges between the primary leakage inductance and the capacitance C_P.

When the secondary current is no longer sufficient to maintain the discharge, the secondary circuit reverts to an open-circuit. The remaining energy in the system is dissipated by lower frequency primary and secondary oscillations. The primary oscillations now take place about the voltage level of the battery negative terminal.

The battery voltage is re-applied to the primary winding when the contacts close and the cycle repeats.

The waveform shown in Fig. 2 exists across both the ignition coil primary winding and the contact-breaker, the only difference being a shift in the zero equal to the battery voltage. This difference is important, however, and the tachometer described must be connected across the coil primary winding. Thus A in Fig. 2 is the zero voltage line and the primary voltage is positive with the contacts closed and has a mean negative component when the contacts open.

Filtering and Limiting

A diode pump circuit is used to give a direct current output whose mean value is proportional to contact-breaker frequency, but the following operations must first be performed on the primary voltage waveform.

(1) *Filtering*

The high frequency, large amplitude components due to primary and secondary oscillations must be removed.

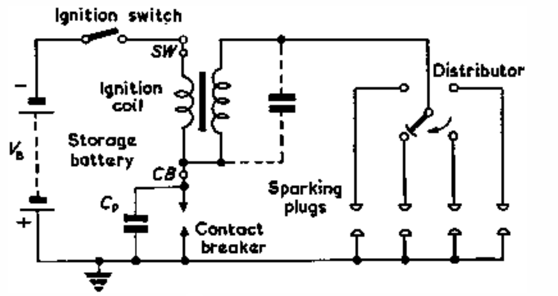

Fig. 1. Automobile ignition circuit

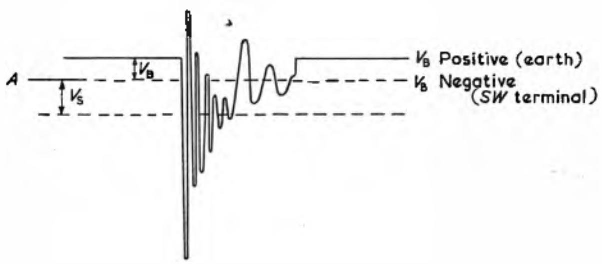

Fig. 2. Voltage waveform across the ignition coil primary winding

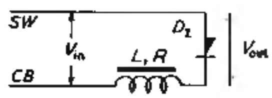

Fig. 3. Filter-limiter circuit

(2) *Limiting*

The terminal voltage of a nominal 12V lead acid battery varies from 16V (fully charged battery on charge), to say, 10V (nearly discharged battery on load). The limiting circuit must give an essentially constant amplitude voltage output over this range.

The circuit elements used must give negligible loading of the primary winding during the initial collapse of flux, otherwise the performance of the ignition system will deteriorate.

A choke and Zener diode in series, see Fig. 3, fulfills the requirements as follows:—

(*a*) The choke has a high reactance at the frequency associated with the spark discharge and so absorbs little energy.

(*b*) The Zener diode is brought into conduction in both the forward and reverse directions and its voltage-current characteristic gives excellent limiting action.

(*c*) The inductance of the choke together with the differential resistance r_d of the Zener diode form a filter that completely suppresses the higher frequencies and yet maintains the steep wave fronts occurring at the fundamental frequency. Fig. 4(b) shows the output waveform for the input of Fig. 4(a) when there is no additional circuit across the Zener diode.

Filter Action

Consider the circuit of Fig. 3 connected across the ignition coil primary winding. Since the terminal *SW* remains fixed in voltage (that of the battery negative terminal), this will be treated as the reference potential (*A* in Fig. 2). The output (V_{out}) is that developed across the Zener diode D_z.

If r_d is the instantaneous slope resistance of the Zener diode, the circuit can be treated as an $L + R$ filter with an instantaneous time-constant, t, given by

$$t = L/(R+r_d) \quad \ldots\ldots\ldots\ldots\ldots \quad (2)$$

where R is the d.c. resistance of the choke.

On the initial closure of the contacts, the Zener diode is non-conducting so that the value of r_d is high (typical value

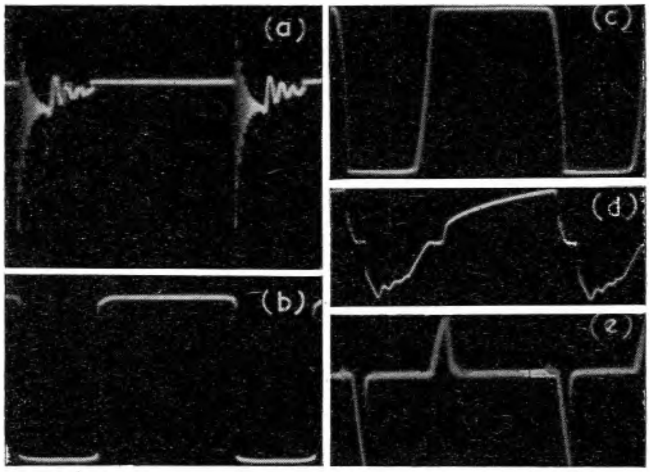

Fig. 4. Circuit waveforms (*8 cylinder engine at 2 000rev/min*)
(*a*) *Ignition coil primary voltage*
(*b*) *Filter-limiter output-voltage with pump circuit disconnected*
(*c*) *Filter-limiter output voltage for complete circuit*
(*d*) *Zener diode current for complete circuit*
(*e*) *Capacitor current*

$= 100\text{k}\,\Omega$) and t therefore small. The output voltage rises rapidly towards the input voltage $+V_B$. When V_{out} equals $+V_R$ (V_R being the reverse breakdown voltage of the Zener diode) however, the diode conducts and clamps the output voltage. r_d is now very small and t correspondingly large. The current flowing in the circuit increases exponentially towards the value I_m given by

$$I_m = (V_B - V_R)/R \quad \ldots\ldots\ldots\ldots\ldots \quad (3)$$

When the contacts open, the voltage across the ignition coil primary winding reverses. Due to the high amplitude of this voltage (up to -300V) the diode current decays rapidly, changes direction and builds up to a high negative value. As the current passes through zero, the Zener diode cuts off and there is a rapid transition in output voltage from the value $+V_R$ to $-V_F$ (V_F being the forward voltage drop of the Zener diode). High amplitude, high frequency input voltage oscillations follow superimposed on a mean negative component. This negative component and the long time-constant of the circuit maintain a unidirectional current flow. Fig. 4(d) shows the gradually decreasing diode current with the high frequency oscillations superimposed. The output voltage (Fig. 4(b)) remains at $-V_F$.

At all but the very lowest engine speeds the contacts close again before the forward current has decayed to zero and voltage $+V_R$ is then re-applied. This causes a more rapid current decay and, after a short time-lag (compare Figs. 4(a) and 4(b)), the diode is cut off and the output voltage changes from $-V_F$ to $+V_R$ when a reverse diode current flows once more.

The output voltage is a square wave of amplitude ($V_R + V_F$) and frequency equal to that of the contact-breaker.

Diode Pump

The Zener diode waveform feeds a diode pump circuit as shown in the complete circuit* diagram of Fig. 5.

During the transition from Zener diode voltage $-V_F$ to $+V_R$ diode D_1 conducts and the potential across the capacitor changes from $-(V_F - v)$ to $+(V_R - v)$ where v is the forward voltage drop of diodes D_1 or D_2. On the next transition diode D_2 conducts and the capacitor potential returns to $-(V_F - v)$. The charge q transferred during each transition is given by

$$q = \pm C.(V_R + V_F - 2v) \quad \dots\dots\dots\dots (4)$$

Distinct periods of time now exist when the Zener diode is non-conducting and the capacitor takes the full choke current. When the capacitor has charged to the appropriate voltage level, the choke current switches into the Zener diode. Photographs of the waveforms are given in Figs. 4(a) to 4(e).

The d.c. milliammeter in series with diode D_1 passes a unidirectional charge pulse once for each cycle of the input waveform. The average meter current i is therefore given by

$$i = q.f = C.(V_R + V_F - 2v).\ f.\ 10^{-3} \quad \dots\dots\dots\dots (5)$$

where i is measured in milliamperes
C ,, ,, microfarads
f ,, ,, cycles per second
V_R, V_F, v are measured in volts.

Since the meter cannot respond to the ripple frequency a steady reading proportional to input frequency (i.e., engine speed) is obtained.

Contact Bounce

The distributor contact-breaker is required to operate at high mechanical frequencies (e.g., 250c/s for a 6-cylinder engine at 5000rev/min) and a small amount of contact bounce usually occurs at the higher engine speeds. This leads to additional voltage fluctuations across the primary winding of the ignition coil. Incorrect instrument readings will be obtained if these fluctuations should reach the frequency measuring circuit.

Referring to the Zener diode current waveform (Fig. 4(d)) the filter choke maintains the flow of current for a short period of time after the contacts have closed. This delay is exploited by ensuring that the circuit inductance is sufficient to hold the Zener diode on its forward characteristic during the expected bounce time.

Although not yet fully investigated, improved rejection of undesirable voltage fluctuations caused by severe bounce is expected by using a choke which saturates at some low current value i_s. Thus, the inductance may be high for currents within the limits $+i_s$ to $-i_s$ but low for currents outside this range. The sequence of events is then as follows. At the instant of closure of the distributor contacts, the choke carries a relatively high current and its inductance is low. A rapid initial fall of current therefore results until the value i_s is reached. This takes the choke into the region of high inductance and a more gradual rate of change occurs while the current decays to zero, reverses, and increases to the value i_s in the opposite direction. The transition to low inductance now occurs, the current increasing rapidly to the value limited by circuit resistances.

Assuming that i_s is small compared with the current required to charge and discharge the capacitor, a 'dead time' is established upon contact closure during which the circuit currents are negligible. At the end of this period, the capacitor charges rapidly to the voltage limited by conduction of the Zener diode on its reverse characteristic.

The above description is somewhat idealized since the inductance of the choke would vary more gradually. However, improved rejection of multiple current reversals is obtained together with faster charging of the capacitor.

Additional advantages of the saturable choke are :

(1) Lighter loading of the ignition system at 'break'. When the distributor contacts open, the voltage across the ignition coil primary winding reverses and the choke is again switched into the region of high inductance while the current is reversing. As a result only a small current flows during the initial stages of spark formation.

(2) The saturable choke is smaller and simpler than a choke designed to give constant inductance over the full range of operating current.

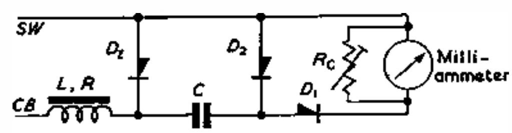

Fig. 5. Complete circuit diagram of the tachometer

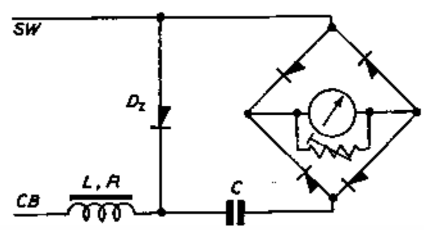

Fig. 6. Diode pump using a full-wave bridge

Modified Diode Pump Circuit

In certain circumstances, a bridge arrangement of diodes (see Fig. 6) may be used to advantage in the frequency measuring circuit. Both capacitor current pulses will then be gated through the milliammeter allowing either:—

(1) An increase in output current.

(2) A reduction in the capacitor value required for a given output current.

The modification is also useful when very low engine speeds (below the usual tick-over value) are of interest since the repetition frequency of the meter current pulses is increased giving less pointer vibration.

Design Details

The choke inductance and resistance are determined by the following considerations:

(1) The resistance R and time-constant L/R must allow capacitor C to charge and discharge completely at the highest speed to be measured.

(2) The time-constant L/R must be long enough to suppress oscillations due to inductive and capacitive components in the ignition circuit.

(3) R must be high enough to prevent burn-out of the Zener diode with V_B permanently applied—ignition on, engine stationary and the contacts closed.

(4) For best linearity the Zener diode reverse current should be allowed to approach closely the maximum

* *Patent applied for*

I_m (see equation (3)) at all speeds. This minimizes variations of V_R with speed.

The choke windings must be insulated to withstand voltage peaks up to 300V.

In equation (5), voltages V_F and v have negative temperature coefficients. The temperature coefficient of the reverse break-down voltage V_R can be negative, zero or positive depending on the value of V_R. A suitable choice will reduce the total temperature coefficient of ($V_R + V_F - 2v$) almost to zero over the expected range. Fortunately, the value required coincides with the value giving minimum slope resistance so that variations of V_R with battery voltage are simultaneously at a minimum.

The following values are suitable for a 12V system

Choke	$L = 1{\cdot}0$H,		$R = 150\,\Omega$
Zener diode	reverse breakdown	$= 6{\cdot}8$V.	
	Dissipation	$= 300$mW.	
	(G.E.C. type SX68 has been found suitable.)		
Milliammeter	f.s.d. = 1mA,		Resistance = $100\,\Omega$
Resistance R_C	1kΩ variable		
Capacitor	$1{\cdot}0\mu$F	—	4-cylinder
	$0{\cdot}7\mu$F	—	6 cylinder
	$0{\cdot}5\mu$F	—	8 cylinder
Diodes D_1 and D_2	Low power germanium diode.		

Performance

The prototype gave the following performance

Linearity

Within $\pm$ 0·5 per cent of full scale deflexion.

Accuracy

At any given engine speed, the scale reading changes less than $\pm 0{\cdot}8$ per cent for battery voltage variations within the range 10 to 16V and ambient temperature variations from 10 to 40°C.

Conclusion

The tachometer is sufficiently accurate for laboratory use while its simplicity and long-term reliability make it suitable for permanent installation in automobiles.

Acknowledgments

The author wishes to thank the Directors of Joseph Lucas Electrical Ltd for permission to publish this article and members of the Electrical Research and Ignition Laboratories, particularly Mr. M. H. Roberts, for their suggestions and assistance.

A Simple Transistor Tester

By G. G. Yates*, M.A.

A simple and inexpensive instrument is described for the measurement of the more useful parameters (I_{co}', β, h_{11}', f_a) of low power a.f. and r.f. transistors. I_{co}' and β are determined by a d.c. method, no additional apparatus being required. For the determination of h_{11}' and f_a a signal generator with a frequency range from below 1kc/s to about 200kc/s and a valve-voltmeter or oscilloscope are required in addition. As even a modestly equipped electronics laboratory may be expected to possess such instruments this is not regarded as a disadvantage.

(Voir page 637 pour le résumé en français: Zusammenfassung in deutscher Sprache Seite 644)

THE increasing use of transistors makes a transistor tester a useful addition to laboratory test equipment. A complete set of characteristic curves for a transistor may be plotted out using quite simple and inexpensive apparatus, but the procedure would be extremely tedious, while an instrument designed to trace out the characteristics say on a cathode-ray tube would be relatively complicated and expensive.

Fortunately, the more useful parameters (I_{co}', β, h_{11}', f_a) can be measured quickly using a simple instrument such as the one described here. The complete circuit is given in Fig. 1. The transistor under test is operated in the common (grounded) emitter connexion when connected to the terminals *E* (emitter), *B* (base) and *C* (collector) of the test instrument. Power for the transistor is supplied by two 4·5V dry batteries incorporated in the instrument, the transistor itself being operated at a collector voltage of $-4{\cdot}5$V. The batteries also supply power to the two stage transistor amplifier which is used when making a.c. measurements.

Measurement of I_{co}' and β

These quantities may be measured without any additional equipment, nor is the internal amplifier used. The switch S_2 (Fig. 1) is closed and the three way selector switch S_1 set in position 2. The transistor base is then open-circuited, while the collector is at $-$ 4·5V, the emitter, of course, being grounded. The meter then reads I_{co}', the 220Ω meter shunt which is in circuit in this position of the switch S_1 gives a f.s.d. corresponding to 200μA collector current. This is generally adequate for most l.f., i.f. and r.f. transistors (except the special high power output types) at ordinary temperatures.

To measure β the selector switch S_1 is moved to position 3. This brings a 22Ω meter shunt into circuit so that the f.s.d. now corresponds to a total current of 1mA. At the same time bias current is fed into the base of the transistor. The 1MΩ 'set zero' variable resistor is now adjusted until the meter reads zero. The collector voltage is then about $-4{\cdot}5$V and the collector current approximately 1mA, this being the current flowing through the 4·2kΩ resistor in the collector circuit. The press switch S_4 is now closed, increasing the base bias current by 10μA, the corresponding increase in collector current is recorded on the meter and gives a direct reading of β. There will be a slight increase in current through the 4·2kΩ resistor but this is quite negligible compared with the current flowing from the 4·5V battery through the meter and its shunt.

Measurement of Input Impedance and Cut-Off Frequency

These are a.c. measurements and require, in addition to the test unit, a signal generator having a range from below

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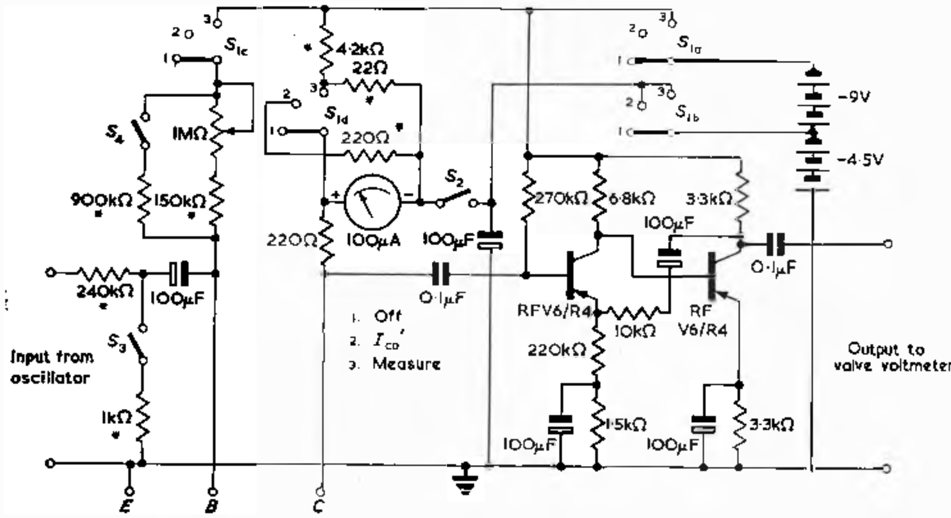


Fig. 1. Transistor test unit

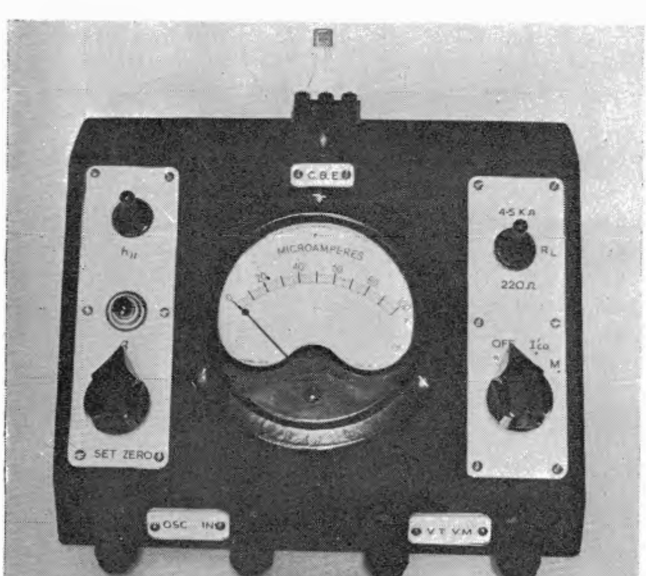
* *High stability resistors (close tolerance)*

1kc/s to 200kc/s, and also a valve-voltmeter with a 1·5V a.c. range. An oscilloscope could be substituted for the valve-voltmeter with suitable changes in procedure. The oscillator and valve-voltmeter are connected to the terminals as indicated in Fig. 1.

The a.c. signal from the oscillator is fed to the transistor through a 240kΩ resistor, so that it may be regarded as a constant current source so far as the transistor is concerned. With the switch S_2 still closed the a.c. collector load is 220Ω. Thus the transistor is operating very nearly with an open-circuit input and short-circuit output. The small a.c. voltage at the collector is amplified by the wide-band, gain stabilized, transistor amplifier. The output from the amplifier is measured by the valve-voltmeter. This output should not exceed about 3V peak-to-peak if distortion is to be avoided.

The following procedure is adopted in the measurement of input impedance. First the switch S_3 is closed, this shunts the input with 1kΩ resistor. Next the oscillator frequency is set to 1kc/s and the oscillator output voltage is adjusted until the valve-voltmeter reads 0·5V. Finally S_3 is opened and the reading on the valve-voltmeter is noted.

A transistor tester



If this latter reading is v then the input impedance h_{11}' is calculated from the formula

$h_{11}' = (2v - 1)$kΩ [see appendix]

If the foregoing measurement is carried out with the switch S_2 open, then the input impedance corresponding to a collector load of 4·4kΩ will be found.

The cut-off frequency of the transistor may be determined as follows. The switch S_2 is closed and S_3 opened. The oscillator is set to 1kc/s and output voltage adjusted until the valve-voltmeter reads 1·0V. The oscillator frequency is now increased, the oscillator voltage being held constant, until the reading on the valve-voltmeter has fallen to 0·7V. The frequency at which this occurs is the β cut-off frequency f_β. The α cut-off frequency is deduced from the relation

$$f_\alpha = \beta f_\beta$$

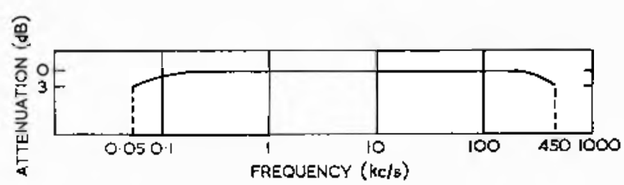


Fig. 2. Response curve of amplifier

The Amplifier

The amplifier, which has a gain of 30dB employs two r.f. transistors (type V6/R4) with cut-off frequencies in the region of 6Mc/s. The use of feedback from the collector of the second transistor to the emitter of the first ensures stable gain, wide band width and a high input impedance. The response is flat from 200c/s to 200kc/s, and this is adequate for the measurement of f_β of audio transistors and the commonly available r.f. types, the latter having f_β based on the amplifier response, in the neighbourhood of 150kc/s. If suitable corrections are applied measurements could be extended to 500kc/s if required. The amplifier response curve is shown in Fig. 2.

Appendix

The Formula for Input Impedance

Let the current (constant) supplied by the generator be i. When S_3 is closed suppose the current taken by the transistor is i_1, then the current flowing in the 1kΩ shunt will be $i - i_1$.

If the input impedance of the transistor is r_{in}(kΩ), then

$$\frac{i - i_1}{i_1} = r_{in} \qquad \ldots\ldots\ldots (1)$$

When S_3 is closed the valve-voltmeter reading is 0·5. When S_3 is open the reading is v, and in this case the current flowing into the transistor is i

$$\therefore\ v/0{\cdot}5 = i/i_1 \qquad \ldots\ldots\ldots (2)$$

From equations (1) and (2)

$$r_{in} = 2v - 1.$$

A Transistor Scaling Circuit with a Short Resolving Time

By B. Collinge*, B.Sc., Ph.D.

Fast scale-of-two circuits are required for a number of applications. In this article the advantages are discussed of coupling one scale-of-two circuit to another by means of a gate circuit.

A circuit diagram and a description are given of a complete unit consisting of a gate circuit and a bistable circuit

(Voir page 637 pour le résumé en français; Zusammenfassung in deutscher Sprache Seite 644)

FAST scale-of-two circuits are required for many applications, for example in pulse counting circuits, frequency meters, and in data handling devices. Many scale-of-two circuits are described in the literature [1,2] and they may be connected together to form registers or counting circuits.

In some circumstances there is some advantage in coupling one scale-of-two to another by means of a gate circuit. The principle of this arrangement is shown in Fig. 1. The transistors form an emitter coupled, bistable circuit and changes of state are initiated by applying pulses of alternate sign to the base of X_3. These pulses are provided by the gate circuit and their polarity is controlled by the state of the bistable circuit to which the gate is directly connected.

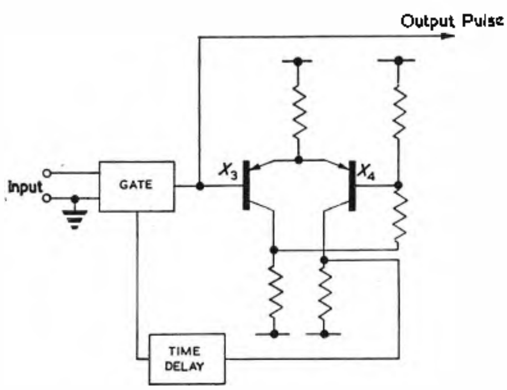


Fig. 1. A gate driven scale-of-two circuit

The advantages of using a gate to drive the bistable circuit are particularly evident when high speed scaling is required. This is because the functions of scaling and memory are located in different parts of the circuit. The time delay, which is shown in Fig. 1 in the connexion between the bistable circuit and the gate, is made larger than the width of the input pulse. The state of the gate is changed at a time when the triggering pulse has been transmitted by the gate and when the bistable circuit has reached its new state. Hence the conditions for reliable triggering are particularly favourable even when the widths of the input pulses are comparable with the time for the bistable circuit to change from one state to another.

The use of a gate to drive a bistable circuit also has advantages over conventional, directly coupled scales-of-two, in certain applications. For example, time delays in a counter or register are reduced. Drive pulses to other circuits are provided by the gate and a particular bistable circuit may be set to a chosen state without interfering with an adjacent bistable circuit.

The Scale-of-Two Circuit

Gate Circuit

A complete scale-of-two circuit is shown in Fig. 2 in which the gate consists of the transformer and the transistors X_1 and X_2. Negative pulses, applied to the base of X_1, turn on a defined current of about 10mA which flows into the primary of the transformer. The base of the transistor, X_2, is connected to the bistable circuit so that the

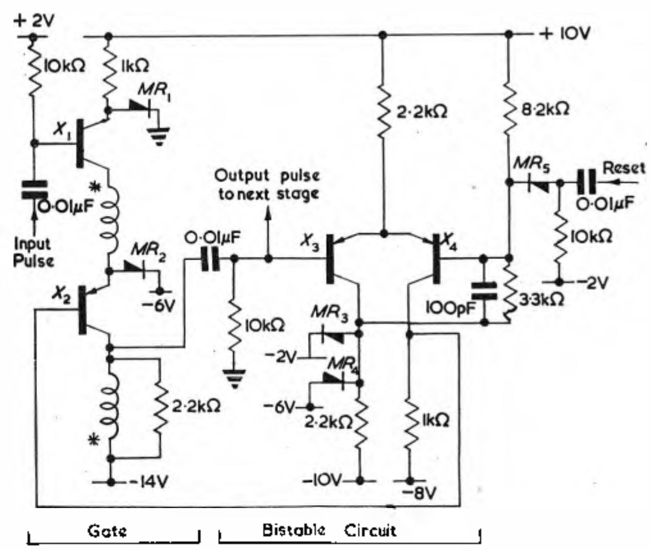


Fig. 2. Scale-of-two circuit

primary current flows into the secondary winding if X_4 is conducting, and into the diode MR_2 if X_2 is not conducting. The phases of the transformer windings and the number of turns are arranged so that a pulse of the correct sign appears at the base of X_3.

Bistable Circuit

The bistable circuit consists of two emitter coupled transistors with a cross connexion between the collector of X_3 and the base of X_4. The collector voltage excursion, is clamped by the diodes MR_3 and MR_4, and it is complete in about 50mμsec. The bistable circuit may be set to zero by a positive signal applied to the base of X_4 through an isolating diode, MR_5.

The time delay shown in Fig. 2 is not necessary if the input pulses have a controlled width of 0·1μsec. Charge storage effects and the transformer provide sufficient delay. Wider input pulses may require the addition of a time delay which may take the form of a small capacitor connected from the collector of X_4 to earth, or a delay network in the position shown. The time delay must always be larger than the input pulse width.

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PERFORMANCE

Sixteen scale-of-two circuits have been constructed with standard components using the circuit diagram shown in Fig. 2. Each unit functioned satisfactorily with pairs of input pulses which were 0·1μsec wide and whose leading edges were separated by 0·2μsec. The amplitude was 4V.

By selecting the diodes MR_3 and MR_4 for minimum collector rise and fall time, it was found that the resolving time was greatly improved. All the circuits would then count pairs of pulses which had a width of 0·1μsec and which were just visibly separated when viewed on a Tektronix 545 oscilloscope.

Acknowledgments

The author wishes to thank Mr. D. Hyde and Mr. D. Wallace who constructed and tested these units.

APPENDIX

Transformer details

One winding of 10 turns and one winding of 20 turns of 36 s.w.g. wound on Mullard ferroxcube cores type FX 1238 and E. K. Cole bobbins type DP 10857.

Transistors

Unselected OC44 transistors, as supplied by Mullard Ltd, were used in these circuits.

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A Sign and Magnitude Display for Use with Reversible Counters

By D. L. A. Barber*, B.Sc.

A circuit is described which may be used with any reversible counter to provide a display of the sign and magnitude of numbers.

(Voir page 637 pour le résumé en français; Zusammenfassung in deutscher Sprache Seite 644)

A COUNTER which will both add and subtract often indicates a negative number as a true complement, for example, a two stage decimal counter indicates -1 as 99.

The disadvantages of this type of display are that an operator may be confused and if such a counter is controlling a servo an incorrect error signal is obtained when passing from 0 to -1, during the time of propagation of the 'carry' signal through the counter. In the example above, as the number indicated changes from 00 through 09 to 99, the error signal would change from zero to $+9$ and finally to -1 (represented by 99).

The circuit to be described allows any reversible counter to provide a display of the sign and magnitude of the count stored, by reversing the connexions of the input lines to the counter, when appropriate, during the counting process. The sign is then indicated by the state of these input connexions.

The requirements of the circuit are:—

(1) If the counter indicates zero an add pulse must indicate a positive sign and a subtract pulse a negative sign.

(2) If a positive sign is indicated, each add pulse must be applied to the add input, and each subtract pulse to the subtract input, of the counter.

(3) If a negative sign is indicated each add pulse is applied to the subtract input, and each subtract pulse to the add input, of the counter.

A logical diagram which satisfies the requirements is shown in Fig. 1.

When a stage of the counter indicates zero a signal is applied to an 'AND' gate A. If all stages are at zero, a signal from gate A is arranged to open a pair of gates B and C, which allow the next add or subtract input pulse to set appropriately the bistable circuit. The output from the bistable circuit, when a positive sign is indicated, opens gates F and G so that the add and subtract lines are connected respectively to the add and subtract inputs of the counter. When a negative sign is displayed gates D and E

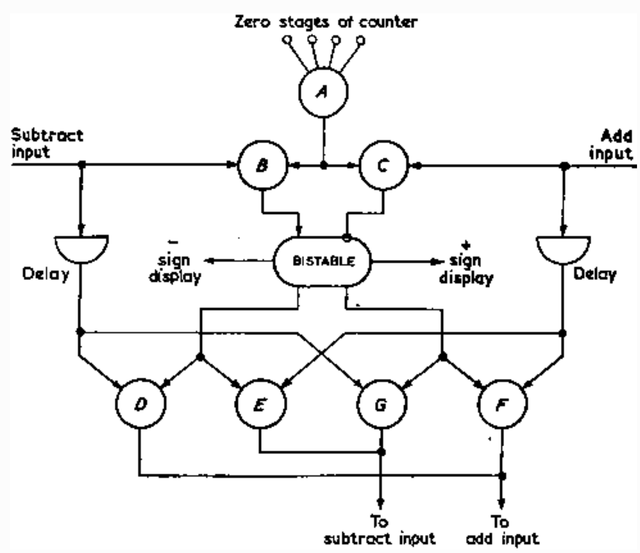

Fig. 1. Logical diagram

are opened and the connexions between the input lines and the counter are reversed. A delay is included in each input line to ensure that the bistable circuit is set before a count is registered.

Fig. 2 shows a practical circuit which uses transistors. The circuit was used in conjunction with a reversible Dekatron counter in which a signal is available which changes from 0 to -12V when zero is indicated[1]. It was con-

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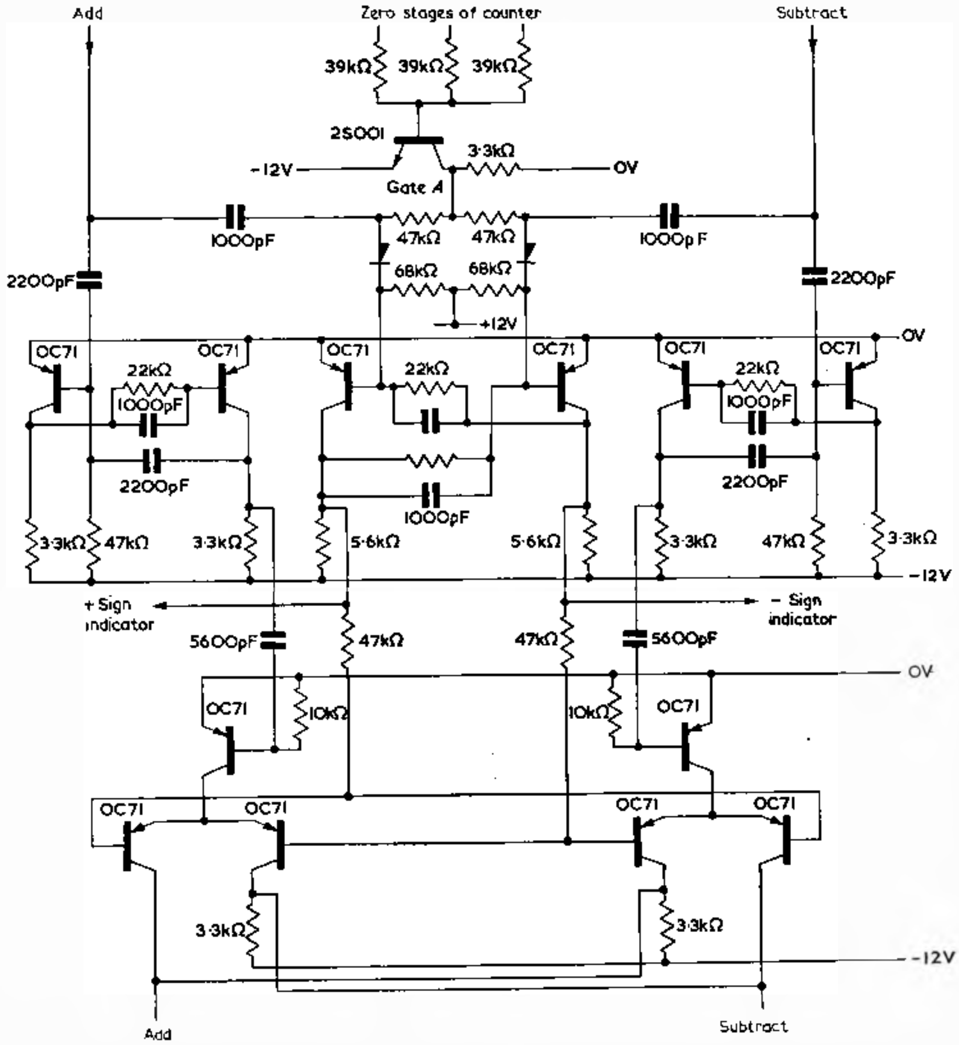

Fig. 2. Practical circuit using transistors

venient therefore to use an npn transistor as gate A. With all inputs at $-12V$ transistor A is cut-off and its collector is near zero potential. The diode gates B and C are then biased so that a positive pulse on an input line will open the appropriate gate and set the bistable circuit. The output gates are arranged to route a pulse formed from one of the monostable circuits to either the add or subtract output depending upon the state of the bistable circuit

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Machine Tool Control in Yugoslavia

During recent years the necessity for increased production has become a problem of growing significance for Yugoslav industry. In this connexion, early in 1956 a firm known as P.A.E. was formed with the object of gradually introducing modern methods of automatic production into Yugoslav factories. The first serious task that was undertaken was the production of a numerically controlled machine tool, a project sponsored by the S.I.K. (Central Chamber of Industry). It was decided to build a prototype machine for demonstration purposes and training the personnel involved and also as a means of developing the specialized experience needed in designing this type of equipment.

For the early work a radial drilling machine of Yugoslav origin was developed and later this was rebuilt as a co-ordinate drilling machine; it being considered that this was the easiest type of machine for numerical control as it only requires accurate positioning on two axes, the depth of drilling being controlled by a system of end relays.

The positioning was performed by separate measuring screws. A slotted disk was attached to each of these and 'measuring' pulses created by means of photo-transistors. A reversible register using Dekatrons was used to compare the 'measuring' pulses with 'command' pulses which were fed into the register from perforated tape. When the table of the machine approaches the correct position the error voltage at the output of the register gradually falls in magnitude. If the correct ordinate is over-run the error signal changes sign and the table may thus take up its correct position by means of a few highly clamped oscillations. When the table comes to rest in the correct position a relay actuates a switch to start the main driving motor of the machine. When the drilling operation has been completed the motor is automatically reversed and switched off and the next positioning commences; the new numerical data already having been fed into the register.

The accuracy of positioning is limited by mechanical errors but, nevertheless, an accuracy of $\pm 20\mu$ has so far been achieved. From the start of this development it was realized that a method involving measuring screws could not provide the highest accuracy but at that time it was not possible to obtain either optical diffraction gratings or linear Inductosyns in Yugoslavia.

Approximate Waveform Solutions for Diodes in Pulse Circuits

By D. C. Dillistone*

Taking an approximate representation of a diode, waveform solutions are obtained for a group of simple circuits to which a rectangular, or (in one case) a triangular waveform is applied. Certain applications of the results are discussed in general terms.

(Voir page 637 pour le résumé en français: Zusammenfassung in deutscher Sprache Seite 644)

ALTHOUGH the diode is by far the most commonly used non-linear circuit element many designers appear to regard its non-linearity as an insurmountable obstacle in the path of analytical investigation. In this article a number of simple circuits containing diodes are collected together and analysed in the hope that the results themselves as well as the method of obtaining them may be of interest.

The diode symbol in the circuits which follow is intended to represent what might be called an 'ideal' diode—that is to say a circuit element which is considered a short- or open-circuit according as the voltage across it is positive or negative. In most cases this approximation is not as sweeping as it may appear since the 'forward' and 'reverse' resistances may often be included as elements in the external circuit. Thus, in effect, the curvature of the diode characteristic in the vicinity of the origin is neglected and it is assumed that the change from conductive to non-conductive state occurs at the origin. Particularly in pulse circuits, where waveforms are large, the few hundred millivolts by which the 'break-point' departs from the origin are generally unimportant.

The approach which has been adopted is to consider separately the two circuits which are formed by open- and short-circuiting the diode. In each circuit the initial conditions are unknown as, in some cases, are the times of transition from one circuit state to another. By assigning arbitrary values to these unknowns a 'waveform' solution for each state is derived—the unknowns may then be eliminated by equating the circuit state at the end of the one interval to that at the beginning of the other.

In what follows a familiarity with the methods of the Laplace transform is assumed. As here used it is defined by:

$$\mathcal{L}f(t) = \int_0^\infty e^{-pt} f(t)\, dt$$

Also in order to avoid the comparative clumsiness of direct appeal to differential equations a change of origin or an impulse function generator is used to insert the appropriate initial conditions.

Series Diode Response to Square Wave

Referring to Fig. 1: (a) shows the circuit under consideration and the applied waveform is indicated in (b) by *a.b.d.e.g.* It is clear that the diode will conduct during the period $0 < t < h$ and be non-conductive during $h < t < 2h$. Thus the waveform of the voltage E_2 will have the general appearance of *m.c.f.*; that of the voltage E_1 being *m.c.d.e.f.* since E_1 equals the applied waveform when the diode is an open-circuit. Thus it remains to find the exact forms of *m.c.* and *c.f.* Arbitrary values A_1 and A_2 are assigned to the starting values of these waveforms as indicated and one can proceed as follows: During $0 < t < h$ Thevenin's theorem gives the equivalent circuit of Fig. 1(c). By a change of origin to the point $(0, A_1)$ one may consider $f(t)$, the time function applied to the circuit, as $(A - A_1) \,.\, H(t)$

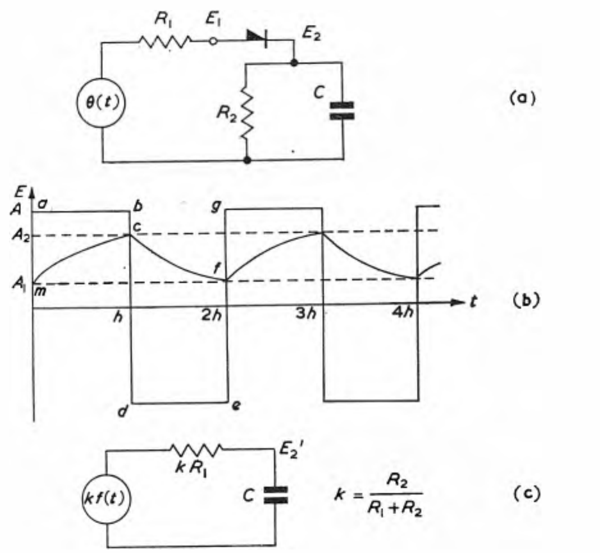

Fig. 1. (a) A typical series diode circuit
(b) An applied square wave and resultant circuit response
(c) A circuit equivalent to (a) when the diode is short-circuited

—i.e., a step function of height $A - A_1$. Then:

$$\mathcal{L}E_2' = k\frac{A-A_1}{p}\cdot\frac{1/pC}{kR_1+1/pC} = k(A-A_1)\frac{1/kCR_1}{p(p+1/kCR_1)}$$

$$\therefore E_2 = A_1 + k(A - A_1)(1 - \exp(-t/kCR_1)) \;\ldots\ldots\; (1)$$

During $h < t < 2h$ it is clear that E_2 is a simple exponential decay towards earth level starting from A_2.

i.e.

$$E_2 = A_2 \exp\left(-\frac{t-h}{CR_2}\right) \qquad \ldots\ldots\ldots\ldots\; (2)$$

Since $E_2 = A_2$ at $t = h$ and $E_2 = A_1$ at $t = 2h$ one obtains from equations (1) and (2) the pair:

$$\left.\begin{aligned} A_2 &= k(A - A_1)(1 - \exp(-h/kCR_1)) + A_1 \\ A_1 &= A_2 \exp(-h/CR_2) \end{aligned}\right\}$$

These yield:

$$A_1 = \frac{A}{1 + \dfrac{\exp(h/CR_2) - 1}{k(1-\exp(-h/kCR_1))}}$$

$$A_2 = \frac{A}{\exp(-h/CR_2) + \dfrac{1-\exp(-h/CR_2)}{k(1-\exp(-h/kCR_1))}}$$

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

Thus the equations (1) and (2) constitute the complete solution when the values for A_1 and A_2 are inserted. It is possibly worth noting that the steady-state response of the circuit has been obtained: the transient response may be calculated in a similar way, treating successive cycles individually.

Shunt Diode Response to Rectangular Wave

Referring to Fig. 2: the waveform of (b) is applied to the circuit shown at (a). The general form of the response is indicated in (c) and the arbitrary values A_1 and A_2 are assigned to the starting points of the waveforms. Clearly

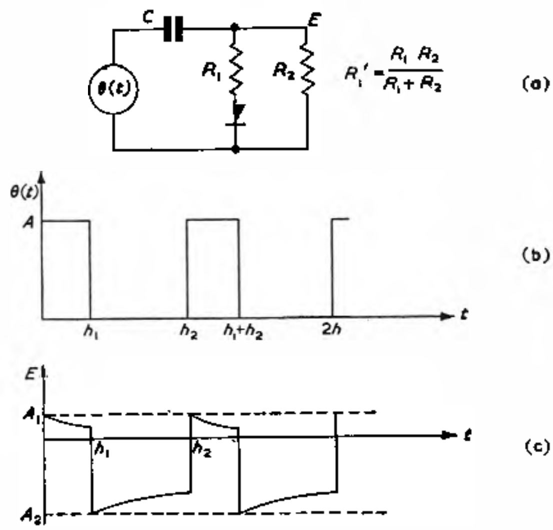

Fig. 2. (a) A typical shunt diode circuit
(b) An applied rectangular waveform
(c) The circuit response

in $0 < t < h_1$ a step-function of height A_1 is being applied to C and R' in series. Thus:

$$\mathcal{L} E = (A_1/p) \cdot \frac{R'}{R' + 1/pC} = \frac{A_1}{p + 1/CR'}$$

$$\therefore E = A_1 \exp(-t/CR') \quad \ldots\ldots\ldots\ldots\ldots \text{(3)}$$

Similarly for $h_1 < t < h_2$:

$$E = -A_2 \exp\left(-\frac{t - h_1}{CR_2}\right) \quad \ldots\ldots\ldots\ldots \text{(4)}$$

From equation (3) with $t = h_1$ and equation (4) with $t = h_2$:

$$\left.\begin{aligned} A_1 \exp(-h_1/CR') + A_2 &= A \\ A_1 + A_2 \exp -\left(\frac{h_2 - h_1}{CR_2}\right) &= A \end{aligned}\right\}$$

$$A_2 = A \frac{1 - \exp(-h_1/CR')}{1 - \exp -\left(\frac{h_2 - h_1}{CR_2} + h_1/CR'\right)}$$

$\therefore$

$$A_1 = A \; \frac{1 - \exp(-h_2 - h_1)/CR_2)}{1 - \exp -\left(\frac{h_2 - h_1}{CR_2} + h_1/CR'\right)}$$

Series Diode Response to 'Sawtooth' Wave

Consider now the slightly more complicated case in which the 'sawtooth' waveform (*a.b.c.d.*) of Fig. 3 is applied to the circuit of Fig. 1(a). An uncertainty now arises as to the point at which the diode begins to conduct and an arbitrary time value, t_1, must be assigned to this point (f in Fig. 3). During $0 < t < t_1$ it is clear that $E_2 = A_1 \exp(-t/CR_2)$ as in previous cases. If one now changes the origin to the point f, i.e., $\{t_1,\ A/h\,(t_1 - h)\}$, one can consider the equivalent circuit of Fig. 1(c) with $f(t) = A/h\,.\,t\,.\,H(t)$.

Then:

$$\mathcal{L} E_2' = \frac{1/kCR_1}{p + 1/1/kCR_1} \,.\, k \,.\, \mathcal{L} f(t) = Ak/h \,.\, \frac{1/kCR_1}{p^2(p + 1/kCR_1)}$$

$$\therefore E_2' = A/hCR_1 \int_0^t \int_0^t \exp(-t/kCR_1)\,(dt)^2 =$$

$$Ak/h \int_0^t (1 - \exp(-t/kCR_1)\ dt.$$

$$\therefore E_2' = Ak/h\,[t - kCR_1(1 - \exp(-t/kCR_1))] \quad \ldots\ldots \text{(5)}$$

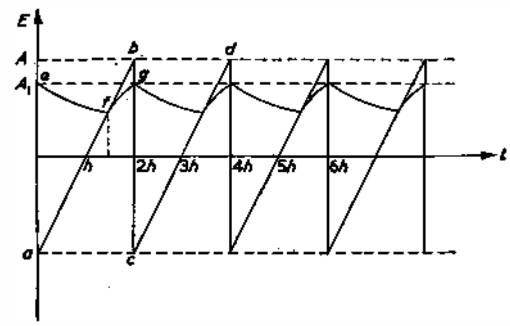

Fig. 3. A triangular waveform (*abcd*) and the response of a series diode circuit (*efg*)

Now the height of the point g relative to f is clearly given by putting $t = 2h - t_1$ in equation (5). Similarly the ordinate of f is given by $A_1 \exp(-t_1/CR_2)$. Thence is obtained the equation pair:

$$\left.\begin{aligned} A[(t_1/h) - 1] &= A_1 \exp(-t_1/CR_2) \\ A_1 &= A[(t_1/h) - 1] + Ak/h\left[2h - t_1 - kCR_1\left(1 - \exp -\frac{2h - t_1}{kCR_1}\right)\right] \end{aligned}\right\}$$

$$\ldots\ldots\ldots\ldots \text{(6)}$$

These equations together with equation (5) and the solution for the curve *e.f.* constitute a complete implicit solution. It will be appreciated, however, that the equations (6) do not permit an explicit solution except in numerical cases where approximate methods may be used. Thus one is reaching the limit beyond which solutions are of little practical use. The author has, in fact, obtained implicit solutions for the case of a triangular wave applied to a shunt diode and for a sine wave applied to a series diode: in both cases when the complexity of the equations corresponding to (6) above is weighed against the ease with which such circuits may be investigated empirically there seems little doubt that the analytical results are of small value.

There exists a class of circuits incorporating diodes which may be considered to be excited by single pulses. These yield readily to analytical treatment as the following two examples show.

Response of Ringing Circuit with Diode Damping

As illustrated in Fig. 4(a) a negative current step function of height I_0 is applied to the parallel combination of an inductor, a capacitor and a diode. In effect one is here considering what might be called the full equivalent circuit of the diode: the forward and reverse resistances of the diode are accounted for by R_1 and R_2 respectively and the diode self-capacitance may be thought to be included in C. Unfortunately it is not possible to take account in this analysis of the variations in diode capacitance which occur with variation in reverse voltage (this applies, of course, only to a junction or point contact diode).

Initially the diode does not conduct and thus:

E_1 is the exponential rise; $A(1 - \exp(-t/C_1R_1))$. At the end of the pulse $t = h$ and the voltage across C_1 is $A(1 - \exp(-h/C_1R_1)) = A_1$ (say).

When the diode does not conduct one may consider the equivalent circuit of Fig. 5(c) in which the current generator is placed across C_1 to insert the voltage A_1 across it instan-

$$\mathcal{L}E = -(I_0/p \cdot \frac{1}{1/R_2 + pC + 1/pL} = -I_0\sqrt{(L/C)}\frac{\omega_0}{p^2 + 2\alpha p + \omega_0^2} \qquad \left[\begin{array}{l}\alpha = 1/CR_2 \\ \omega_0^2 - 1/LC\end{array}\right.$$

$$\therefore e^{\alpha t}E = -\mathcal{L}^{-1}I_0\sqrt{(L/C)}\frac{\omega_0}{p^2 + \omega_0^2 - \alpha^2} = -\frac{I_0}{\sqrt{(1/L^2 - 1/R^2)}} \cdot \sin\sqrt{(\omega_0^2 - \alpha^2)}t$$

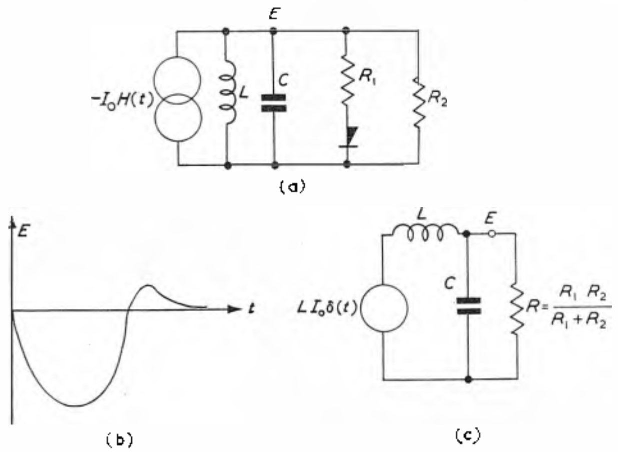

Fig. 4. (a) A ringing circuit with diode damping
(b) The response of the circuit to a current step function
(c) The equivalent circuit when the diode is short-circuited

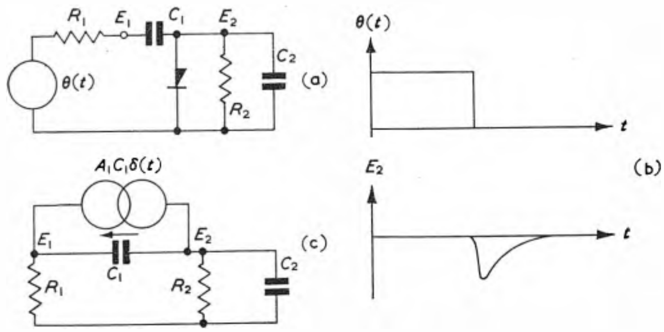

Fig. 5. (a) A shunt diode differentiating circuit
(b) The applied square pulse and the single resulting sharp pulse
(c) The equivalent circuit when the diode does not conduct
The current generator inserts the charge that has been built up across C_1 during the period of diode conduction

Since only one half cycle is under consideration it is justifiable in almost all cases to neglect R_2 so that

$$E = -I_0\sqrt{(L/C)}\sin\omega_0 t \quad \ldots\ldots\ldots\ldots \quad (7)$$

Also when $t = \pi/\omega_0$ the current in the inductor

$$= 1/L\int_0^{\pi/\omega_0} E dt = -2(I_0/L)\sqrt{(L/C)} \cdot \sqrt{(LC)} = -2I_0$$

However a 'quiescent' current $-I_0$ exists in the circuit due to the applied step function so that the effective current origin is at $-I_0$. With respect to this origin a current $-I_0$ exists in the inductor at $t = \pi/\omega_0$.

Thus when the diode conducts the equivalent circuit is that shown in Fig. 4(c) in which the impulse function generator is present to insert the desired initial current in the inductor. Proceeding from that circuit:

$$\mathcal{L}E = LI_0\frac{1(1/R + pC)}{pL + 1/(1/R + pC)} = I_0/C\frac{1}{p^2 + 1/CR\ p + 1/LC}$$

$$\therefore E = I_0/C\ e^{-\alpha t}\sin h\ \beta t$$

$$\text{where } \alpha = 1/2CR\ \beta^2 = \alpha^2 - 1/LC$$

or

$$E = I_0/C\ t\ e^{-\alpha t} \text{ when } \beta = 0.$$

In both cases the diode does not conduct again since the voltage does not fall below zero. In the worst case when $\beta = 0$ the height of the positive pulse is given by:

$$dE/dt = 0 = e^{-\alpha t} - \alpha t\,e^{-\alpha t} \text{ giving } t = 1/\alpha = 2CR$$

$$\therefore E_{max} = 2RI_0/e$$

Shunt Diode Response to Rectangular Pulse

Referring to Fig. 5: (a) shows the circuit to be investigated and in (b) are indicated the forms of the applied pulse and the expected response at E_1 and E_2.

Clearly for $0 < t < h$ the diode conducts, $E_2 = 0$ and

taneously at time $t = 0$ (taking now $t = h$ as new origin). Then, writing down the nodal equations for the circuit (with bars indicating that the transform has been taken):

$$A_1C_1\begin{bmatrix}1\\-1\end{bmatrix} = \begin{bmatrix}1/R_1 + pC_1 & -pC_1\\ -pC_1 & 1/R + p(C_1 + C_2)\end{bmatrix}\begin{bmatrix}\overline{E_1}\\ \overline{E_2}\end{bmatrix}$$

Solving for $\overline{E_2}$

$$\overline{E_2} = -A_1C_1 \cdot \frac{1/R_1}{1/R_1R_2 + p(C_1/R_2) + p\dfrac{C_1 + C_2}{R_1} + p^2C_1C_2}$$

Now if one puts:

$$\omega_1 = \frac{R_1 + R_2}{C_2R_1R_2} \qquad \omega_2 = \frac{1}{(R_1 + R_2)C_1}$$

Then provided $\omega_1 \gg \omega_2$ there results:

$$-\overline{E_2} \simeq A_1/C_2R_1\frac{1}{(p + \omega_1)(p + \omega_2)} = \mathcal{L}\frac{A_1R_2}{R_1 + R_2}[\exp(-\omega_2 t) - \exp(-\omega_1 t)]$$

$$\therefore E_2 = -\frac{A(1 - \exp(-h/C_1R_1))R_2}{R_1 + R_2}[\exp(-\omega_2 t) - \exp(-\omega_1 t)] \quad \ldots\ldots \quad (8)$$

Some Applications of the Results Derived

In almost all applications of diodes to waveform circuits, i.e. in detectors, d.c. restorers, clamps, gates, and the like, errors arise in the sense that the d.c. levels established by the diode conduction differ, usually by a small amount, from the value they would have if the time-constants involved were infinite. As an example of the approximate assessment of this error consider the analysis of the saw-tooth waveform and series diode made previously. If desired no approximations need be taken and a graphical solution might be obtained to any desired degree of accuracy. If a more practical formula is desired one may take the approximation $\exp(-t/CR_2) \simeq 1 - t/CR_2$ and neglect the exponential term in $t_1 < t < 2h$ altogether. This is equivalent to taking the waveform during the non-conducting period as a linear fall and assuming that when the diode conducts the source impedance is sufficiently low to establish the linear rise immediately. Then if a milliammeter is inserted in series with R_2 the mean diode current is obtained as:

$$I_m = A/R_2 [1 - (h/CR_2)]$$

by using the indicated approximations in the results of the section dealing with the series diode response to a saw-tooth waveform. Such results are particularly useful when it is desired to obtain greater accuracy than that available in the average oscilloscope.

Consider next a case in which it is desired to measure the repetition rate of a series of rectangular pulses of fixed height and width. Such a requirement might arise, for example, in the demodulation stage of a frequency—or phase-modulation system or in the design of a 'p.r.f.-meter' for laboratory use. Referring to Fig. 2(b): in effect it is desired to measure h_2 while assuming A and $h_2 - h_1$ constant. If again a milliammeter is placed in series with R_2 the mean current I_m is given by:

$$I_m = 1/R_2h_2 \int_0^{h_2} E\,dt$$

$$= 1/R_2h_2 \int_0^{h_1} A_1 \exp(-t/CR')\,dt - 1/R_2h_2 \int_0^{h_2-h_1} A_2 \exp(-t/CR_2)\,dt$$

$$= -(1/h_2) \frac{ACR_2}{R_1+R_2} \cdot \frac{(1-\exp(-h_1/CR'))\,(1-\exp(-(h_2-h_1)/CR_2))}{1 - \exp-((h_2-h_1)/CR_2 + h_1/CR')}$$

$$= -(B/h_2) \cdot \frac{1-\alpha}{1-\alpha D}$$

where B and D are constants. It is clear that by making α sufficiently small, i.e., by making the time-constants CR' sufficiently short, the requirements will be met.

The application of the results of the section dealing with ringing will be familiar to most readers. In circuits in which the object is to produce a short, large pulse—for example 'differentiating' pulse transformers and the so-called 'multiar' circuit—such arrangements are common.

It frequently occurs that a single pulse of short duration must be derived from a comparatively wide rectangular pulse. One common method of doing this is illustrated in Fig. 6(a): C_1R_1 'differentiates' the square pulse and the diode permits only the negative 'spike' to pass into the output circuit. A serious disadvantage where the rise-time of the applied square pulse is short is the fact that the diode capacitance permits a sizeable proportion of the unwanted 'spike' to pass to the output. The only remedy is to reduce R_2, but this conflicts with other circuit requirements. Thus the circuit of Fig. 6(b) is preferable since the undesired pulse is suppressed completely. The operation is sufficiently exemplified by reference to Fig. 5 and by consideration of equation (8). R_1 may be thought to represent the amplifier output impedance and C_2 the stray capacitance. Roughly speaking, R_1 and C_2 define the rise-time of the circuit, C_1 and R_2 define the differentiating time-constant: it is also necessary to make R_1C_1 sufficiently small to enable C_1 to acquire sufficient charge during the diode conduction. The arrangement may be easily adapted to provide positive or negative outputs from the leading or trailing edges of the input pulse. There are no great changes in the analytical results in each case.

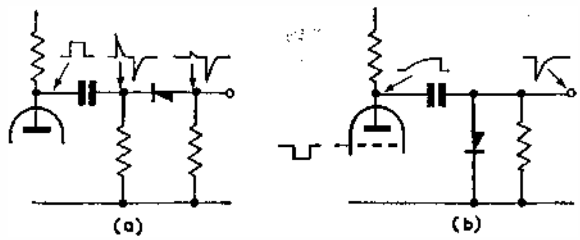

Fig. 6. (a) A common method of obtaining a single spike from a square pulse
(b) A better method of achieving the same effect

Acknowledgment

The author is indebted to the Directors of E.M.I. Electronics Ltd. for permission to publish this article.

A Development in H.F Transmitters

Marconi's Wireless Telegraph Company Ltd have announced the successful development and engineering of a new system of high-frequency radio transmission. A way has been found of providing wideband amplification over the whole h.f. band, thereby making it possible to dispense with all forms of operational tuning in the transmitter's h.f. amplifier stages. At the same time it permits the simultaneous radiation of two or more independent transmissions from one equipment.

The new technique overcomes what has always been a major drawback with conventional transmitters—the complex stage-by-stage retuning of the h.f. amplifier whenever it is desired to change frequency. The dual-transmission aspect is a further important feature in that it enables a service to be transferred to a new frequency before the old one is relinquished.

The first transmitter in the world to embody these features is the Marconi HS113, a 1kW wideband h.f. communication equipment. This has been designed to provide either single/independent sideband telephony, low-level double-sideband telephony, frequency-shift telegraphy or c.w. on-off telegraphy. It will operate at any frequency (or simultaneously on a number of frequencies) in the range of 2 to 24Mc/s.

It provides a transmitter which is extremely simple to operate and one in which mechanical failure is virtually impossible because there are no moving parts in the amplifier. By reason of its use of multi-wave techniques the failure of a valve (other than by short-circuit) does not put the equipment out of service but merely results in a slight reduction of transmitter output.

Basically, the HS113 consists of a distributed amplifier and its associated drive equipment. The unique feature is the distributed amplifier, for although this type of circuit is well known as a low-level voltage amplifying device (for instance, in oscilloscopes), the principle has never before been successfully used for amplification for this type of service.

An inherent feature of the design of the distributed amplifier is that it is insensitive to the nature of the load impedance even when reactance is present to a degree corresponding to quite high standing wave ratios. As a practical instance, a 2 : 1 standing wave ratio on the output feeder is quite permissible, producing only a very slight reduction in output power.

To change frequency on this transmitter involves nothing more than the operation of a switch on the drive unit. Aerial commutation in the transmitter automatically selects the correct array for the desired frequency.

As the modulation is applied at low level, it is possible to install the drive equipment at a considerable distance from the amplifier, with distribution by coaxial cable. This is particularly advantageous for comprehensive station working, as the drives can be installed at a control centre and the amplifiers at any convenient point. One engineer can then control a number of services from the central position. In this connexion of comprehensive station working the HS113 is an attractive replacement for existing high-speed and channelized transmitters (especially for single-sideband or telegraph traffic) on the grounds of speed of operation, lower cost, reliability and ease of maintenance.

A Transistor Blocking Oscillator Frequency Divider

By F. Butler, B.Sc., M.I.E.E., M.Brit.I.R.E.

The frequency divider includes a 'staircase' waveform generator designed to produce a multiple step waveform in which all the voltage increments are equal in amplitude. After a prescribed number of steps, the output is used to trigger a blocking oscillator which gives a recurrent pulse output at a fixed sub-multiple of the input step frequency. The frequency divider operates from a source giving a sinusoidal output waveform which is squared and limited in order to make the mode of operation independent of the amplitude of the input signal. An output amplifier follows the blocking oscillator. Transistors and diodes of readily available types are used throughout.

(Voir page 637 pour le résumé en français: Zusammenfassung in deutscher Sprache Seite 644)

AN earlier contribution[1] described a transistor regenerative modulator frequency divider giving outputs of 10kc/s and 1kc/s from a 100kc/s crystal oscillator. For use at lower frequencies a triggered blocking oscillator is simpler, more reliable and is less affected by variations in the characteristics of transistors due to temperature or supply voltage changes. Blocking oscillators work well when the output frequency is a small sub-multiple of the drive frequency but are liable to miscount at high division ratios. The reliability of triggering can be much improved by making use of a circuit which converts the sinusoidal primary drive source into a stepped 'staircase' waveform. Two known circuits enable the required transformation to be effected. In one of these the steps in the waveform show an exponential decrease in amplitude. In the second circuit, provision is made for generating a linear staircase in which all the steps are of equal amplitude.

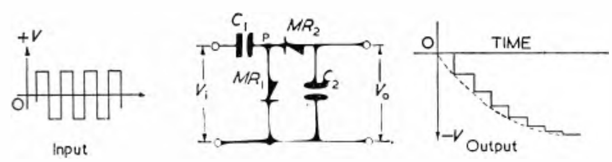

Fig. 1. Diode pump circuit and waveforms

The Diode Pump Circuit

Fig. 1 shows the simplest possible version of the diode integrator circuit[2]. The input is a square wave voltage of fixed amplitude which can be derived from a sinusoidal source in a number of different ways. The first negative excursion of the input voltage causes MR_2 to conduct, placing C_1 and C_2 in series across the input signal source. The two capacitors carry equal charges but share the input voltage in the inverse ratio of their capacitances. On the positive half cycle of the input voltage there is a rise in the potential of the point P, MR_2 is cut off, isolating C_2 from the input and allowing it to retain its initial charge. Meanwhile, MR_1 conducts and discharges C_1. The action is repeated but owing to the retention of charge on C_2 the successive increments of voltage across it get less and less. The waveform across C_2 is in fact a quantized exponential. By using a cathode-follower valve, or a transistor in a common collector mode, it is possible to equalize the voltage increments across C_2, giving a linear 'staircase' waveform[3]. The initial action is the same as in the diode pump arrangement. On the positive half cycle the transistor in Fig. 2 conducts, behaves as an emitter-follower and keeps the point P at a potential almost equal to that of Q. On the arrival of the next negative input pulse, MR has a negligibly small reverse bias to oppose the input so that the full input voltage is available for charging C_1 and C_2. This condition holds no matter what the potential across C_2 may be and as a result all the voltage increments are equal.

With a pnp transistor and with the diode polarity shown in Fig. 2 the output staircase waveform is negative with respect to the earth line. A positive waveform can be produced by reversing the diode connexion and using an npn transistor.

For use as frequency dividers it is possible to devise a number of pulse generator circuits which can be triggered recurrently after any prescribed number of cycles of the input drive voltage to the staircase waveform generator. Thyratrons, multivibrators and blocking oscillators are good examples of such circuits. Quite clearly, the reliability of triggering at the sub-multiple frequency, and hence the reliability of frequency division, is much greater in the case of a linear staircase waveform than in the case where the voltage increments diminish exponentially. This is particularly true when large division ratios are desired.

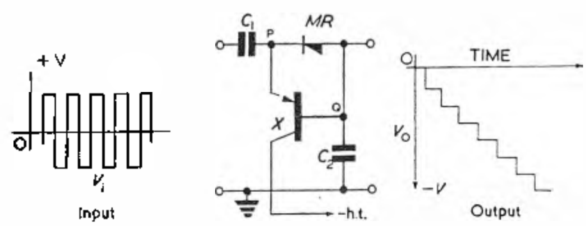

Fig. 2. Transistor circuit to give linear staircase waveform

The transistorized escalator circuit is particularly suitable for working in conjunction with a transistor blocking oscillator, and the combination makes an extremely reliable divider circuit.

Practical Frequency Divider Circuit

The complete circuit diagram of a frequency divider giving a pulse output at 100c/s from an input at 1kc/s is shown in Fig. 3.

The first transistor stage is a clipper-amplifier which delivers a squared output voltage of constant amplitude regardless of possible wide variations of the input signal level. The input is a sine wave voltage at 1kc/s.

The stepped waveform generator driven by X_1 comprises the transistor X_2, the diode MR_1 and the two capacitors C_1 and C_2. The negative staircase waveform appears across C_2.

The transistor X_3 acts in conjunction with the transformer T as a blocking oscillator. Sufficient reverse bias to prevent untriggered self-oscillation is provided by the 50kΩ variable resistor and a fixed resistor of 1·5kΩ connected in the emitter circuit.

When an input signal is applied to X_1, the potential across

C_2 increases in a succession of equal negative steps until the negative base bias on X_3 reaches a critical level sufficient to allow collector current to flow. Regenerative feedback from the collector winding to the base winding of the transformer T then operates to drive X_3 into collector current saturation. The transistor is bottomed and C_2 begins to discharge through the transformer secondary winding, reducing the collector current and inducing a reverse voltage in the primary winding. The diode MR_2 serves to suppress this voltage since an output in the form of a train of unidirectional pulses is required. In practice, the discharge of C_2 takes place in a time which is shorter than the duration of a single cycle of the input trigger voltage. The output pulses from the blocking oscillator are of a correspondingly short duration.

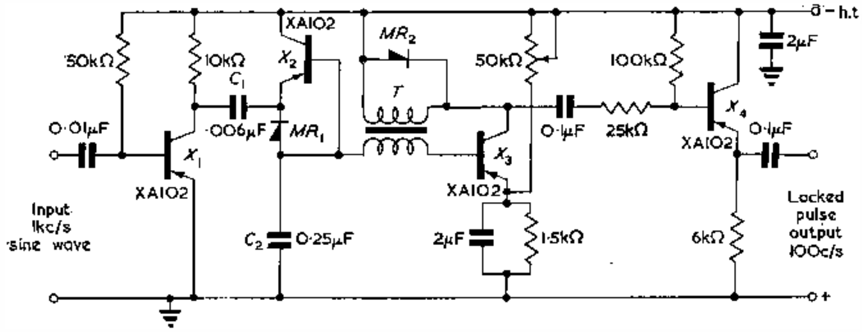

Fig. 3. Blocking oscillator triggered by staircase waveform generator

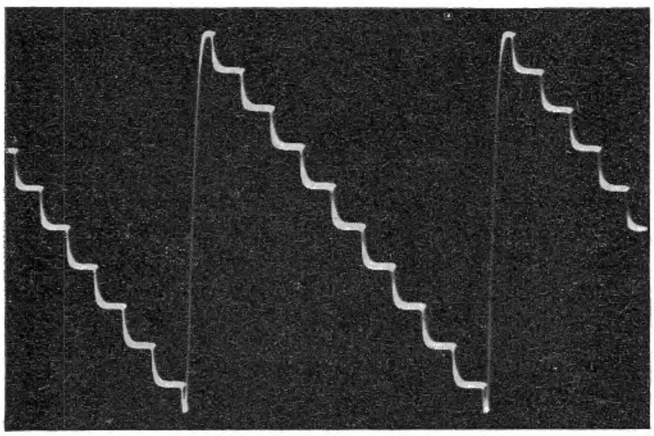

Fig. 4. Output of stepped waveform generator (*Input 1kc/s, division ratio 10*)

Referring again to Fig. 3 it will be seen that an additional transistor stage X_4 follows the blocking oscillator X_3. This last stage is an emitter-follower amplifier having a low output impedance. It serves to isolate the blocking oscillator so that its performance is almost unaffected by any externally connected load.

If desired the output amplifier may be replaced by a pair of transistors connected as a triggered multivibrator and arranged to deliver a square wave output of unity mark-to-space ratio instead of the train of short duration pulses derived from the blocking oscillator. The output from such a multivibrator is suitable for driving another staircase waveform generator with an associated blocking oscillator. Operation down to 10c/s is entirely satisfactory with standard components and transistors.

Practical tests on the circuit of Fig. 3, operating with an input of 1kc/s and a division ratio of 10, show that the performance is reliable over a large range of ambient temperature and is not sensitive to supply voltage changes. Where extreme variations of battery voltage are to be expected it might be worth stabilizing the supply by using a Zener diode in a simple regulator circuit.

Fig. 4 shows the stepped waveform across C_2 in Fig. 3. The extreme linearity of the staircase voltage and the sharpness of the trigger action will be apparent.

Selection of Components

Referring again to Fig. 3 there are only three components which are in any way critical. These are the capacitors C_1 and C_2 and the blocking oscillator transformer T. The absolute value of C_1 is determined by the input frequency while the ratio C_1/C_2 depends on the required division ratio. For an input of 1kc/s and a division ratio of 10, values of 0·006 and 0·25μF will normally be suitable. With higher input frequencies, proportionately smaller capacitances will be required. At lower frequencies, correspondingly larger capacitances are necessary.

As regards the transformer, any component having a collector winding inductance of a few henries will prove suitable for operation at 100c/s. The winding resistance is unimportant. The base circuit winding should be of lower inductance than the primary coil, the transformer turns ratio being about 2 : 1. The core material may be silicon steel or nickel alloy. For operation at 10c/s a standard intervalve audio frequency transformer with a turns ratio of 3 : 1 or 5 : 1 may be employed, provided that the winding resistances are not excessively high.

The diodes MR_1 and MR_2 may be of almost any available type. Point contact germanium units were used in the prototype circuit but germanium or silicon junction diodes are equally suitable. There are no stringent requirements to be met as regards forward or reverse resistance and the peak inverse voltage is very small. The transistors are all Ediswan-Mazda Type XA102 but any equivalent types may be substituted.

Alternatives to the Blocking Oscillator

Staircase waveform generators have been used to trigger a biased thyratron instead of a blocking oscillator[4]. Development work in the U.S.A. has led to the production of semiconductor devices having thyratron-like characteristics. In the U.K., work on this subject is currently being carried out and it seems likely that such a device would be readily adaptable for use in a frequency divider circuit of the type already described. Use of a semiconductor thyratron would eliminate the need for the transformer which is the one objectionable feature of the blocking oscillator divider.

Attempts to check on this possible application of the new device must await the availability of sample production quantities.

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Diode Phase-Sensitive Detectors with Load

R. Chidambaram*, M.A., M.Sc., A.M.I.R.E., and S. Krishnan*, M.Sc.

A theoretical investigation of the operation of the simple diode push-pull phase-sensitive detector with load is carried out. The transfer ratios for the two diodes are found to vary considerably with the signal. The non-linearity in the output due to these variations is evaluated and a table is given from which the suitability of a given detector may be judged immediately. Experiments confirm quantitatively the theoretical results.

(Voir page 637 pour le résumé en français: Zusammenfassung in deutscher Sprache Seite 644)

THE simple push-pull phase-sensitive detector† finds application in various fields such as the generation of very slow sine waves, nuclear magnetic resonance studies, balance indication on a.c. bridges, etc.[1,2]. Though for some purposes, linearity of the detector is not essential, there are other cases where the linearity is of prime importance. The connexion of a load (which may be the measuring instrument itself) between the output terminals of the detector introduces a non-linearity, an effect not mentioned in the literature. This article deals with a theoretical investigation of this effect and the experimental verification of the conclusions reached.

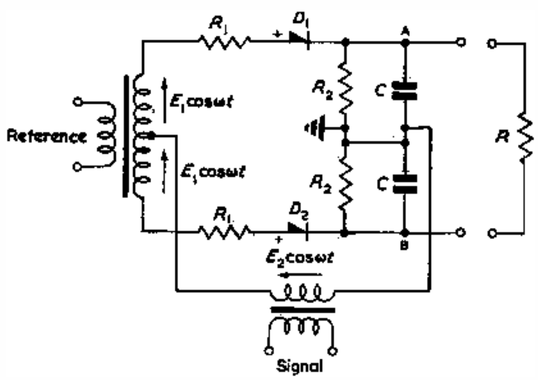

Fig. 1. Circuit of the simple push-pull phase-sensitive detector

The Simple Push-pull Phase-sensitive Detector

Fig. 1 shows the circuit of the detector. E_1 and E_2 are the amplitudes of the reference and signal voltages respectively, assumed to be in phase, as is customary. The sum of the two voltages is applied to one of the diodes, D_1, and the difference to the other diode, D_2. R_1 are the effective series resistances in the two diode circuits. If the value CR_2 is very large compared to the period of the applied sine waves, the voltages E_A and E_B developed at the output nodes, A and B, may be assumed to be steady. The output voltage is denoted by

$$E_o = E_A - E_B \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (1)$$

In the absence of the external load R

$$E_A = k_o\,(E_1 + E_2),\; E_B = k_o\,(E_1 - E_2),\; E_2 < E_1$$

$$E_o = 2k_oE_2 \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (2)$$

The transfer ratio

$$k_o = \cos\,\theta_o. \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (3)$$

where $2\theta_o$ is the angle of conduction of either diode, θ_o being given by the following equation derived by Farren[3].

$$R_2/R_1 = \frac{\pi\,\cos\,\theta_o}{\sin\,\theta_o - \theta_o\,\cos\,\theta_o} \quad \ldots\ldots\ldots\ldots \quad (4)$$

* *Indian Institute of Science, Bangalore.*

† *This detector has been commonly called a push-pull detector even though the output in general is not balanced.*

If now the external load R is connected across the output terminals, a current will flow through it from node A to node B. The effective resistance between node A and ground will consequently decrease while that between node B and ground will increase. This will make the transfer ratio for D_1 decrease and that for D_2 increase.

Suppose that the diodes, D_1 and D_2, conduct respectively over angles $2\theta_1$ and $2\theta_2$ during a cycle, their transfer ratios, k_1 and k_2, can be written down as

$$k_1 = \cos\,\theta_1,\; k_2 = \cos\,\theta_2.$$

$$E_A = k_1(E_1 + E_2),\; E_B = k_2(E_1 - E_2) \quad \ldots\ldots\ldots \quad (5)$$

and

$$E_o = k_1(E_1 + E_2) - k_2(E_1 - E_2) \quad \ldots\ldots\ldots \quad (6)$$

Since, in general, k_1 and k_2 are unequal, the output E_o is not proportional to the signal amplitude E_2.

Writing $x \equiv E_2/E_1$, equation (6) becomes

$$E_o = E_1\left\{ k_1(1 + x) - k_2(1 - x)\right\} \ldots\ldots \quad (7)$$

The average current flowing through D_1 into node A is

$$\frac{1}{2\pi R_1}\int_{-\theta_1}^{\theta_1} E_1(1 + x)\,(\cos\,\omega t - k_1)\,d(\omega t)$$

$$= \frac{E_1(1 + x)}{\pi R_1}[\sin\,\theta_1 - k_1\theta_1]$$

This current may be equated to the current flowing out from node A through R_2 and R.

$$\frac{E_1(1 + x)}{\pi R_1}\cdot(\sin\,\theta_1 - k_1\theta_1) = \frac{k_1E_1(1 + x)}{R_2} + \frac{k_1E_1(1 + x) - k_2E_1(1 - x)}{R} \ldots\ldots \quad (8)$$

Equation (8) and the corresponding equation for node B can be written as

$$n_1/\pi\,(\sin\,\theta_1 - \theta_1\cos\,\theta_1) = \cos\,\theta_1(1 + n_2) - wn_2\cos\,\theta_2 \quad \ldots \quad (9)$$

and

$$n_1/\pi\,(\sin\,\theta_2 - \theta_2\cos\,\theta_2) = \cos\,\theta_2(1 + n_2) - (n_2/w)\cos\,\theta_1,\; w \neq 0 \quad \ldots\ldots\ldots \quad (10)$$

where

$$n_1 \equiv R_2/R_1,\; n_2 \equiv R_2/R \;\text{ and }\; w \equiv \frac{1 - x}{1 + x} \ldots \quad (11)$$

n_1 and n_2 may for obvious reasons be called the *source impedance factor* and the *load factor* respectively.

For given values of n_1, n_2 and w (or x), a pair of simultaneous transcendental equations have to be solved in θ_1 and θ_2. The method of solution is given in the Appendix.

Behaviour of the Transfer Ratios k_1 and k_2

The calculations have been made for the source impedance factor $n_1 = 1{\cdot}46\pi$, this value being recommended[4] for the elimination of third harmonic distortion because when used in equation (4) it leads to an angle of conduction $2\theta_o = 120°$. (As will be clear below, however, this condition is not really satisfied for either diode for any finite value of the signal).

Fig. 2 gives the variation of k_1 and k_2 against the signal-to-reference voltage ratio x for different values of the load factor n_2. It will be noted that in all cases, in the absence of the signal (i.e. $x = 0$), $k_1 = k_2 = k_0$ (equation (3)). This is also otherwise obvious since $E_A = E_B$ here and no current flows through R. As the signal voltage (or x) is increased, k_1 decreases and k_2 increases continuously. The deviations of k_1 and k_2 from k_0 are more pronounced for larger values of the load factor n_2 (i.e. lower load resistances) and, incidentally, also for smaller values of the source impedance factor n_1.

For any value of n_2, there can be seen to exist a value of x at which k_2 reaches unity, corresponding to zero angle of conduction of D_2. This occurs when the voltage transferred

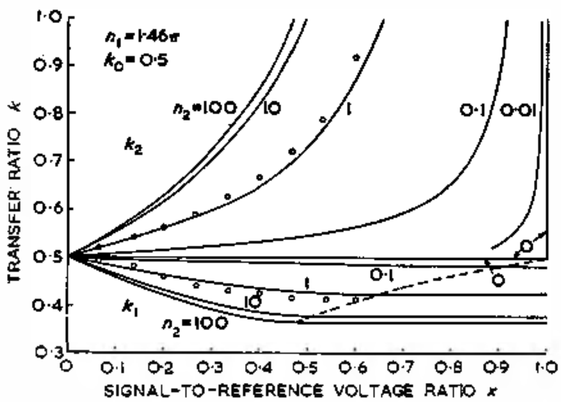

Fig. 2. Variation of the transfer ratios of the two diodes with the signal

The broken line is the locus of $k_{1(crit)}$ the value of k_1 where D_2 stops conducting. Experimentally observed points for $n_2 = 1$ are indicated by circles

from the node A to the node B through R equals the peak value of the voltage applied to D_2. For still higher values of x, D_2 continues to be non-conducting and E_B is obtained from E_A through potential division between R and R_2. E_0 is now just a fraction, $1/(n_2 + 1)$, of $k_{1(crit)} E_1 + x$), where $k_{1(crit)}$ is the value of k_1 at the point where D_2 stops conducting.

From equation (7), it can be seen that the ratio of the output d.c. voltage E_0 to the signal voltage amplitude E_2, which may be called 'sensitivity' of the detector, is given by

$$\alpha \equiv E_0/E_2 = k_2 + k_1 - \frac{k_2 - k_1}{x} \quad \ldots\ldots\ldots (12)$$

It is interesting to note that even though k_1, $k_2 \to k_0$ as $x \to 0$, α does not tend to $2k_0$ because $(k_2 - k_1)x$ tends to a finite positive limit. The limit $\alpha_{x=0} \equiv \alpha_0$ has been theoretically evaluated in the Appendix.

Non-Linearity of the Detector

It might appear from an examination of Fig. 2 that the wide variations in k_1 and k_2 might cause correspondingly large variations in the sensitivity α (making the detector highly non-linear). However, it turns out that the non-linearity is not pronounced so long as the diode D_2 does not cease conducting. This is due to the fact that the variations in k_1 and k_2 tend to compensate each other partially.

It has been found that the graphical method described in the Appendix, although accurate enough for the determination of k_1 and k_2, is not satisfactory for estimating this small non-linearity. It is, however, possible to evaluate accurately the sensitivity, α, at $x = 0(\alpha_0)$ and at $x = x_{(crit)}(\alpha_{(crit)})$, where D_2 stops conducting. This has been done in the Appendix. Since it is positively undesirable to use the detector for $x > x_{(crit)}$, the total non-linearity N_0 may be defined as

$$N_0 \equiv 100\,\frac{\alpha_0 - \alpha_{(crit)}}{\alpha_0} \text{ per cent} \quad \ldots\ldots\ldots\ldots (13)$$

Values of N_0 calculated in this manner are given in Table 1 for various values of n_1 and n_2, along with the corresponding values of α_0 and $x_{(crit)}$. The suitability of a detector can be judged immediately from this Table.

Experimental Verification

All the significant results of the above analysis have been verified experimentally for the value of source impedance factor $n_1 = 1{\cdot}46\pi$ and load factor $n_2 = 1$.

DETERMINATION OF TRANSFER RATIOS k_1 and k_2

Diode-connected 6AC7's were used for D_1 and D_2 because of their low forward resistance. The signal and reference voltages were obtained from the 50c/s mains.

TABLE 1

n_1 \ n_2	100	30	10	3	1	0·3	0·1	0·03	0·01
1	0·00309 0·774 5·5%	0·0101 0·776 5·0%	0·0290 0·783 4·7%	0·0835 0·807 4·2%	0·181 0·851 3·2%	0·306 0·917 1·8%	0·381 0·964 0·8%	0·417 0·988 0·3%	0·429 0·996 0·15%
1·46 π ..	0·0125 0·470 9·8%	0·0405 0·477 9·6%	0·113 0·495 9·3%	0·297 0·564 7·5%	0·558 0·654 5·1%	0·808 0·809 2·4%	0·926 0·916 0·9%	0·977 0·973 0·4%	0·993 0·990 0·15%
10	0·0239 0·322 11.8%	0·0763 0·330 11·4%	0·205 0·352 9·8%	0·498 0·421 8·0%	0·843 0·555 4·9%	1·11 0·753 2·1%	1·23 0·892 0·8%	1·27 0·964 0·3%	1·28 0·988 —
30	0·0550 0·174 13·5%	0·170 0·183 12·4%	0·421 0·213 10·7%	0·870 0·292 7·2%	1·25 0·454 3·8%	1·48 0·696 1·6%	1·56 0·846 0·7%	1·59 0·954 —	1·60 0·985 —
100	0·127 0·084 13·9%	0·363 0·095 12·5%	0·777 0·125 9·4%	1·29 0·213 5·0%	1·60 0·391 2·2%	1·74 0·659 0·9%	1·78 0·849 0·3%	1·80 0·948 —	1·81 0·982 —

The set of three numbers in each square gives in that order the limiting sensitivity α_0, rounded off to three significant figures, the critical value of x, x_{crit}, where the diode D_2 stops conducting and the total non-linearity N_0 in the range $O < x < x_{crit}$.

The component values were:

$$R_2 = 240\text{k}\Omega$$
$$R_1 = 52{\cdot}4\text{k}\Omega$$
$$R = 240\text{k}\Omega$$
$$C = 8\mu\text{F}$$

The experimental points are marked in Fig. 2 and are seen to fit the corresponding theoretical curves closely.

Determination of Non-linearity

In this case, the signal voltage was maintained constant and the reference voltage varied since the change in the output voltage then indicates directly the non-linearity. Also the meter does not need to be switched to different ranges with consequent variations in loading and sensitivity. The sensitivity α has been plotted in Fig. 3. From

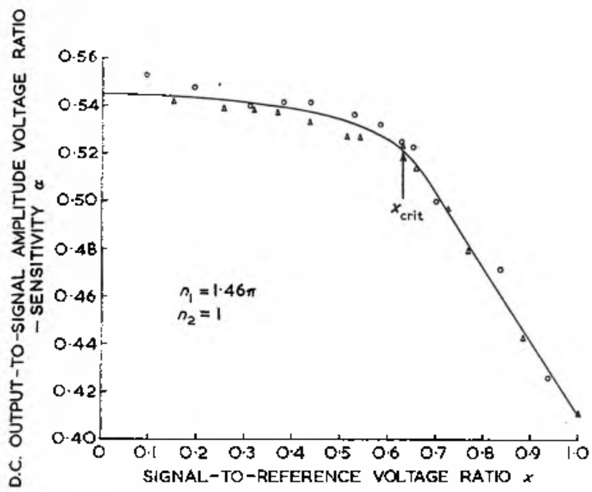

Fig. 3. Variation of the sensitivity α with the signal. *The value of $x_{(crit)}$ indicated has been obtained from Fig. 2*

the curve, using the value of $x_{(crit)}$ obtained in the previous experiment

$\alpha_o = 0{\cdot}545 \quad \alpha_{(crit)} = 0{\cdot}522,$

total non-linearity $N_o = 4{\cdot}2$ per cent.

The corresponding theoretically predicted values (taken from Table 1) are

$\alpha_o = 0{\cdot}558 \quad \alpha_{(crit)} = 0{\cdot}530, \quad N_o = 5{\cdot}1$ per cent.

Appendix

(a) Solution of the Equations (9) and (10) for $\cos\theta_1, \cos\theta_2$

From equations (9) and (10), one immediately obtains

$$w = -(F(\theta_1)/F(\theta_2)) \quad \ldots\ldots\ldots\ldots (14)$$

and

$$n_2 = 1 \Big/ \left\{\frac{\cos\theta_1}{F(\theta_1)} + \frac{\cos\theta_2}{F(\theta_2)}\right\} \quad \ldots\ldots\ldots\ldots (15)$$

where

$$F(\theta) \equiv (n_1/\pi)(\sin\theta - \theta\cos\theta) - \cos\theta \quad \ldots\ldots (16)$$

It is not possible to solve for θ_1 and θ_2 for given n_2 and w in a straightforward manner. Hence a graphical method of solution is adopted. For a given value of n_1, the functions $F(\theta)$ and $F(\theta)/\cos\theta$ are plotted against $\cos\theta$ for $0 \leqslant \theta \leqslant \pi/2$. It is known that for $x = 0$, the angles of conduction of the two diodes are equal and, as seen from equations (9) and (10), are the single root of the equation

$$F(\theta) = 0, \quad 0 \leqslant \theta \leqslant \pi/2 \quad \ldots\ldots\ldots\ldots (17)$$

From the physical arguments provided in the text, it is clear that for any n_2 ($\neq 0$) and $w(\neq 1)$

$$\theta_1 > \theta_o > \theta_2 \quad \ldots\ldots\ldots\ldots\ldots\ldots (18)$$

Hence to obtain a plot such as is given in Fig. 2, a value of $\cos\theta_1$ is chosen such that $\theta_1 > \theta_o$, the corresponding $F(\theta_1)/\cos\theta_1$ is read from the curve and using equation (15), $F(\theta_2)/\cos\theta_2$ is calculated for the chosen value of n_2 and the corresponding value of θ_2 is read from the curve. The value of w consistent with this set (θ_1, n_2, θ_2) is determined by reading the values of $F(\theta_1)$ and $F(\theta_2)$ from the $F(\theta)$ plot and using equation (14).

Note: $F(\theta)/\cos\theta$ and not $\cos\theta/F(\theta)$ is plotted because the latter has a discontinuity at θ_o, approaching the limits $+\infty$ and $-\infty$ from the θ_o+ and θ_o- sides respectively.

(b) Evaluation of $\alpha_o \equiv \underset{x=0}{\text{Limit}} \dfrac{k_1(1+x) - k_2(1-x)}{x}$

For a given small value of x, let the changes, $\Delta\theta_1$ and $\Delta\theta_2$, in the semi-angles of conduction of D_1 and D_2 be δ and $-p$ respectively where δ and p are small positive quantities. The signs have been so chosen because of condition (18).

Applying Taylor's theorem, it is easily shown that

$$F(\theta_o + \Delta\theta) \simeq a\Delta\theta + b(\Delta\theta)^2 \quad \ldots\ldots\ldots\ldots\ldots (19)$$

and

$$\frac{F(\theta_o + \Delta\theta)}{\cos(\theta_o + 6\Delta\theta)} \simeq c\Delta\theta + d(\Delta\theta)^2 \quad \ldots\ldots\ldots\ldots (20)$$

neglecting higher powers of $(\Delta\theta)$, where

$$\left.\begin{aligned}
a &\equiv ((n_1\theta_o/\pi) + 1)\sin\theta_o \\
b &\equiv \tfrac{1}{2}\left\{(n_1/\pi)\sin\theta_o + ((n_1\theta_o/\pi) + 1)\cos\theta_o\right\} \\
c &\equiv ((n_1\theta_o/\pi) + 1)\tan\theta_o \\
&\text{and} \\
d &\equiv \tfrac{1}{2}\left\{(n_1/\pi)\tan\theta_o + ((n_1\theta_o/\pi) + 1)(\tan^2\theta_o + \sec^2\theta_o)\right\}
\end{aligned}\right\} (21)$$

From equations (14) and (11)

$$x \simeq \frac{a(p - \delta) - b(p^2 + \delta^2)}{a(p + \delta) - b(p^2 - \delta^2)} \quad \ldots\ldots\ldots\ldots (22)$$

From equation (5), using Taylor's theorem

$$\left.\begin{aligned}
k_1 &\simeq k_o - (\sin\theta_o)\,\delta - k_o\delta^2 \\
k_2 &\simeq k_o + (\sin\theta_o)\,p - k_o p^2
\end{aligned}\right\} \quad \ldots\ldots\ldots\ldots\ldots\ldots (23)$$

From equations (22) and (23) and the definition of α_o

$$\alpha_o = 2k_o - \underset{\delta\to 0}{\text{Limit}} \frac{a(p + \delta)^2 \sin\theta_o}{a(p - \delta) - b(p^2 + \delta^2)} \quad \ldots\ldots\ldots\ldots (24)$$

From equations (15) and (20) one can solve for p in terms of δ to obtain

$$p = \delta + ((c/n_2) + (2d/c))\,\delta^2 + 0(\delta^3) \quad \ldots\ldots (25)$$

From equations (24), (25) and (21)

$$\alpha_o = \frac{2k_o}{1 + \dfrac{2n_2}{((n_1\theta_o/\pi) + 1)}} \quad \ldots\ldots\ldots\ldots (26)$$

It may be mentioned that the procedure for the accurate determination of θ_o is similar to the one described below for calculating $\theta_{1(crit)}$.

(c) Evaluation of $\alpha_{(crit)}$

It will be noted that at $x = x_{(crit)}$ when the diode D_2 stops conducting, the equivalent load resistance, R_{eq}, for the diode D_1 consists of the parallel combination of R_2 and $(R + R_2)$. The semi-angle of conduction of D_1, $\theta_{1(crit)}$, at this point is a solution of equation (4) with R_{eq}/R_1 for R_2/R_1. This equation has also to be solved graphically[3]. A single curve giving $\theta_{1(crit)}$ against R_{eq}/R_1 can serve for all phase-sensitive detectors. The value of $\theta_{1(crit)}$ obtained graphically may be refined using Taylor's theorem.

One can now make use of the fact, already mentioned, that at the critical point, the voltage at node B obtained through potential division between R and R_2 is equal to the peak value of the voltage applied to the diode D_2. That is

$$k_{1(crit)}\,(1 + x_{(crit)})\,\frac{R_2}{R + R_2} = (1 - x_{(crit)})$$

remembering that $k_{2(crit)} = 1$. Using equation (12),

$$x_{(crit)} = \frac{2k_{1(crit)}}{n_2 + 1 - n_2 k_{1(crit)}} \qquad \text{.......... (27)}$$

Acknowledgments

The authors wish to thank Prof. R. S. Krishnan for his kind interest and encouragement and Dr. G. Suryan for many helpful discussions. One of the authors (S.K.) is grateful to the Council of Scientific and Industrial Research for a research assistantship and the other (R.C.) to the Department of Atomic Energy for a research fellowship.

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Television News Films by Transatlantic Telephone Cable

The BBC Engineering Division has developed a system for transmitting brief television news picture sequences and other short television film sequences of up to one minute's duration over a circuit of the transatlantic telephone cable normally used for sound. These picture transmissions can be sent over the cable in both directions.

The process employs a slow-speed flying-spot film scanner, the video signal from which is used to modulate a carrier for transmission over the cable. At the receiving end the signals are demodulated and used to operate a slow-speed film telerecording equipment.

Consideration of the characteristics of the Atlantic cable indicated that a maximum video frequency of 4·5kc/s could be used. It has therefore been necessary to effect as many economies in the bandwidth of the video signal as are compatible with acceptable picture quality. These economies are:

(1) Restriction of the horizontal definition to that corresponding with a bandwidth of 1·75Mc/s in the 405-line television system.
(2) A reduction to 200 lines using sequential scanning.
(3) The scanning at the transmitting end of only alternate film frames with each frame-scan reproduced on two adjacent film frames at the receiving terminal.

These measures result in reducing the 3Mc/s bandwidth of the British television system to approximately 450kc/s, the remainder of the bandwidth reduction being obtained by a decrease of the scanning speed until the maximum video frequency corresponds with the available 4·5kc/s upper limit. The time required to scan the film is approximately 100 times normal.

16mm film has been chosen because it is almost universally used for television news. The average length of television newsreel sequences is less than half a minute and the new system will make it possible to transmit facsimiles of these sequencies in much less time than it would take to fly them across the Atlantic. The speed of transmission of each film frame is about 75 times faster than previous methods of sending still pictures by facsimile transmission.

The effective picture repetition frequency of 12½ per second results in satisfactory reproduction of most material excepting that in which rapid movement occurs. In this case special provision is made for the transmission of every frame of the original film. This of course doubles the transmitting time.

The new system uses a channel of the type normally used for transmitting music over the cable; such a channel has a nominal bandwidth of 6·4kc/s. In order to limit the variation in the group delay/frequency characteristic to a value which can be corrected, it is necessary to restrict the usable video bandwidth to 4·5kc/s.

Vestigial sideband transmission is used with a special form of negative-going amplitude modulation. The carrier frequency is 5kc/s and the whole of the lower sideband is transmitted, the vestige of the upper sideband extending from 5kc/s to 5·5kc/s.

An additional problem results from the need to remove from the circuit the volume-range compressors and expanders which are normally used. This makes it more difficult to achieve a satisfactory signal-to-noise ratio and therefore a special form of amplitude modulation has been used in which the maximum depth of modulation considerably exceeds 100 per cent. That part of the modulation envelope which extends beyond the normal 100 per cent, or zero carrier, condition is "folded back" in a positive-going direction. This method of modulation results in an increase in the effective depth of modulation and thus also in the signal-to-noise ratio of the system. In order to achieve the synchronous detection needed with this type of signal it is necessary to use a re-generated carrier at the receiving terminal and this must be locked in phase to the original transmitted carrier. The necessary bursts of carrier which occur during the synchronizing signals are used for this purpose.

Another problem which has necessitated careful design in the equipment results from the fact that the line scanning frequency of 25c/s makes it impossible to reduce the effects of mains hum by means of clamping circuits. It has therefore been necessary to keep hum through the system at an extremely low level. Should trouble from hum on the transatlantic circuit be experienced some benefit can be achieved by a further reduction of line scanning frequency at the cost of a slight increase in transmission time. The precise choice of frequency will depend upon whether the hum is predominantly 50c/s or 60c/s.

As in other television systems a synchronizing signal is transmitted at the beginning of each line-scanning period, in this case the full amplitude of the video signal being utilized for the triggering edge of the synchronizing signal. The field synchronizing signal consists of four similar pulses and protection is provided against these pulses interfering with the bursts of reference carrier which are used for oscillator locking.

Negative film will normally be used at the sending terminal, but the equipment can deal with either negative or positive film.

Identical film equipments are used at both terminals of the system. At the sending end this apparatus operates as a flying-spot scanner while at the receiving end it functions as a telerecording channel. The same cathode-ray tube is used for both purposes and is enclosed in a double mu-metal shield in order to minimize mains frequency interference.

The time required for each field scan is approximately 8sec; a separate monitor tube with a long persistence phosphor reproduces a recognizable picture.

The special film traction mechanism, which is operated by the synchronizing signal, pulls down two film frames at a time. Twin optical systems are needed to telerecord simultaneously on two adjacent film frames and very small lenses with a focal length of one inch and an aperture of $f/8$ were developed specially for this purpose.

Short News Items

Mullard Equipment Ltd are building a 4 MeV linear accelerator for X-ray treatment of deep-seated tumours for the Cancer Institute Board of Victoria. Valued at £60 000, it will be installed at the Board's Peter MacCallum Clinic in Melbourne in the middle of next year. The contract represents the fourth medical linear accelerator to be built by the Company, the first three having been for U.K. hospitals. The accelerator will give an X-ray output more than ten times greater than that of a normal, 250kV therapy machine, and will enable treatments to be given at the rate of 60 a day—about three times the number possible with the conventional apparatus. The higher voltage rating of the accelerator also produces an X-ray beam of greater penetration, making it possible to reach tumours seated too deep for effective treatment with conventional machines. Damage to the patient's skin, which can be a serious problem when low-voltage X-rays are used, is virtually eliminated, and there is less side-scattering of the beam during its passage through the body. The accelerator is mounted on a rotatable gantry, which enables the X-ray beam to be directed precisely to the diseased area and at exactly the required angle.

The European Airlines Electronics Committee held their first 'maintenance meeting' in London on 22 and 23 September. It was attended by radio and electronic representatives of eleven airlines, namely, Air France, Alitalia, DLH, Finnair, KLM, SAS, Sabena, Swissair, UAT, BEA and BOAC. By invitation, representatives attended from the Air Transport Electronics Council, the Electronic Engineering Association and the Society of British Aircraft Constructors. A representative of the Collins Radio Company of England was also present. The meeting was organized on behalf of the committee by British Overseas Airways Corporation, with the participation of British European Airways. The chairman was Mr. R. D. Jones, electrical systems development superintendent of BOAC.

A Demonstration of the Lear Automatic Pilot was recently given at the De Havilland airfield at Hatfield, which is equipped with a Pye Instrument Landing System. The demonstration showed the i.l.s. coupled approach flying an aircraft with precision down to a height of 20ft above the runway. The Pye i.l.s. installation at Hatfield incorporates the latest modifications resulting from developments in co-operation with the Blind Landing Experiment Establishment, Ministry of Supply and Ministry of Transport. These include the replacement of the dipole aerials by slot aerials on the glide path which gives linearity to the glide slope virtually to ground level and improved monitoring and automatic control of the localizer beam to give indication of the centre of the runway on landing to ±20ft on a 2 000yd runway.

Open Days at the Electrical Engineering Department of the University of Birmingham were organized by Professor D. G. Tucker on September 18-19 to show the activities of the Department since its move to the new building some two years ago. Information on the Department's main undergraduate courses, together with its postgraduate courses on Information Engineering and Theory of Electrical Machines, took the form of displays and demonstrations of its teaching apparatus. Among the research projects being demonstrated was one being carried out in co-operation with the National Institute of Oceanography and consisting of an electronic sector-scanning asdic system for the location and identification of fish. Other projects included an experimental system for the halving of bandwidth of a television signal and the development of a special purpose digital computer using Dekatrons for arithmetic operations and crystal diodes for switching.

Mr. P. D. Canning of The Plessey Company Limited, has been appointed Chairman of a new sub-committee, 12-7, formed by the International Electrotechnical Commission to deal with Climatic and Durability Testing of Telecommunications Equipment. The new sub-committee has recently met under Mr. Canning at Ulm, in Western Germany, where he is also leading the U.K. delegation to similar sub-committees for electronic components, and acting as Secretary of sub-committee 40-5 on Basic Testing Procedures for Electronic Components. The International Electrotechnical Commission (also known as C.E.I., la Commission Electrotechnique Internationale), with headquarters in Geneva, is affiliated to the International Organization for Standardization as its electrical division. Its main aim is to facilitate co-ordination and unification of national electrotechnical standards.

Mr. T. E. Goldup, C.B.E., M.I.E.E., died at his home on Tuesday, 6 October. He started at the Mullard Company in 1923 from Signal School, Portsmouth, where he had been engaged on radio research work in the thermionic valve field for the Admiralty. The early days of his industrial career were spent in the Mullard valve factory at Balham where he was responsible for the manufacture of transmitting and receiving valves for the Government. While there he also assisted in the organization of the valve development laboratories. In 1928 Mr. Goldup was transferred to Head Office to set up a Technical Service Department to deal with valve application problems for Mullard customers, and in this capacity he became widely known throughout the Industry. In 1938 he was appointed a Director of Radio Transmission Equipment Ltd (now known as Mullard Equipment Ltd) and two years later he became a Director of the Mullard Radio Valve Co. Ltd. He joined the Board of Mullard Ltd, the parent Company, in 1951. Mr. Goldup has taken a prominent part in affairs outside his Company activities, and has been particularly concerned with all matters pertaining to technical education. He was Chairman of the Ministry of Supply College of Electronics at Malvern, and served on various committees of the Radio Industry Council, the British Standards Institution and other organizations connected with the electronic industry. He was awarded the C.B.E. in 1954, and was President of the Institution of Electrical Engineers for 1957-58.

A Radar Simulator Course for senior deck officers to a syllabus approved by the Ministry of Transport and Civil Aviation after full discussion with Shipowners' and Seafarers' Organizations and with radio manufacturers opened at the Sir John Cass College, London, early in October. A similar course is expected to be available in the near future at the South Shields Marine and Technical College and it is hoped that other Colleges will introduce courses during 1960. Officers completing the course will be issued by the College with a certificate and their certificates of competency may be suitably endorsed upon application to the Registrar-General of Shipping and Seamen at Cardiff either direct or through the Superintendent of a Mercantile Marine Office.

The British Standards Institution held their annual general meeting in London on 30 September and Mr. R. E. Huffam was re-elected for a second term of office as the Institution's president. Mr. Huffam was, until his recent retirement, United Kingdom Co-ordinating Director of Unilever Ltd. He also holds office as Chairman of B.S.I.'s General Council. The three deputy-presidents of the Institution were also re-elected. They are: Sir Roger Duncalfe, Sir Herbert Manzoni and Mr. John Ryan.

BOOK REVIEWS

Junction Transistor Electronics

By R. B. Hurley. 473 pp. 120 figs. Medium 8vo. John Wiley & Sons Inc., New York. Chapman & Hall Ltd, London. 1959. Price 100s.

IN the last two years the impetus of commercial application of junction transistors has increased enormously. More and more technicians are becoming involved in transistor electronics and the majority have had to acquire the necessary 'know how' without formal training. Unfortunately there has been no really suitable text book to ease the path of learning.

Most of the text book material offered was either by experts in semiconductor technology or workers in advanced circuit research. In the one case an involved description of transistor operation cloaked the circuit principle which the beginner required; in the other, the field covered tended to be rather limited.

Professor Hurley's text derives from his considerable lecturing experience in the subject. It is probably true to say that it is the first text on the subject by a practising educationalist—one who has met and studied the difficulties which the newcomer to the subject experiences. He has reduced the sections on device physics to a minimum qualitative description; and the material covers all the basic types of circuits in which junction transistors are being usefully applied.

One point which frequently puzzles the beginner is the multiplicity of equivalent circuits and attendant parameters which are in use. Professor Hurley has adopted the more readily recognizable and understood *T*-network for the simpler low-frequency small signal discussions. Later for non-linear, high frequency and broadband applications more complicated equivalents are derived. The treatment here is simplified yet still retains practical significance. However, the treatment of *h*-parameter networks is inadequate considering their importance to the circuit designer.

The first half of the book deals with low frequency linear applications. The development of circuits which give useful gain is allied with adequate qualitative description of the physical functioning of the device. The gain and impedance formulae are tabulated in exact and approximate form and appreciation of their use is encouraged by repeated application to later circuit problems. It is in this first half of the book that Professor Hurley's teaching experience is most evident. His is believed to be the first text in which an explanation of the negative collector voltage region of common-base characteristics has appeared.

The second half of the book begins with a consideration of high-frequency equivalent circuits. Gain and impedance formulae using *h*- and *T*-parameters are tabulated and *T-h* conversion formulae given. Further chapters deal with the prevention of self-oscillation in high-frequency amplifiers, the design of video frequency amplifiers, tuned amplifiers (including methods of gain control) and sinusoidal oscillators.

No transistor text would be complete without a treatment of switching applications. Professor Hurley gives a useful chapter in switching theory and follows with a discussion of non-regenerative and regenerative switching circuits. Finally he describes the transformer coupled multivibrator and its applications in computers, control systems, convertors and telemetry.

The book can be recommended to those familiar with thermionic valve techniques as an introduction to the more advanced textbooks. No problem sets are given but this is no handicap. Useful bibliographies are inserted at the end of each chapter to fill in the inevitable gaps and permit extension of knowledge in the direction of particular interest to the student. However, the majority are taken from American sources and may not be readily obtainable in Great Britain.

Generally, the style of the book is pleasing and interesting. Unfortunate phrases such as "properties that would overly complicate the circuit" are not so frequent as to annoy—but the reader should be warned.

P. ROBINSON.

Dynamische Vorgänge in Linearen Systemen der Nachrichten- und Regelungstechnik (Dynamic phenomena in linear systems of communication and control engineering).

By Hans Kaufmann. 208 pp. 105 figs. Medium 8vo. R. Oldenbourg, München. 1959. Price DM 26.50

THIS slender volume is a valuable contribution to the literature on the theory of communication and control engineering. It can be warmly recommended to advanced students and practical engineers with the necessary mathematical background and with a sufficient knowledge of technical German.

Throughout the book special emphasis is laid on the correspondence between the treatment on a time basis and on a frequency basis. This correspondence is well illustrated in the figures 2.6.1 and 2.6.2 on page 37. The reviewer was particularly pleased with the lucid explanation of the relative merits of the Fourier transformation and the Laplace transformation in dealing with the problems of linear networks.

As may be expected the treatment is highly mathematical throughout. As far as the necessary background is concerned a certain familiarity is expected with the elements of the theory of functions, with integral transforms, convolution integrals, Laguerre functions, filter theory and the like. The book serves the dual purpose of analysis and of synthesis or design of a network for a certain purpose. Clearly drawn diagrams and many numerical examples illustrate the text.

An introductory chapter deals with the origin and significance of the theory of linear systems, the connexion between excitation and response in electrical and mechanical translational and rotational systems, and the mathematical equivalents of treatment on a time and on a frequency basis. In the second chapter, weight function and frequency response are discussed as characteristics of the properties of the networks, the weight function relating the response to the excitation on a time basis while the frequency response is the corresponding relation on a frequency basis. The third chapter introduces the transfer function as a Laplace transform of the weight function. It is then treated as the quotient of a two power series which leads to its development on the lines of the theory of functions with its zeros and poles, and the significance of their location is discussed. In the fourth chapter the frequency response and, in particular, its logarithmic representation are dealt with. The fifth chapter is devoted to algebraic and geometrical criteria of stability. Approximation methods used for the analysis and synthesis of networks are discussed in the sixth chapter. Here Laguerre functions are introduced for calculating the transient function, and a brief section is devoted to the root locus method. The seventh chapter deals with characteristic values for transient phenomena, e.g. transit time, rise-time, bandwidth and error constants. In the eighth and final chapter sampled data systems and their application in systems containing digital computers are discussed.

Tables of the functions $\sin x/x$ and $\cos x/x$, of Laguerre functions, of Laplace transformations and of **L**- and **Z**-transforms are given in four appendices.

It may be regretted that in some instances the same symbol is used for different meanings, e.g. the symbol M for mutual inductance (p. 42), for torque (p. 44) and for a finite number characteristic for stability (p. 35), and the symbol s for the complex variable in Laplace transformation (passim) and for second (p. 187). The derivation of equation (5.4.1) on p. 107 might have been given in not more than four additional lines. But these are very minor blemishes in such a stimulating book containing a wealth of useful information.

A brief bibliography and a combined name and subject matter index conclude this well produced volume.

R. NEUMANN.

Vacuum-Tube and Semiconductor Electronics

By J. Millman. 644 pp. 300 figs. Medium 8vo. McGraw Hill Publishing Co. Ltd. 1958. Price 77s. 6d.

AT last a book has been written which brings together the internal physics and the external circuit characteristics of both thermionic valves and semiconductor devices. The resulting chapter by chapter comparison enables their similarities, wherever they exist, to be made as clear as their differences. Wherever possible the similarities of circuit design technique are emphasized; for example it is reassuring to find the technique of the load-line applied to the characteristics of both the triode valve and the transistor, and to be able to compare their equivalent circuits.

This is not a handbook for the sole use of design engineers, it is a text book for students of electronic engineering written by someone who is not afraid to explain the physics and mathematics to an engineer. The physics are extremely clear, while the mathematics need deter-no one who has reached 'A' level, or the ordinary National Certificate. Very good use is made in all the chapters of pictorial and graphical explanation.

The first two chapters deal with the principles and some applications of charged particles moving under the influence of magnetic or electric fields, while the third chapter discusses such particles moving in metals and semiconductors.

The next two chapters describe the physics and circuit characteristics of the vacuum diode and the semiconductor diode respectively. There are two chapters devoted to vacuum triode characteristics and equivalent circuits, while two chapters immediately following are devoted similarly to the transistor, including a comparison of the performance of the three possible circuit connexions, in terms of the physical constants. The hybrid perameters are also defined and briefly discussed.

A chapter on untuned voltage amplifiers touches all the many aspects of cascading several l.f. stages, and is followed by a very thorough chapter on the audio power stage; in both these chapters the valve is dealt with in more detail than the transistor.

The broad scope of this book is best emphasized by mentioning that eight other chapters include photoelectric devices, electrical discharges in gases with commercial examples, rectifiers, power supplies, feedback amplifiers and sinusoidal oscillators.

There is one chapter in which the characteristics of multi-electrode valves and h.f. transistors are discussed, but the book does not include any radio-frequency electronics. Throughout the book, worked numerical examples demonstrate the text, and references covering more than twenty years appear at the end of each chapter.

There are nine Appendices including physical constants, conversion factors, and the mks System of units.

The clarity and consistent treatment of the subject through all the chapters indicates the objective, successfully achieved, in the mind of the author, to provide an un-muddled basis for any electronic engineer, so that he can afterwards study more detailed design techniques restricted to his particular part of the subject.

W. Ian Heath.

Topics in Electromagnetic Theory

By D. A. Watkins. 118 pp. 84 figs. Medium 8vo. John Wiley & Sons Inc., New York. Chapman & Hall Ltd, London. 1958. Price 120s.

THE author of this slender volume is professor of electrical engineering and director of the Electron Devices Laboratory at Stanford University in the United States. He is also the inventor of the low-noise travelling-wave tube and the helitron oscillator. The book has been developed from the lecture material of a short course for selected electrical engineering graduates whose previous work in the subject has been based on the text book *Fields and Waves in Modern Radio* by Ramo and Whinnery.

The first of the book's four chapters deals with the physical and mathematical aspects of periodic transmission systems of the kind employed in linear accelerators, travelling-wave amplifiers, and backward-wave oscillators. With Floquet's theorem as the basis, the subject is developed with the aid of the ω-β diagram. Both the field and the equivalent network methods are employed. Helical structures are discussed in the second chapter. The concepts of sheath helix and tape helix are introduced and lead to a study of multi-wire helices, helical aerials, and helical delay lines. The third chapter is devoted to Pierce's coupled-mode theory which is applied to a variety of microwave problems. The final chapter deals with anisotropic media. Among the topics discussed are Polder's tensor permeability for a saturated ferrite, wave propagation in an infinite medium, the principles of ferrite devices, Faraday rotation and field displacement devices. Associated with each chapter is a set of problems.

The book is lucidly written and much of its contents has not previously been available in text-book form. It will be of considerable interest and value to workers in the field of microwave engineering.

S. R. Deards.

Patents for Engineers

By L. H. A. Carr and J. C. Wood. 98 pp. Demy 8vo. 1st Edition. Chapman & Hall Ltd. 1959. Price 18s.

THIS is a short book dealing with the law in simple form, avoiding wherever possible technical legal phraseology, which also deals especially with the engineering and commercial aspects. The authors hope that as well as preventing the engineer from taking any step which might put him in a difficult position later, it will also tend to make any discussion with a legal adviser more profitable, since the engineer will be more likely to know in advance the information required from him.

High Speed Computing Methods and Applications

By S. H. Hollingdale. 239 pp. 71 figs. Demy 8vo. The English Universities Press Ltd. 1959. Price 25s.

THIS is one of the very few books written for those who are not yet acquainted with the art of computers which does not merely recite a list of wonders which may be performed. Its merit lies in really getting down to brass tacks of how work can be performed.

The type of material presented has already been published all too often. What is different is that many other authors have assumed that the non-professional reader wishes to be amazed or humbled but never on any account instructed or made to think.

The book does, however, have its shortcomings which may be considered by many to be minor. Both the machines reviewed in detail have now been superseded by other more efficient machines but it does of course follow that experience of these machines is somewhat less. One can say that there is every reason for including one of the forerunners namely EDSAC and the author is of course quite right to discuss the machine with which he is most familiar. Unfortunately the coincidence does not allow for the introduction of a more modern machine with the elaborate array of organizational and 'red tape' instructions and facilities which are now commonplace.

Another defect of the work is that

although its title promises equality of treatment, it is in fact biased towards methods rather than applications. It is perhaps greedy to expect full treatment of both aspects but it would be less confusing for publishers to adjust titles to fit contents.

The real value of this book which outweighs all criticism lies in its inclusion of lucid descriptions of physical, logical and programming design while remaining compact and very readable.

H. DENYER.

Proceedings of the 1958 Electronic Components Conference

222 pp. 132 figs. Medium 8vo. Engineering Publishers, New York. Interscience Publishers, London. 1959. Price $6.00.

EACH year a conference is held in the United States of America on some particular application of components. In 1958 the theme was "Reliable Applications of Component Parts" and the complete series of papers presented at this Symposium has now been published.

There are some twenty-five papers covering developments in resistors, capacitors and dielectrics, transistors and solid state devices, valves, economic aspects of components, reliability, applications of components and progress with materials. The papers vary somewhat in their quality. Some of the better ones present an excellent review of the 'state of the art' in 1958—others are either highly theoretical or cover developments which are now obsolescent. While it is not possible to give a complete list of the papers, the more interesting ones are given below in order that the reader may have an idea of the subjects covered:

- Elimination of Mica Films as Dielectric Material for High Temperature Capacitors
- High Temperature Precision Wire-wound Resistors
- Reliability Data on Solid Tantalum Capacitors
- Design Limitations of Semiconductor Components
- Predicted Reliability of a Transistorized Amplifying System
- Revolutionary Electronic Equipment Design Concept
- Component Application Stress Analysis
- Capability of Commercially available Coaxial Cables when exposed to Extreme Environments
- Transient Coil Current as a Means of Relay Evaluation
- Component Reliability Research in the United Kingdom
- Status Report on the Ceramic Receiving Tube Development
- Glass for High-Temperature Components
- Evaporated Electronic Components and Microcircuitry
- Cine-Radiography—A Unique Approach to the Development of Reliable Components
- Multiconductor Flat Wire Cable

There is little doubt that the book is of value to electronic equipment design engineers and component specialists but it should be pointed out that there has been another Component Symposium (held on 5 to 7 May, 1959) and the Proceedings of this conference have also been published.

G. W. A. DUMMER.

Elektrokeramik (Electroceramics)

By A. Hecht. 264 pp. 200 figs. Medium 8vo. Springer-Verlag, Berlin. 1959. Price DM 41.40.

THIS book deals with the basic materials, the production, the testing and the applications of ceramics for electrotechnical purposes. It intends to be a mediator between the ceramic industry and the manufacturer of electrical equipment in that it teaches the ceramic specialist the requirements of the electrical industry and the electrical designer the properties of ceramic materials and the limitation of their use. Reference is frequently made throughout the book to the German Standard Specification DIN 40 685 "Ceramic insulation materials for electrical engineering", and a copy of this specification is contained in a pocket at the end of the book.

A brief introductory chapter deals with the significance of the electrical and the ceramic industries. The second chapter is devoted to the basic materials and the technology of ceramic materials, in particular to the shaping, drying, glazing and burning, the machining and the connecting of ceramic bodies with metal and glass. The brief third chapter discusses the shrinking of, and the tolerances for, ceramic products. In the fourth chapter the electrical, mechanical, thermal and other physical and chemical properties of ceramic materials are dealt with at some length. Testing methods are described in the fifth chapter. The sixth chapter deals in greater detail with the shaping and the construction of ceramic building elements as used for h.v. insulators, l.v. insulators, installation materials and heating elements (a special feature here being a comparative table of some 12 pages showing on opposite columns correct and incorrect designs), and for capacitors. The final chapter gives a survey on the various applications of ceramic materials in electrical engineering including high frequency installations.

An extensive bibliography and a subject matter index conclude this well produced volume which contains a wealth of useful information and may be recommended to all those interested in this field.

R. NEUMANN.

Solid State Magnetic and Dielectric Devices

Edited by H. W. Katz. 536 pp. 374 figs. Medium 8vo. John Wiley & Sons Inc., New York. Chapman & Hall Ltd, London. 1959. Price 108s.

UNLIKE many books which represent the joint efforts of several authors, this volume is well-written throughout and has clearly been carefully edited with regard to detail. The authors expressly exclude treatment of devices whose action depends primarily on the control of charge flow, as is the case with transistors and similar semiconductor-based circuit elements. The subject matter may be described as 'the theory of solid state circuit elements which utilize the magnetization and polarization properties of materials'. Particular reference is made to the ferrites and titanates. The properties of the more important of these and other pertinent materials are tabulated at the end of the book.

It is stated in the preface that "the compilation of a textbook in this embryo field, encompassing many diverse areas in theory and applications is difficult". This is certainly true. The task has been accomplished with a large measure of success. The text could however, have been made somewhat less disjointed by a more suitable juxtapositioning of chapters. For example, chapters 3 and 5 which deal respectively with "Electrostrictive and Magnetostrictive Systems" and "Electrochemical Applications" clearly form an independent section of the book. They are in fact separated by a chapter on "Nonlinear Magnetic and Dielectric Materials". The first two chapters review 'Electrostatic and Magnetostatic Field Theory" and the "Origin of Magnetic and Dielectric Properties" Apart from chapter 11, in which some less well-known "Magnet and Dielectric Measurements" are discussed, the remainder of the book deals with the theory underlying the operation and construction of devices. Practical details of design are not discussed.

In chapter 5, which has already been referred to, the main topics are ceramic high-voltage transformers and ceramic filter elements. Chapter 6—"Small-Signal Applications" is devoted mainly to ferrite applications below "microwave frequencies" while chapter 7 is specifically concerned with "Ferrites at Microwave Frequencies". The remaining chapters deal with "Magnetic and Dielectric Amplifiers", "Digital Techniques Employing Square Loop Materials" and "Magnetic Recording". There are references to relevant material at the end of each chapter.

The book certainly fills a gap in the literature of the solid state. It will be very useful to the reader who wishes to gain insight into the underlying theory of the various devices discussed.

F. J. HYDE.

Semiconductor Abstracts Volume 4

Compiled by Battelle Memorial Institute. Edited by E. Paskell. 456 pp. Demy 4to. John Wiley & Sons Inc., New York, Chapman & Hall Ltd, London. 1959. Price 96s.

THE effort which has been put into the preparation, editing and review of these abstracts, as in the three previous issues, has all been voluntary. As with the previous issues, this year's compilation of abstracts has been prepared jointly by members of the Solid State Devices and the Physical Chemistry Division of Battelle Memorial Institute. The ex-expanded coverage initiated with the 1955 issue has been continued this year by including papers presented at meetings of the American Physical Society and the Electrochemical Society.

LETTERS TO THE EDITOR

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

A Power Supply Unit for Improving Small Electric Motors

DEAR SIR,—The running characteristic of a small electric motor, with permanent magnet or separately excited field, can be represented by a diagram such

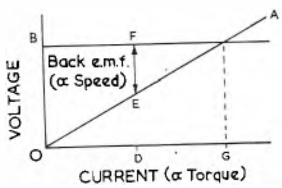


Fig. 1. Typical running characteristic of small motor

as Fig. 1. The line OA represents the voltage dropped across the armature, brush and external circuit resistance for any value of load torque. If the motor is fed from a constant-voltage supply OB, then at a value of load torque D, the vertical height FE represents the back e.m.f., and the speed will be proportional to this. The diagram therefore shows how the motor speed will vary with load, and in particular that it will stall at torque G.

Suppose now the motor be fed from a supply whose voltage BHC in Fig. 2

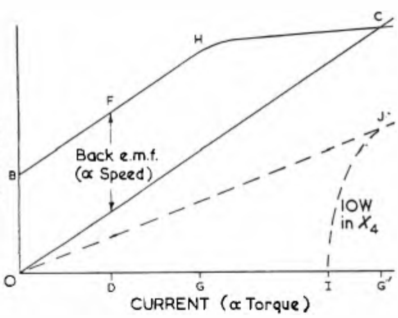


Fig. 2. Motor supply voltage against current

rises linearly—over a range—with output current. Then if the slope of BH is nearly as great as the slope of OA, for moderate values of load torque the speed will be practically independent of load. The motor will maintain its no-load speed up to a torque where previously, on a constant voltage supply, it stalled. It behaves as if it were much larger physically than it actually is. Looked at another way, the motor is supplied from a constant voltage supply in series with a negative resistance. The negative resistance is arranged to cancel the positive resistance of the armature, brushes and circuit wiring.

A supply unit for small electric motors to make use of this effect has been built (Fig. 3). S_1 and RV_1 set the no-load output voltage in the range 3 to 13V, and may be labelled coarse and fine

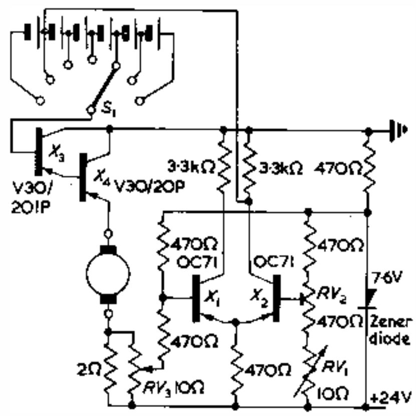


Fig. 3. Power supply unit for small motor

speed. RV_3 determines the negative resistance, continuously variable in the range 0 to -50Ω. RV_2 is preset so that X_2 is just not bottomed with the motor running light. The available collector swing at X_2 limits the 'boost' voltage (BH in Fig. 2) to 8V. The maximum power output of the device is therefore limited by the maximum full-load output voltage (11V at low speed, 21V at high speed); also by power dissipation in X_4. The permitted 10 watts here is represented by the curve IJ in Fig. 2. The straight line OJ, drawn tangential to IJ, represents the lowest value of armature resistance which may be stalled without exceeding this figure, about 15Ω. Motors having armatures of resistance higher than this may be stalled with impunity. Armatures of lower resistance should be loaded with circumspection.

Yours sincerely,

P. E. K. DONALDSON and
J. G. ROBSON,
Physiological Laboratory,
Cambridge.

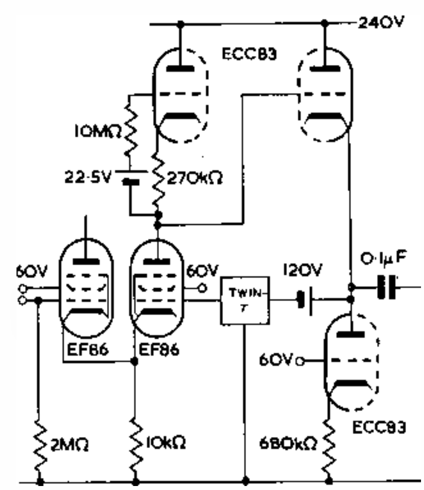
A Narrow Band Amplifier

DEAR SIR,—It is frequently necessary to phase detect small periodic signals in the presence of a large amount of noise, possibly several orders of magnitude larger than the signal. In order to prevent the pre-detector amplifier from being driven far off its working point it is then necessary to use a narrow band amplifier tuned to the wanted signal.

It is well known that a convenient way of achieving the desired selective amplification is to use an amplifier with a feedback filter network having the inverse characterstic, that is, zero transmission at the signal frequency. The reciprocal relationship between the feedback factor and the amplifier gain applies under ideal conditions in which the feedback network sees an infinite impedance and is fed from a zero impedance source. In the case of a normal audio frequency *RC* coupled amplifier, the latter condition is difficult to fulfil, with the result that low and high frequencies suffer attenuation apart from that due to the filter. This has the undesirable effect of increasing the bandwidth, given by $\int_0^\infty |G/A|^2 df$, often by a large factor. Here G and A are the gain with and without feedback respectively.

In the present case these difficulties are overcome by applying well-known d.c. amplifier techniques to the design of an a.c. selective feedback amplifier. A high gain single stage d.c. coupled feedback amplifier with cathode-follower output is supplied with a twin-T type rejection network (zero transmission 300c/s). This network feeds into the grid of the pentode amplifying stage and is itself fed from the cathode-follower output. The full feedback factor of the rejection network is therefore available at all frequencies up to those attenuated by the valve electrode capacitances. The amplifier stage is cathode fed from the input cathode-follower. Advantage is taken of the high μ of a pentode at low currents to obtain a large single-stage gain. The anode resistance of the EF86 pentode at 0·1mA and anode voltage, V_a = 120V is of the order of 20MΩ and the corresponding g_m = 0·26mA/V. The effective anode load impedance of the device shown is R_a =

Fig. 1. D.C. coupled narrow band amplifier



$r_a + (\mu + 1) R_k$ where μ is the triode gain and r_a its anode resistance. The ECC83 used under these conditions has a μ = 90 so that R_a is of the order of 24MΩ. Thus the single stage gain of the amplifier is 2 800 and the Q = 700. This gain will not be achieved in practice unless an output stage cathode-follower of input impedance higher than 10MΩ is used, since otherwise the high load impedance of the amplifier stage will be shunted. This may be achieved either by under-running the filament of the output cathode-follower or by using a constant current device as cathode load. The circuit diagram is shown in Fig. 1 and the amplitude-frequency response is shown in Fig. 2. Tolerances in the twin-T com-

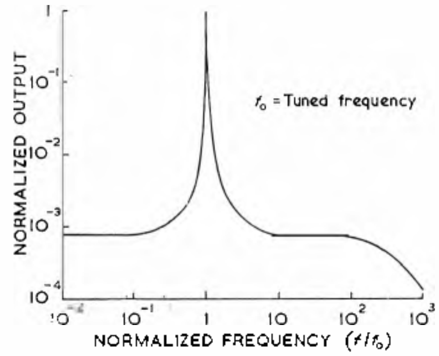


Fig. 2. Amplitude frequency response of amplifier

ponents reduced maximum amplification to 1 400, which was adequate for the present purpose. At 300kc/s the output is further attenuated as no precautions were taken to minimize the effect of stray and interelectrode capacitances.

Hearing aid batteries were used for the 22·5V and 120V floating potentials and the low working current of 0·1mA for all valves made possible the use of h.t. batteries as power supplies.

Because of the direct coupling the amplifier is particularly suitable for use at very low frequencies (e.g. 10c/s). Similar amplifiers have been in use on various experiments in these laboratories including crossed beam methods of measuring electron collision cross-sections[1] and electron energy distribution measurements in gas discharges[2].

Yours faithfully,

R. L. F. BOYD and J. B. THOMPSON,
Department of Physics,
University College, London.

REFERENCES

1. BOYD, R. L. F., GREEN, G. W. Electron Ionization Cross-Section using Crossed Beams. *Proc. Phys. Soc. 71*, 351 (1958).
2. THOMPSON, J. B. Negative Ions in the Control of Oxygen Discharge. *Proc. Phys. Soc. 73*, 818 (1959).

Analysis of Impact Noise

DEAR SIR,—We have read with interest the article which appears on pages 200-203 of your April 1959 issue of ELECTRONIC ENGINEERING.

This article appears to be based on previous work published by Dr. Arnold P. G. Peterson in this country and appearing both in *Noise Control* and in the *General Radio Experimenter*.

Under the circumstances it would have been appropriate for you to include the following references among those listed in the article in order that proper credit be given to earlier work in the field:

1. PETERSON, A. The Measurement of Impact Noise. *Noise Control*, pp. 46-51 (March, 1956).
2. PETERSON, A. The Measurement of Impact Noise. *General Radio Experimenter*, pp. 1-7 (February, 1956).

Yours faithfully,

C. E. WORTHEN,
General Radio Company,
Cambridge, Massachusetts, U.S.A.

The Author Replies:

DEAR SIR,—With regard to my article which appeared in the April 1959 issue of ELECTRONIC ENGINEERING, I very much regret that the valuable references given in Mr. Worthen's letter were not mentioned, as they certainly do describe earlier work by Dr. Peterson in this field.

Yours faithfully,

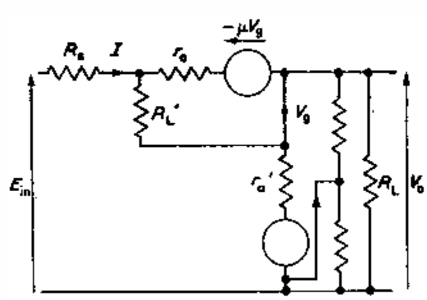
F. M. SAVAGE,
Dawe Instruments Ltd,
Ealing.

Regulated Power Supplies

DEAR SIR,—In their analysis of the closed loop regulator, Fig. 5, I think Mr. Collins and Mr. Smith failed to show the disadvantage of that particular circuit connexion. I refer to the return of the pentode anode load to an unregulated voltage which may considerably reduce the performance of the circuit. While the performance may be regained by suitable 'forward feedback' to the pentode screen I think it important to stress the fact that an unregulated supply for the pentode anode circuit is not the best arrangement. In the analysis given below, it is shown that the stabilization and output impedance may differ widely from the values given by the authors' equations. The main difference in these equations is due to the definition of $-M$ which I have defined as the voltage gain from the pentode grid to anode with no variation in E_{in} or in current through R_s. In addition to the authors' assumptions I have assumed that R_s is negligible compared with the pentode anode load..

The equivalent circuit of the regulator is shown in Fig. 1.

Fig. 1. Equivalent circuit of regulator



Let $P = \dfrac{r_a'}{r_a' + R_L'}$

$$V_g = -M\beta\Delta V_o - P(\Delta I R_s - \Delta E_{in}) - \Delta V_o$$
$$= -M\beta\Delta V_o - \Delta_o P\, R_s/R_L + P\Delta E_{in} - \Delta V_o \quad \text{.......... (1)}$$

and $\Delta E_{in} - \Delta I(R_s + r_a) + \mu V_g - \Delta I R_L = 0$

$$\Delta E_{in} - \Delta V_o \frac{R_s + r_a}{R_L} + \mu V_g - \Delta V_o = 0$$

and from equation (1)

$$E_{in} - \Delta V_o \frac{R_s + r_a}{R_L} - \mu M\beta\Delta V_o -$$
$$\mu P\, R_s/R_L\, \Delta V_o + \mu P\Delta E_{in} \quad \mu\Delta V_o - \Delta V_o = 0$$
$$\Delta E_{in}(1 + \mu P) = \Delta V_o \quad \frac{R_s + r_a}{R_L} \quad + \mu M\beta + \mu$$
$$R_s/R_L\, P + \mu + 1]$$
$$\frac{\Delta E_{in}}{\Delta V_o} = \frac{\frac{R_s + r_a}{R_L} + 1 + \mu(1 + M\beta + P\, R_s/R_L)}{1 + \mu P}$$

Since $P\, R_s/R_L$ will normally be small, the expression for stabilization is reduced by approximately a factor of $1 + \mu P$ due to voltage variation at the 'supply' end of R_L.

For output impedance, $E_{in} = 0$ and the current I will flow in the opposite direction.

$\therefore\ \Delta V_o = \Delta I(R_s + r_a) + \mu V_g$ .... (2)

and $V_g = -M\beta\, V_o + P\, I R_s - V_o$ .... (3)

from equations (2) and (3)

$$\Delta V_o = \Delta I(R_s + r_a) - \mu M\beta\Delta V_o + \mu P R_s \Delta I$$
$$- \mu\Delta V_o$$
$$\therefore\ \Delta V_o\,(1 + \mu M\beta + \mu) = \Delta I(r_a + R_s)(1 + \mu P)$$
$$\therefore\ R_o = \frac{r_a + R_s(1 + \mu P)}{1 + \mu + \mu M\beta}$$

The effect on R_o is therefore to increase the effective source impedance by $(1 + \mu P)$.

As the factor $(1 + \mu P)$ may be say, 2 to 10, I think it earns a place in the analysis.

Yours faithfully,

R. THOMPSON,
Cheltenham.

The Authors Reply:

DEAR SIR,—We are indebted to Mr. Thompson for his comments. We must admit to having failed to emphasize this particular disadvantage, but would point out that we had kept our analysis to simple conditions by making a number of overall assumptions. It is interesting to note that Mr. Thompson has made an exception of one particular assumption, although he has retained several others, and we feel that if, in the limit, no overall assumptions were made, the resulting analysis would be somewhat different from both previous analyses. In fact, an individual analysis would have to be made for each particular design.

We agree with a range for factor P of 2 to 10, but feel that the majority of designs will centre at $P \simeq 4$, and stress that this could be considered as another variable (in the same category as valve parameter variation, change of loop gain with change of dynamic conditions etc.) which is catered for by an extra requirement of loop gain.

The effect is of more consequence from the point of stabilization than from output resistance, since the factor P only applies to one term of the output resistance expression. For the higher frequencies, the effect does not exist if a filter of the form R_2C_2 in our Fig. 7 is included.

In conclusion, we note that, though the effect and its cause are relatively obvious, not so the cure. We feel that, in general, it is preferable to design for increased loop gain than to attempt to stabilize the return point for the anode load.

Yours faithfully

D. J. COLLINS and J. E. SMITH,
Solartron Research
and Development Ltd,
Dorking.

Rectangular Hysteresis-Loop Magnetic Cores

DEAR SIR,—I was very interested in the article by Mr. Kaposi (May 1959) concerning the rectangular hystersis loop cores, and the concept of inductance used in obtaining performance data over both the reversible and irreversible portions of the loop.

An additional requirement to those stated in the article for validity of the equation,

$$\tau = \frac{S_w}{H_m - H_0}$$

is that H_m must be equal or greater than[1] $2H_0$. If as Mr. Kaposi suggested, H_m be made very much greater than H_0

$$\tau = S_w / H_m$$

From this an interesting result may be obtained. The back e.m.f. generated during switching is given by

$$V = N\phi_t / \tau \times 10^{-8} \text{ volts}$$

where N number of turns ϕ_t total flux switched giving

$$V = \frac{N \; \phi_t \; H_m \; . \; 10^{-8}}{S_w}$$

volts

or for a given core $V = K_1 \; H_m = K_2 \; I$ (I is the applied current).

The result of this is that during the switching period, under constant current conditions, and providing $H_m \geqslant 2H_0$, the core generates an e.m.f. proportional to the magnitude of the applied current and not its derivative. The concept of inductance whereby voltage is related to rate of change of current is thereby violated.

The equivalent circuit of the resistance with a back e.m.f. is related to this effect.

Yours faithfully,

K. E. WOOD,
Baltimore, U.S.A.

REFERENCE

1. BROWN, D. R., BUCK, D. A., MENYUK, N. A Comparison of Metals and Ferrites for High-Speed Pulse Operation. *Trans. Amer. Inst. Elect. Engrs.* 1, 73, 631 (1955).

The Author replies:

DEAR SIR,—The voltage generated by the switching of the core is indeed proportional to the applied field

$(H_m - H_0) = S_v \; (\mu V_1)$

where S_v is a constant for the material

μV_1 is the peak output voltage

if $H_{w(max)} < H_m < H$ $(t_r/\tau < 0{\cdot}3)$, where

$H_{w(max)}$ critical field required to move all the mobile domain walls

t_r = rise-time of square wave input pulse

τ = switching time.

This was shown by Messrs. Menyuk and Goodenough (reference 1 and 2 in my article).

This result follows from a physical model which is to explain the switching of a core, i.e. to explain that a finite output voltage can exist in spite of the fact that the rate of change of the switching current $di/dt = 0$.

Thus the concept of inductance is not violated if one considers that the inductance is not constant during switching.

Yours faithfully,

JANOS KAPOSI,
Ericsson Telephones Limited
Beeston, Nottingham.

A Transistor Characteristic Curve Tracer

DEAR SIR,—With reference to the article by Mr. J. F. Young in the June 1959 issue it is doubtful whether the use of Zener diodes is necessary to obtain a flat topped stepped waveform and to avoid the slight pairing effect due to difference in the counting tube discharge paths as referred to in the article.

It was found, during work[1] on this same method of displaying transistor and valve characteristics in 1954, that the particular type of cold-cathode counting tube and circuit used produced a satisfactory flat topped stepped waveform and, although the cathodes were paired as in the article, no splitting of the lines or pairing effect was noted. However, the hysteresis effect was noted and this appeared to be some function of display frequency since the effect was more pronounced at frequencies higher than the 50c/s normally used. A delay in the collector sweep voltage, before rectification, was incorporated to ensure that the current step had reached a steady value; this avoided sweeping the collector voltage during the rise-time (approximately 2msec) of the stepped current waveform.

Yours faithfully,

E. G. BOROUGHS,
Sperry Gyroscope Co. Ltd,
Bracknell, Berks.

REFERENCE

1. Patent Specification, 809, 375.

The Author replies:

DEAR SIR, As mentioned in the article, the pairing effect is usually small, and, with many tubes, not troublesome unless a large oscilloscope amplification is used. The effect with a particularly bad tube is shown in the use of potentiometers in the cathode circuits, though this is a complex solution. Replacing the cathode resistors with Zener diodes as stated would eliminate the effect and have the additional advantage that the control voltage for the constant current generator would be little affected by supply voltage variations.

As stated in the article, the hysteresis effect seen on the photograph (Fig. A)

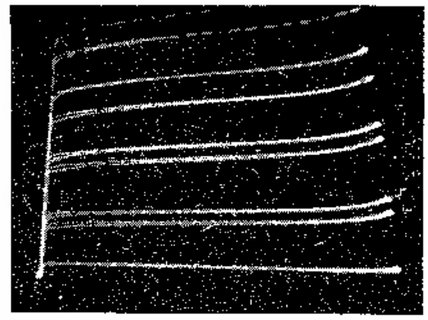


Fig. A. Hysteresis effect

has been noted by other workers in the field, and this is normally small.

In the instrument described, the alternating drive for the Dekatron pulse generator grid was phase advanced as shown in Figs. 8 and 9 of the article, since the phase shift circuit here works into a fixed load. The alternative mentioned by Mr. Boroughs of using a delay in the collector voltage circuit is less convenient, since the load there depends on the settings of the switches.

Yours faithfully,

J. F. YOUNG,
Great Barr,
Birmingham 22a.

The Electroforming of Waveguide Components

DEAR SIR,—Further to the article by Mr. P. Andrews (March 1959, p. 150) your readers may be interested in the following additional details on electroforming waveguide components which I have found to be essential if uniform results are to be obtained.

It is well known in the literature of electroplating that the addition of minute quantities of animal matter such as gelatin improves the surface finish and permits faster plating through the use of higher currents. Usually a few parts per million by weight of gelatin is sufficient and this can be determined by experiment.

Slow rotation of the mandrel will serve to agitate the solution and reduce edge effect by the formation of a virtual cathode as well as permitting the use of higher currents.

It is conventional to use stainless steel mandrels; however, molybdenum offers several advantages. The coefficient of linear expansion for molybdenum is about $5 \times 10^{-6}/°C$, which is about one quarter that of either copper or stainless steel. Heat treatment materially aids in the removal of the mandrels when molybdenum is employed, whereas little advantage is derived with stainless steel. Moreover, it is easier to obtain surfaces with irregularities that are less than 5μin with molybdenum than with stainless steel. Both these factors reduce the force required to remove the mandrel from the finished component thus minimizing the possibility of distortion.

Yours faithfully,

H. A. BUCKMASTER,
University of Alberta,
Edmonton, Alberta.

THE AIRCRAFT EXHIBITION

A description, compiled from information supplied by the manufacturers, of a few of the electronic exhibits shown at the recent exhibition of the Society of British Aircraft Constructors.

(Voir page 631 pour le résumé en français: Zusammenfassung in deutscher Sprache Seite 638)

BLACKBURN ELECTRONICS LTD

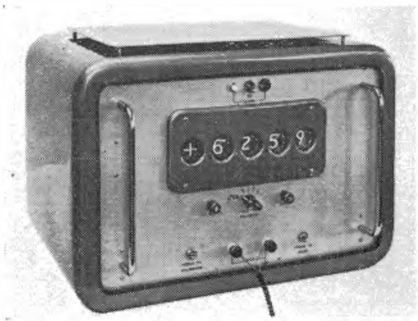
Brough, Yorkshire.

THREE AND FOUR DIGIT VOLTMETERS

(Illustrated below)

These electronic digital voltmeters have been designed for applications in fields requiring accurate measurements of d.c. voltages. These include computer inputs and outputs, reference supply measurements, strain gauge measurements, production testing of electronic and electrical components and data handling and display (i.e. analogue to digital conversion).

The circuit design of these instruments gives a high factor of reliability and accuracies of ± 0·1 per cent model B.I.E. 2113 and ± 0·5 per cent model B.I.E.



2114, on any of the three ranges of the voltmeter. These ranges are 0 to ± 9·60, ± 9·60 to ± 96·0, ± 96·0 ± 999V d.c. on B.I.E. 2113, and 0 to ±9 600, ±9·600 to ± 96·00, and ± 96·00 to ± 999·9V d.c. on B.I.E. 2114. The input impedance of both instruments is 50MΩ in the lower range and 5MΩ in the upper ranges. Range changing and sign selection are automatic and both are indicated on the visual display which employs neon number tubes. The readout times of the instruments are 13msec for the three digit unit B.I.E. 2113, and 103msec for the four digit unit B.I.E. 2114.

In operation the instruments compare the input voltage with a ramp which is linear within ± 0·01 per cent. Between the time the ramp voltage passes zero volts and the amplitude of the input voltage, the number of cycles generated by crystal controlled oscillator are counted. This count is the numerical value of the voltage and is displayed directly on the neon number tubes. Calibration referred to a standard Weston cell is built-in in both instruments. 80V negative going pulses are available to drive an in-line printer, electric typewriter, paper punch or magnetic tape recorder or remote display units.

High speed scanning switches and remote display units incorporating storage facilities are available. These units enable large quantities of data to be displayed continuously, in digital presentation at any point remote from the voltmeter.

Both instruments have built-in stabilized power supplies and can be supplied either for rack mounting or in an instrument case.

EE 14 751 *for further details*

DATA MONITORING AND LOGGING EQUIPMENT

This equipment has been designed for use with continuous process plant, to monitor automatically, information regarding plant performance. It will detect faulty conditions should they occur, or log the data for research and development purposes.

The input signals which are derived from thermocouples, pressure transducers, flowmeters etc, are scanned repetitively. The voltage representing each input parameter, after amplification, is compared with a preset upper and lower limit. In the event of any parameter departing outside the pre-set limit, the condition is detected and aural and visual warning given. The faulty parameter is digitized with three decimal digit precision (an accuracy of ±0·1 per cent full scale) and recorded by an inline printer or tape punch which also indicates the identification of the parameter and the time of the faults.

The equipment can also log all the data at pre-determined time intervals, on demand, or continuously.

EE 14 752 *for further details*

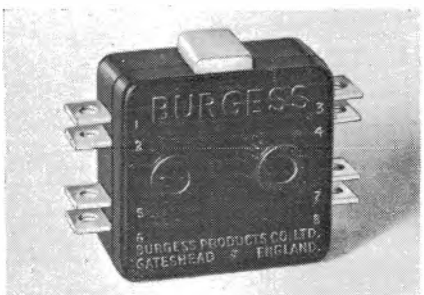
BURGESS PRODUCTS CO. LTD

Dukes Way, Team Valley, Gateshead, 11.

MICROSWITCHES

(Illustrated below)

The M1 is a new four circuit, bistable sub-miniature microswitch. It is a phenolic cased model with a nylon operating button and eight solder terminals of pure silver. Over operating button and terminals, its approximate dimensions are 1·0in wide, 0·9in high and 0·39in deep. Operating force is 20oz maximum, movement differential is 0·020in maximum and the mechanical life rating is 10 million operations minimum.



The main purpose of the M1 switch is to provide, in a very limited space, instantaneous double pole changeover switching facilities. This it does with the attendant advantages of snap-action, repeat accuracy, small movements and long life. It will make or break or change-over two completely separate circuits in the same instant of time and can be used for all precision double pole requirements, even the control of three-phase equipment by the switching of two phases.

Another feature of the M1 is its ability to control direct currents of considerably higher value than those associated normally with sub-miniature precision switches. Suitably connected to the load or loads, it can be used either as a two-gap switch with a d.c. rating of 5A at 30V, or as a four-gap single pole switch with a d.c. rating of 1A at 250V. These ratings are achieved without any protective circuits. The general a.c. rating is 5A at 250V.

EE 14 753 *for further details*

EKCO ELECTRONICS LTD

Southend-on-Sea, Essex.

AIRBORNE WEATHER RADAR

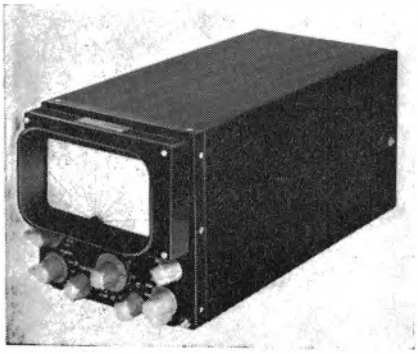
(Illustrated above right)

Ekco Electronics exhibited a new transistorized airborne weather radar with a weight of only 56 lb which is less than half that of previous systems of similar performance. Only three main units are employed—indicator, transmitter/receiver and scanner—thus allowing a considerable saving in installation costs with a further saving in cable and racking weight. Space requirements and power consumption are also greatly reduced.

This new radar provides all the facilities of the best weather radar systems, including a fully stabilized 24in antenna, 150 miles range and a high-brightness indicator display. The reduction in weight has been achieved by the extensive use of transistors and re-arrangement of the system layout to eliminate the conventional synchronizer and control units.

The indicator conforms to the A.R.I.N.C. standard dimensions and includes all operational controls. A 5in × 3in rectangular tube, operating at 18kV, displays the full range throughout the 180° azimuth sweep. The high operating voltage together with an angled polaroid filter, ensure exceptional display visibility even under bright ambient light conditions. Either single or dual-indicator installations can be provided.

The transmitter/receiver employs the



latest microwave receiver techniques, including a ferrite load isolator. It is an X-band unit rated at 15kW and is housed in a short-half A.T.R. case, conforming to A.R.I.N.C. standards.

The standard scanner uses a 24in antenna with switchable pencil/cosec^2 beam distribution. Line-of-sight stabilization is used with a velocity-corrected servo system which maintains stabilization accuracy throughout the 180° sector scan.

EE 14 754 *for further details*

FERRANTI LTD

Hollinwood, Manchester.

Airborne Radar and Fire Control Equipment

Ferranti Airborne Radar and Fire Control Equipment, known as AIRPASS, was shown publicly at Farnborough for the first time.

AIRPASS stands for Airborne Interception Radar and Pilot's Attack Sight System. The two basic components, the interception radar and the attack sight, have been developed as a complete equipment to enable a fighter pilot to find and destroy his target, even though it may be invisible to him and travelling at supersonic speed.

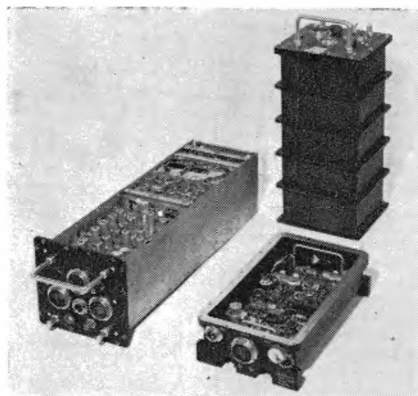
The radar is a powerful search-and-track set, and includes a computer for working out the correct course for the fighter to follow in order to reach a suitable position for the final attack. This information is presented to the pilot in his sight which also ensures that the weapons carried by his fighter are aimed correctly.

During the search phase, the radar scans a sector of sky covering a wide angle horizontally and vertically on either side of the flight-path, and extending many miles ahead of the fighter. When the pilot detects a target in this sector, he can make the radar beam 'lock-on' to the target. The radar will then track it automatically and the approach computer will start to operate.

Once the radar has started to track the target, all the information required to complete the interception is presented to the pilot, who can carry out 'blind' attacks using this equipment. Provision is also made for visual attack should that be desirable. As a precaution, a warning light is provided to tell the pilot to break-off the attack, should he approach close enough to be in danger of colliding with the target.

In the English Electric Lightning, the radar is housed as a single unit inside the centre-body of the engine air-intake. Although this raised some mechanical design problems in the early stages, it has ensured that the radar is a very compact unit. The single-package layout has many other advantages, not the least of which is the saving of weight, by the elimination of the usual number of separate boxes with their interconnecting plugs and cables.

EE 14 755 *for further details*



THE GENERAL ELECTRIC CO. LTD

Magnet House, Kingsway, London, W.C.2.

Remote Control of Aircraft

(Illustrated above)

Following the successful development of an 80Mc/s equipment for the remote control of the Jindivik and other target aircraft, the G.E.C. has now developed a more compact equipment for operation at 400Mc/s.

A complete installation would usually consist of an eight tone receiver, a two tone receiver, and two power units. The eight tone receiver is used for normal flight control, the two tone receiver is used as an emergency 'destroy' facility, and the two separate power units are used to ensure that, in the event of a failure of the main aircraft supply, the 'destroy' facility is still available.

The eight tone receiver consists of a 400Mc/s f.m. high gain receiver together with audio tone selector circuits and relays. With eight audio tones used in pairs, 24 combinations can be transmitted and made to close 24 different command circuits at the receiver output.

The two tone receiver is a simplified version of the eight tone receiver and is intended for the reception of a single command only.

The power unit is used to supply the h.t. and l.t. power (totalling 60W) to drive each receiver. It consists of a transistorized regulator and convertor and is driven from the 28V aircraft supply.

In the illustration the eight tone receiver is shown on the left, the two tone receiver at the rear and the power pack on the right.

EE 14 756 *for further details*

Missile Attitude Indicator

(Illustrated below)

The G.E.C. missile attitude indicator has been developed for G.W. Department, R.A.E., to measure the attitude in roll, pitch and yaw of a Skylark research rocket during flight.

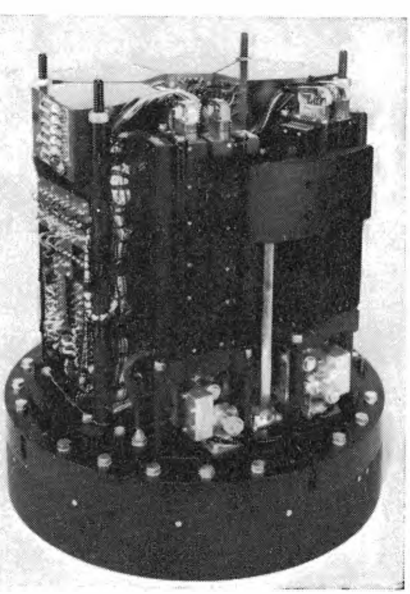
Two high power c.w. X-band transmitters are stationed on either side of the range centre line to illuminate the rocket. A special bay in the rocket, mounted on a strong ring, contains the circuits and three flush mounted microwave aerials, arranged to form two orthogonal interferometers. The microwave circuit is almost entirely contained in the strong ring.

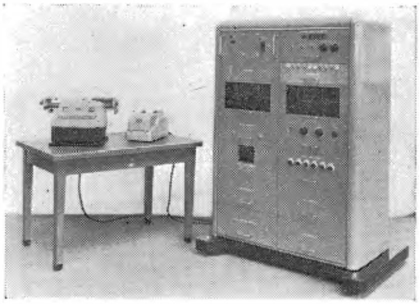
By a phase comparison technique, the angles between each interferometer base line and each transmitter are measured. Any three of these four angles can be combined with the knowledge of the rocket's position in space to determine roll, pitch and yaw to within a few minutes of arc. The interferometer measures the phase difference between signals received at two aerials from a single source.

The G.E.C. system avoids the difficulty of maintaining equal phase responses in two amplifier channels. By means of a microwave single-sideband modulator, a low frequency ω is added to the local oscillator input to one mixer.

The frequency and phase differences are preserved during mixing and, after adding the two signals, subsequent amplification is in a common intermediate frequency amplifier. After detection and further amplification, the resultant low frequency is compared in phase with a reference frequency ω, derived directly from the modulator, to give a voltage proportional to the phase difference existing at the receiving aerials. This is then telemetered back to the ground.

EE 14 757 *for further details*





HONEYWELL CONTROLS LTD

Ruislip Road, East Greenford, Middlesex.

Data Handling System

(Illustrated above)

The Honeywell data handling system will accept a number of different electrical signals from thermocouples, pressure transducers, etc. In one operation it measures and digitizes these variables and presents the information in a digital form in lights on the front of the unit and, via an electric typewriter, on a log sheet. In addition to this straightforward function it will also linearize non-linear inputs where necessary and check inputs against high and low level set-point alarms. The system includes a very convenient method of changing alarm levels and transducer spans by means of pin boards.

The system scans against high and low alarm levels at six points per second, logging all variables on an electric typewriter at one point per second. The accuracy of measurement is ±0·1 per cent of reading, or one digit.

The system is useful wherever a large number of variables (say 40 or more) are required to be recorded accurately and in digital form and where it is necessary to check a large number of points against alarm levels.

EE 14 758 *for further details*

Automatic Wave Analyser

The automatic wave analyser is a highly accurate heterodyne type analyser designed to produce automatically plotted frequency analyses of complex data signals having components in the range of 3c/s to 2kc/s. By virtue of its extreme selectivity, it provides a method of analysing data below the normal lower frequency limit of many wave analysers. The automatic control circuits make possible the analysis of large volumes of data with only occasional attention by the operator.

The equipment is suitable for use with magnetic tape loops, which permit analysis of signals having durations as short as a few seconds. For such applications, Honeywell offer suitable magnetic tape recording and playback equipment.

The analysis may be amplitude versus frequency, power versus frequency and in some instances the analyser may be used to provide an amplitude-versus-time analysis of a signal having time-varying frequency components.

EE 14 759 *for further details*

MARCONI'S WIRELESS TELEGRAPH CO. LTD

Chelmsford, Essex.

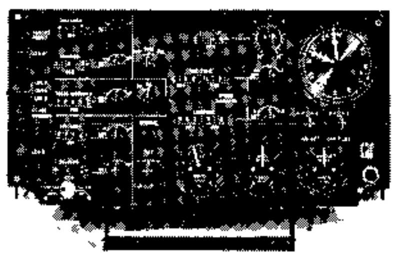
Doppler Navigator

(Illustrated below)

The AD2300 series of Doppler navigators employs a continuous-wave, frequency modulated system operating at a frequency of 800Mc/s (±50) at a power of 1W. In order to meet a wide variety of requirements the equipment is available either as a sensor incorporating an indicator drive unit (giving the basic information of ground speed and drift angle, together with the distance flown) or as a sensor plus either of two coputers, to give much more information. The three variants are indicated by the suffix 'A,' 'B' or 'C' as under:—

TYPE AD 2300A (sensor and indicator drive unit), which provides:—
(a) Ground speed and drift angle
(b) Distance flown

TYPE AD 2300B (using computer type 4381 instead of the indicator drive unit). This provides:—
(a) Ground speed and drift angle as for type AD 2300A



(b) Miles to go along a selected leg
(c) Error in miles (left or right) across the selected leg
(d) Total miles to go to destination
(e) Plan of next selected leg to be flown
(f) Steering information to pilot's direction indicator in flight system (or to auto-pilot).

TYPE AD 2300C (using computer type 4382 (illustrated) instead of the indicator drive unit). This provides:—
(a) Ground speed and drift angle
(b) Wind speed and direction
(c) Present position in latitude and longitude
(d) Magnetic variation
(e) Total miles to go for destination
(f) Miles to go on selected leg being flown
(g) Plan of next selected leg to be flown
(h) Steering information to pilot's direction indicator in flight system (or to auto-pilot).

The three variants indicated above cover the requirements of all categories of aircraft. If desired, a roller map presentation of the aircraft track can be provided with the latter two equipments and such a one (manufactured by Kelvin & Hughes Ltd) was shown on the stand in conjunction with the Marconi AD 2300B Doppler navigator.

The AD 2300 comes into operation at take-off and is fully effective up to a ceiling of 50 000ft, and angles of bank up to 30°. Full accuracy is maintained with a drift angle of up to 30° (either port or starboard). The ground speed range lies between 80 and 900 knots.

An electro-mechanical tracking system is incorporated, whereby the equipment is rendered 'fail-safe'—that is, as long as a usable Doppler signal is present the tracker must give a correct answer. Should the signal fail the tracker automatically reverts to 'memory,' locking on to the signal again when the level is restored. After exhaustive flight trials over many world air-routes it has been found that even with surface winds of less than 5 knots and a smooth sea below, an ample margin of signal keeps the equipment fully operational even at heights in excess of 45 000ft.

EE 14 760 *for further details*

Single Channel V.H.F. Direction Finder

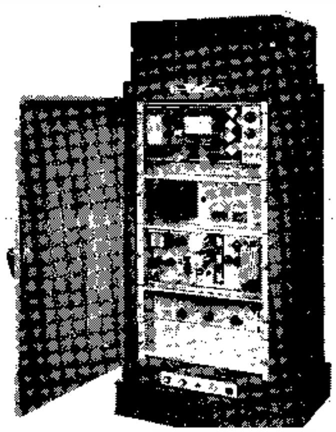
(Illustrated below)

The Marconi AD200 series of ground-based v.h.f. direction finders are installed all over the world, well over 200 being in constant daily use. This series is now being superseded by the AD 210 series.

The type AD 210C, which was shown in operation at Farnborough, caters specially for the requirements of the smaller airport in providing d.f. facilities in the simplest and least expensive form. Although basically a single channel equipment, multi-channel working can be effected by the introduction of local equipments connected in cascade.

Particular attention has been paid in the design of this equipment to the achievement of the highest operational standard. Operation is in the frequency range 100 to 156Mc/s; any one of five crystal-controlled frequencies are available at the main display by push-button selection.

The presentation of bearings is entirely automatic on an 8in diameter bearing indicator. The display units are normally desk-mounted or, where desired, incorporated in the airport's main control desk. A slave display may be operated in addition to the main dis-



play and can be situated up to 500ft from it.

A simple remote system, operating over standard telephone lines, permits remote operation up to a distance of 8 miles.

The bearing accuracy (when receiving a vertically polarized wave) is extremely good, the operational error being of the order of $1\frac{1}{2}°$ to $2°$ over the whole band.

EE 14 761 *for further details*

MICROCELL LTD

56 Kingsway, London, W.C.2.

HIGH SPEED SEQUENTIAL SWITCH

Microcell Electronics, a division of Microcell Limited have recently solved, in collaboration with the Ministry of Supply, the problem of hazard-free high-speed sequential unlatching of electro-mechanical locks.

Electrical reasons prohibit the unlatching of the locks at the same time, and a requirement was thus established for a high-speed sequential switch which would unlatch one at a time—the time interval between unlatching being 10msec.

Severe restrictions were placed on the unit: physical size is limited, yet it must not be a 'throw away' device, but one capable of undergoing normal electronic test routines, and repair if damaged.

This problem has been solved by a fully-transistorized electronic switch. This switch incorporates a number of sequential stages—each stage switching a current pulse of 5A to the electro-mechanical locks. The design ensures that no stage operates before the preceding one has completed its functions. The switch has the ability to wait at a sticking lock for a limited time; and special safety devices then operate, to allow the sequence to continue should the lock refuse to unlatch.

The electronic switch scores over a similar electro-mechanical device, in that there are no moving parts to jam or wear.

EE 14 762 *for further details*

MULLARD EQUIPMENT LTD

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

ELECTROMAGNETIC SURVEY APPARATUS

A Scottish Aviation 'Twin Pioneer' aircraft specially adapted for geophysical air survey work has been fitted with electromagnetic survey apparatus supplied by Mullard Equipment Limited.

The aircraft has been built and equipped for Rio Tinto, an international group of companies with extensive mining interests, and was exhibited by Scottish Aviation at Farnborough.

The electromagnetic apparatus is designed to detect zones of conductive minerals, such as massive deposits of metallic sulphides, lying near the earth's surface. It consists essentially of a transmitter, a receiver and a display unit, together with ancillary apparatus that provides additional information which may be useful in interpreting the primary data.

The output of the transmitter is used to produce an alternating magnetic field in a tuned coil carried on an extension to the starboard wing-tip of the aircraft. This field induces a voltage in a second coil similarly mounted on the port wing and coupled to the receiver. Both the transmitter and receiving coils may be oriented to bring their axes horizontal or vertical to the aircraft's normal line of flight, depending on the more likely plane of the ore bodies in the area being surveyed.

When the aircraft is flying beyond the range of possible ground influences the induced voltage in the receiving coil is backed-off by a voltage of equal magnitude but opposite phase derived from a secondary winding on the transmitting coil, and the system is thus in balance. However, when the aircraft reaches the proximity of a large conductive body, eddy currents are induced in the body which modify the magnetic field in the receiving coil. As a result, the balance of voltages at the receiving coil is disturbed and an unbalance signal produced. This signal is pre-amplified and fed to the receiver where it is further amplified and resolved into in-phase and quadrature components. The resulting output signals are then amplified for recording on the multi-channel recorder carried in the aircraft.

The recording of both in-phase and quadrature signals, combined with a suitable choice of transmitter frequency, enables the apparatus to distinguish between potentially valuable deposits and those of poor conductivity, such as swamp and saline overburden.

In addition to the in-phase and quadrature outputs, a third channel, operating on a nearby frequency, indicates the amount of atmospheric noise, and a fourth provides for recording the strain in the wing struts to reveal the presence of any wind gusts which might produce spurious electromagnetic readings. Further channels are provided for recording the responses of other survey instruments carried in the aircraft, namely, a magnetometer, a radio altimeter, and a scintillometer for detecting gamma radiation from potential sources of uranium and thorium.

EE 14 763 *for further details*

REDIFON LTD

Broomhill Road, London, S.W.18.

H.F. TRANSMITTERS

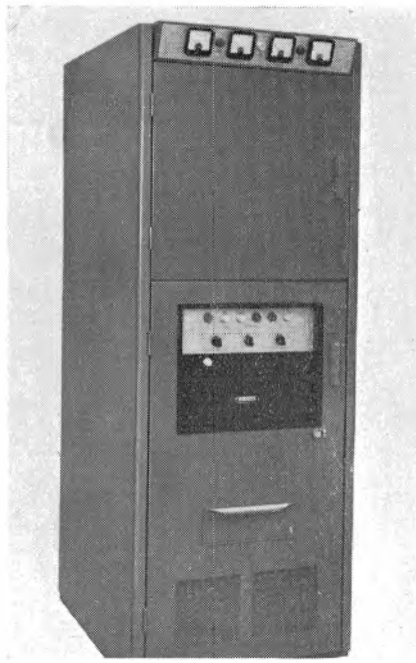
(Illustrated above right)

Redifon have announced a completely new range of h.f. transmitters to meet the many different requirements for aeronautical communications. This range, known as the G.420 series, is based on an s.s.b. transmitter of 1·5kW p.e.p. covering the frequency range 1·5 to 30Mc/s, and including models for A1, A2, A3, A3a, A3b and F1, F2, F3, F4 and F6 with external drive. Control can be manual, local or extended, or fully remote with servo tuning. Unit construction has been adopted so that a change in service requirements can be made at any time simply by the addition of the appropriate drive unit. Up to eight spot frequencies are available in one r.f. unit or, alternatively, where immediate changes of frequency are required, up to eight pre-tuned r.f. units can be supplied from common drive circuits.

The G420 series of transmitters are available in three basic forms:

(a) *Manual*, for single channel operation and requiring manual adjustment for change of frequency. Remote control facilities can be provided, but not to change frequency.

It is practical to add additional units to an existing transmitter to convert from manual to servo control.



(b) *Servo control*, for single channel operations and providing for selection of up to eight spot frequencies remotely. Full remote control facilities can be provided.

(c) *Multi-bay*, for single channel operation, and providing for very rapid remote selection of up to eight radio frequency units. This version is particularly intended for use where frequency change time must be minimized and where a large number of changes will be required daily.

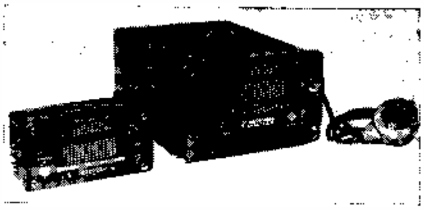
EE 14 764 *for further details*

V.H.F. RADIOTELEPHONE

(Illustrated in next column)

The GR.318 is an amplitude modulated v.h.f. radiotelephone of advanced design for speech communication between mobile vehicles and a suitable main station. It is a compact 4 to 5W equipment of low power consumption and great flexibility in mounting arrangements; these features contribute to its particular suitability for airport mobile ground services.

Increased reliability and economy in



battery consumption are obtained by judicious use of transistors. Power conversion for h.t. is provided by an efficient transistor convertor unit—in addition the modulator which also uses transistors, serves the dual function of receiver output stage.

Because of its small size the GR.318 may be conveniently installed under the dashboard or the cab head of the vehicle driving compartment.

To maintain complete flexibility in installation an extension control unit is available. This unit duplicates all the controls of the main unit including channel selection for the dual channel model. Mounting trays for both installation arrangements are supplied with the equipment, together with a robust hand microphone. A telephone handset or microphone and headphones, however, may alternatively be used.

Models are available for seven frequency bands in the range 72 to 175Mc/s.

EE 14 765 *for further details*

ROYSTON INSTRUMENTS LTD

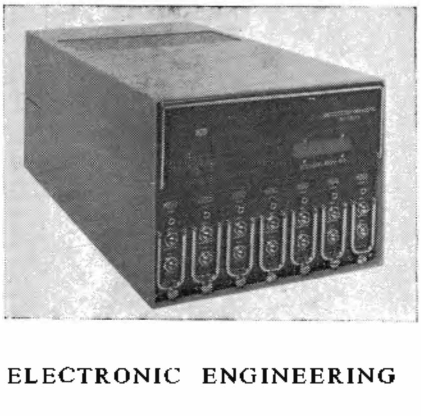
Canada Road, Oyster Lane, Byfleet, Surrey.

FLIGHT DATA RECORDING EQUIPMENT

(Illustrated below)

The MIDAS comprehensive routine flight data recording system using magnetic tape has been developed from the background of this Company's experience in providing recording and processing equipment for research and testing purposes on aircraft. Its function is to provide a complete, accurate and permanent record of every important operating parameter, from 'run up' to landing, throughout the whole life of an aircraft. Speeded up automatic data processing permits the record to be translated into a compact, edited and usable form.

The recorder consists of a special low speed magnetic tape data recorder, based upon the principles which have been tried and proven in the MIDAS system over many years. The recorder has a total of seven independent signal channels operating in parallel, one of which is reserved for G.M.T., date, flight number, etc. The recording capacity of the remaining six



channels is increased by a time sharing switch which can select a large number of parameters in sequence. In fact, there are six switch banks, continuously driven by a common motor, each one of which is feeding a corresponding recording channel and so combining a number of different parameters thereon. The recorder is a completely standard item and will, unless otherwise specified, be identical in all aircraft, irrespective of the installation and the parameters to be recorded. The time sharing switch, however, while being standard in construction and detail, is wired to the muti-way input connexions in such a way as to suit the particular aircraft instrumentation. Physically, the time sharing switch will be mounted at the convenient centre of transducer wiring (i.e. probably in the cabin) and it may be attached physically to the rear of the recorder via unitor type plugs and sockets. Alternatively, where the user desires an emergency ejection facility, the recorder only can be fitted in the tail and a tenway cable is sufficient to connect the two units.

The time sharing switch will determine the total number of parameters to be sampled and the sampling rate. The multi-way switch will make one rotation in five seconds so that every parameter will be sampled at a minimum rate of once every five seconds. The sampling rate may be doubled or trebled by paralleling sections of the switch at discrete intervals, so that these signals will be sampled two or three times in each revolution of the switch. Therefore, any number of channels and combinations of sampling rates may be secured within a maximum of 300 channels, each of which is then sampled every five seconds.

The magnetic tape spools which form the recording medium are preloaded into a cassette which can be inserted or removed by unskilled personnel. This factor, together with a recording capacity of 160 hours, means that there is no difficulty in retaining a complete record on the longest of flight circuits, even when the recording tape is replaced only on return to main base. The recorder is designed for indefinite life, will allow a very high standard of accuracy, is of small size (1ATR) and weighs less than 40lb.

The fact that the recorder has a very high input impedance and has built-in facilities for the conversion of signals from three wire synchro's into analogue voltages means that in nearly every case simple connexions on to existing instrumentation are all that is required in the way of installation.

Playback and data processing equipment is available for use with this recorder.

EE 14 766 *for further details*

THERMIONIC PRODUCTS (ELECTRONICS) LTD

Hythe, Southampton.

MULTI-CHANNEL TAPE RECORDER

(Illustrated above right)

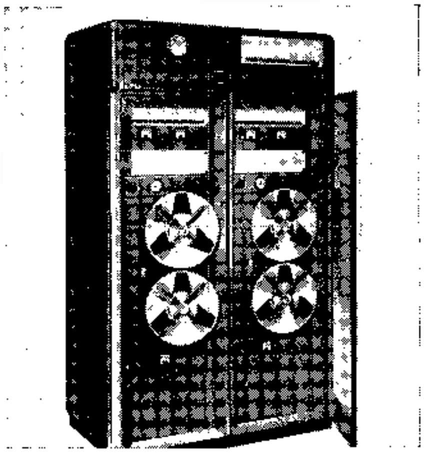
These recorders are built up of a number of separate units and by using the required selection a recorder to suit most applications can be supplied.

The recorder illustrated is the model 1A/1A and provides for the simultaneous recording of any number of speech and/or coded signals up to a maximum of 20. This normally includes one time coded channel and one automatic spare channel.

A voice/carrier stop start unit is also fitted allowing a maximum of 24 hours unattended operation, extended in proportion to the total time during which signals are not being received.

Continuous monitoring is fitted on all channels, the one spare being switched into operation on a first failure without interference with the continuity of recording. Full replay facilities are provided on the 'stand by' bay.

The equipment is fully transistorized and, where suitable, printed circuits have



been used to ensure reliability. The clock and coding unit is crystal controlled for accuracy and has plug-in stages to simplify servicing if needed.

EE 14 767 *for further details*

THORN ELECTRICAL INDUSTRIES LTD

105-109 Judd Street, London, W.C.1.

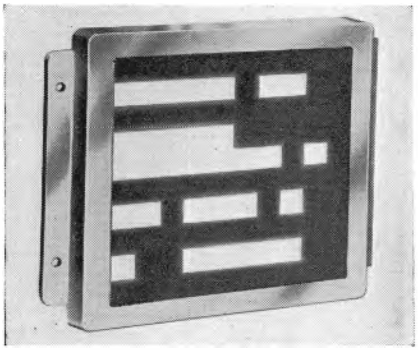
ELECTROLUMINESCENT INDICATOR

(Illustrated above right)

A 100 dot electroluminescent indicator block now being manufactured by the Aircraft Components and Connector Division of Thorn Electrical Industries Ltd, applies a new principle to digital read-out and progress chart systems.

This component has a depth of only 0·265in and occupies a fraction of the space used by lamp indicators. No filaments are employed; the unit thereby generates virtually no heat at all. The component consists of 100 rectangular phosphor dots sandwiched between two electrically conducting surfaces. By applying alternating voltage to these dots they are excited to luminescence. Variation in the pattern of illuminated dots enables the indicator to read-out any desired character or figure or to provide an accurate reading of progress.

The brightness rating of the indicator is 15 foot lamberts. To develop the in-



tensity required it is desirable to use a high frequency supply of a voltage of about 350V.

EE 14 768 *for further details*

ULTRA ELECTRIC LTD

Western Avenue, London, W.3.

CRASH POSITION INDICATOR

This is an interesting extension of the SARAH search and rescue equipment. Its use would ensure that even in conditions of fog or low cloud, or in isolated conditions on land or at sea, the position of survivors or wreckage could be swiftly and accurately detected.

The crash position indicator (c.p.i.) is a beacon transmitter constructed over a thin plate aerial. It is embedded in tough plastic foam, which is shaped to form an aerofoil with a laminated skin of nylon.

In use, the c.p.i. is incorporated as an integral section of an aircraft fuselage. A sensitive release mechanism, operated by wires stretched along the aircraft structure, detects the first stage of collision or abnormality in the structure. In less than the 100msec in which a high speed modern aircraft can be totally destroyed, the c.p.i. will be ejected clear, delivered a safe distance from the wreck and commence a beacon transmission.

The transmission can be detected under normal and favourable conditions up to 30 miles away, with 70 miles as a maximum range. Under severe conditions, for example with the c.p.i. buried under snow in a deep valley, a range of 5 to 10 miles is still maintained.

The transmitter will work immersed in water and has an uninterrupted operational life of over 100 hours.

On rock or ice the c.p.i. will land safely at speeds up to 40mile/h and at several times this speed on snow or soft ground. It will tumble to earth at approximately 26mile/h and float 80 per cent out of water.

The c.p.i. signal is compatible with SARAH search equipment, which is standard equipment with the military forces of the British Commonwealth and most N.A.T.O. countries.

EE 14 769 *for further details*

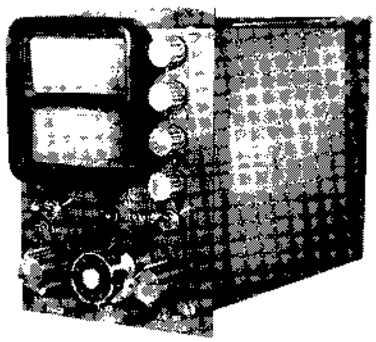
ENGINE CONDITION ANALYSER

(Illustrated above right)

This is a new instrument which has been developed to display information of engine temperature and vibration in a form which can be assessed at a glance. It is a versatile system which can be used for pre-flight checks, flight monitoring and routine maintenance.

Temperature indications from up to 40 thermocouples are displayed in turn across the upper cathode-ray tube in the display unit. These temperature indications are divided into four groups, i.e., one group per engine. By means of a front panel control the ten temperature points of any individual engine may be displayed across the full width of the screen. Two straight horizontal lines are traced across the temperature display: one of these is the temperature base line and the other is a calibration line. Also on the left-hand side of the temperature display is a scale with markers at intervals of 25°C. The sensitivity is 300°C/in, and the unit is adjusted before installation so that the nominal engine running temperature is at mid-screen.

The base line is fixed in one of four positions, selected by a front panel switch. The four positions are for Take



Off, Climb, Cruise and Descent. With correct functioning of the engine the base line passes through the mean of the engine temperature indications: temperature drift at any of the check points is thus immediately obvious. The position of the line for each of the four conditions is preset to the particular type of engine before the analyser is installed.

The temperature calibration line can be moved up or down by a front panel control: the temperature of any point can be read to within ½ per cent of full scale by adjusting the calibration line until it coincides with the point, and reading the scale on the adjusting knob. If only an approximate reading is required, the temperature can be estimated from the cathode-ray tube scale alone.

Indications from two vibration transducers on each engine are continuously displayed on the lower cathode-ray tube in the display unit. A vibration danger line is traced across the display. The height of this line is adjusted to the particular engines before the analyser is installed, and is set so that the indications remain below the line if the engines are functioning correctly. The points of the display correspond to first order amplitudes of vibration: vibrations at frequencies other than that corresponding to the engine rotation speed are not shown.

EE 14 770 *for further details*

WAYNE KERR LABORATORIES LTD

Roebuck Road, Chessington, Surrey.

DISTANCE METER

(Illustrated below)

This instrument operates in conjunction with non-contacting probes to provide a basically linear method of measuring distance from 50μin to 0·5in. Distance is read directly from a moving-coil meter, the full-scale range depending on the probe in use. Four probes are available for full-scale ranges of 1, 10, 100 and 500 thousandths of an inch respectively.

Readings are accurate to ± 2 per cent of full-scale deflexion and the discrimination is better than 0·5 per cent of full-scale deflexion.

The length of the probe cable is 10ft, enabling the instrument to be situated well clear of the test structure. An outlet from the front-panel is provided for connexion to recorders or servo-control systems.

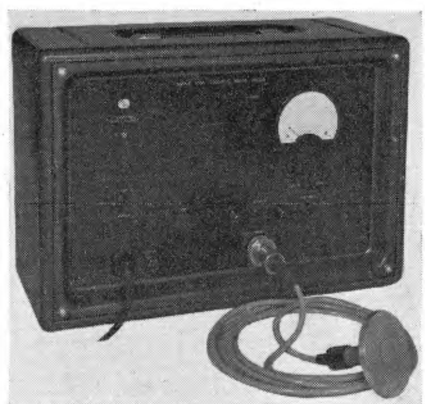
The instrument provides a rapid and straightforward means of measuring distance in precision grinding processes and can readily be adapted for such applications as the dynamic balancing of helicopter blades and the measurement of verticality of control rods in nuclear power stations.

Direct indication of expansion is obtained without imposing any load on the item under observation and changes in dimensions can be continuously monitored or recorded against time or temperature.

Other applications include any distance measurements of up to 0·5in where direct-reading and continuous monitoring facilities are required, the routine inspection of piece parts and all *in-situ* measurements during machining operations. The instrument can be set up to give automatic rejection of out-of-tolerance parts.

The design is centred on a high-gain amplifier with a negative-feedback loop completed by the capacitance between the probe and test structure. The amplifier input is a reference current derived from an internal oscillator. The output from the amplifier is applied to a feedback voltmeter circuit, and the meter reading is directly proportional to the probe/structure separation.

EE 14 771 *for further details*



Meetings this Month

THE BRITISH INSTITUTION OF RADIO ENGINEERS

All London meetings will be held at the London School of Hygiene and Tropical Medicine, Keppel Street, Gower Street, London W.C.1 at 6.30 p.m. unless otherwise stated.
Date: 28 October.
Inaugural Meeting. Addresses on:
Radio—Its Impact on Shipping.
By: Captain J. D. F. Elvish.
A Historical Survey of Radar and Radio Aids to Aircraft Navigation.
By: Air Marshal Sir Raymund G. Hart.
Date: 18 November. Time: 3 p.m. and 6 p.m.
Half-day Symposium: Electronic Digitizing Techniques.
Opened by: G. J. Herring.

Merseyside Section

Date: 2 November. Time: 7 p.m.
Held at: University Club, Liverpool.
Lecture: Electronics in the Auto-Pilot of the Firestreak Missile.
By: A. Bedford.
Date: 10 November. Time: 7 p.m.
Held at: University Club, Liverpool.
Lecture: The Use of Transistors in Communications and Control.
By: E. Wolfendale.

North Eastern Section

Date: 11 November. Time: 6 p.m.
Held at: Institution of Mining and Mechanical Engineers, Neville Hall, Westgate Road, Newcastle-on-Tyne.
Lecture: Electronic Welding Controls.
By: C. R. Bates.

North Western Section

Date: 12 November. Time: 6.30 p.m.
Held at: Reynolds Hall, College of Technology, Sackville Street, Manchester 1.
Lecture: Progress in Permanent Magnet Materials.
By: J. E. Gould.

South Western Section

Date: 18 November. Time: 7 p.m.
Held at: School of Management Studies, Unity Street, Bristol 1.
Lecture: Data Recording and Presentation.
By: D. W. Thomasson.

West Midlands Section

Date: 11 November. Time: 7.15 p.m.
Held at: Wolverhampton and Staffordshire College of Technology, Wulfruna Street, Wolverhampton.
Lecture: Recent Developments in Semi-Conductor Rectifiers

THE INSTITUTION OF ELECTRICAL ENGINEERS

All London meetings will be held at Savoy Place commencing at 5.30 p.m.

Electronics and Communications Section

Date: 2 November.
Lecture: Some Comments on the Classification of Waveguide Modes.
By: A. E. Karbowiak.
Date: 9 November.
Lecture: Theory of the Travelling-Wave Parametric Amplifier.
By: Professor A. L. Cullen.
Lecture: The Gain of Travelling-Wave Ferromagnetic Amplifiers.
By: P. J. B. Clarricoats.
Lecture: Some Properties of Travelling-Wave Resonance.
By: J. R. G. Twistleton.
Lecture: Saturation Effects in a Travelling-Wave Parametric Amplifier.
By: A. Jurkus and P. N. Robson.

Measurement and Control Section

Date: 3 November.
Lecture: An Analogue Electronic Multiplier using Transistors as Square-Wave Modulators.
By: P. Gleghorn.

Cambridge Electronics and Communications Group

Date: 3 November. Time: 8 p.m.
Held at: Cavendish Laboratory, Free School Lane, Cambridge.
Section Chairman's Address: The Development of Eurovision.
By: M. J. L. Pulling.

Mersey and North Wales Centre

Date: 11 November. Time: 6 p.m.
Held at: The Temple, Dale Street, Liverpool.
Lecture: Vision and Position—Two Electronic Aids to Marine Navigation.
By: R. B. Mitchell and C. Powell.

North-Eastern Measurement and Electronics Group

Date: 2 November Time: 6.15 p.m.
Held at: Rutherford College of Technology, Newcastle-on-Tyne 1.
Lecture: The New Post Office 700-Type Telephone.
By: F. E. Williams and F. C. Carter.
Supported by a paper on. The Design of an Automatic Sensitivity Control for a new Subscriber's Telephone Set—The British Post Office 700-Type Telephone.
By: F. E. Williams and F. A. Wilson.

North Lancashire Sub-Centre

Date: 11 November. Time: 7.15 p.m.
Held at: North-Western Electricity Board Demonstration Theatre, North Road, Lancaster.
Lecture: H.V. D.C. Transmission, with Special Reference to the Cross Channel Cable.
By: F. J. Lane.

North Scotland Sub-Centre

Date: 12 November. Time: 7 p.m.
Held at: Electrical Engineering Department, Queen's College, Dundee.
Lecture: The Application of Transistors to Line Communication Equipment.
By: H. T. Prior, D. J. R. Chapman and A. A. M. Whitehead.
Date: 13 November. Time: 7.30 p.m.
Held at: Robert Gordon's Technical College, Aberdeen.
As above.

North-Western Electronics and Communications Group

Date: 11 November. Time: 6.15 p.m.
Held at: Engineers' Club, Albert Square, Manchester.
Lecture: Ultra-Sound Image Camera.
By: C. N. Smyth and J. Sayers.

Reading

Date: 23 November. Time: 7.15 p.m.
Held at: George Hotel, King Street, Reading.
Lecture: An Introduction to Electronic Computers.
By: R. C. M. Barnes.

South-East Scotland Sub-Centre

Date: 3 November. Time: 7 p.m.
Held at: Carlton Hotel, North Bridge, Edinburgh.
Lecture: Silicone Electrical Insulation.
By: J. H. Davis.

Southern Centre

Date: 11 November. Time: 6.30 p.m.
Held at: S.E.B. Showrooms, 17 New Canal, Salisbury.
Lecture: The Planning and Installation of a Television Transmitting Station.
By: D. B. Weigall.

Rugby Sub-Centre

Date: 11 November. Time: 6.30 p.m.
Held at: Rugby College of Technology and Arts.
Lecture: The Universe Explored by Radio Astronomy.
By: F. C. Smith.

South Midland Electronics and Communications Group

Date: 23 November. Time: 6 p.m.
Held at: James Watt Institute.
Lecture: Learning Machines.
By: P. Huggins.

South-Western Sub-Centre

Date: 12 November. Time: 3 p.m.
Held at: Electric Hall, Upper Union Street, Torquay.
Lecture: Subscriber Trunk Dialling.
By: D. A. Barron.

South-West Scotland Sub-Centre

Date: 4 November. Time: 6 p.m.
Held at: Institution of Engineers and Shipbuilders, 39 Elmbank Crescent, Glasgow C.2.
Fiftieth Kelvin Lecture on: The Geophysical Year 1957-58.
By: Sir David Brunt.

Western Centre

Date: 9 November. Time: 6 p.m.
Held at: Electricity House, Bristol.
Lecture: Subscriber Trunk Dialling.
By: D. A. Barron.

THE INSTITUTE OF PHYSICS

Electronics Group

Date: 10 November. Time: 5.30 p.m.
Held at: 47 Belgrave Square, London S.W.1.
Lecture: Electronic Properties of Diamonds.
By: Dr. F. C. Champion.

Manchester and District Branch

Date: 16 November. Time: 7 p.m.
Held at: College of Science and Technology, Manchester.
Lecture: High Altitude Research during the International Geophysical Year.
By: Dr. R. L. F. Boyd.

North-Eastern Branch

Date: 6 November. Time: 6.15 p.m.
Held at: King's College, Newcastle-on-Tyne.
Lecture: The Scientific uses and Rockets and Satellites.
By: Professor H. S. W. Massey.

THE JUNIOR INSTITUTION OF ENGINEERS

Date: 30 October. Time: 7 p.m.
Held at: Pepys House, 14 Rochester Row, Westminster, London S.W.1.
Lecture: Planning a Computer Job.
By: M. G. Ferrand.
Date: 13 November. Time: 7 p.m.
Held at: Pepys House, 14 Rochester Row, Westminster, London S.W.1.
Chairman's Address: Electronic Components.
By: W. C. C. Ball.

THE TELEVISION SOCIETY

All meetings will be held at the Cinematograph Exhibitors' Association, 164 Shaftesbury Avenue, London W.C.2 commencing at 7 p.m.
Date: 6 November.
Lecture: Deflection Techniques for 110° Picture Tubes.
By: B. Eastwood.
Date: 20 November.
Lecture: Television Film Production.
By: J. K. Byers.

PUBLICATIONS RECEIVED

COLD CATHODE VOLTAGE STABILISERS is an application report giving a survey of all the characteristics which are likely to be of importance when these tubes are in normal use. Standard Telephones and Cables Limited, Footscray, Sidcup, Kent.

THE EFFECT OF TECHNOLOGICAL PROGRESS ON EDUCATION is a new classified bibliography published by The Institution of Production Engineers which has been compiled from British sources 1945/57. The bibliography is not comprehensive, but is intended to be a guide to the development of technical education during the period covered. Independently of its use as a bibliography, it should serve as a summary of opinion on the subject. Publications Department, The Institution of Production Engineers, 10 Chesterfield Street, London W.1. Price 14s. 6d. plus 1s. postage.

A MAP OF THE MOON has been made available by the Missile and Space Vehicle Department of the General Electric Company, Room 4C, 3198 Chestnut Street, Philadelphia 4, Pennsylvania, free of charge. Printed in large size, the moon map makes all charted geography of the visible side of the moon easily identified and located.

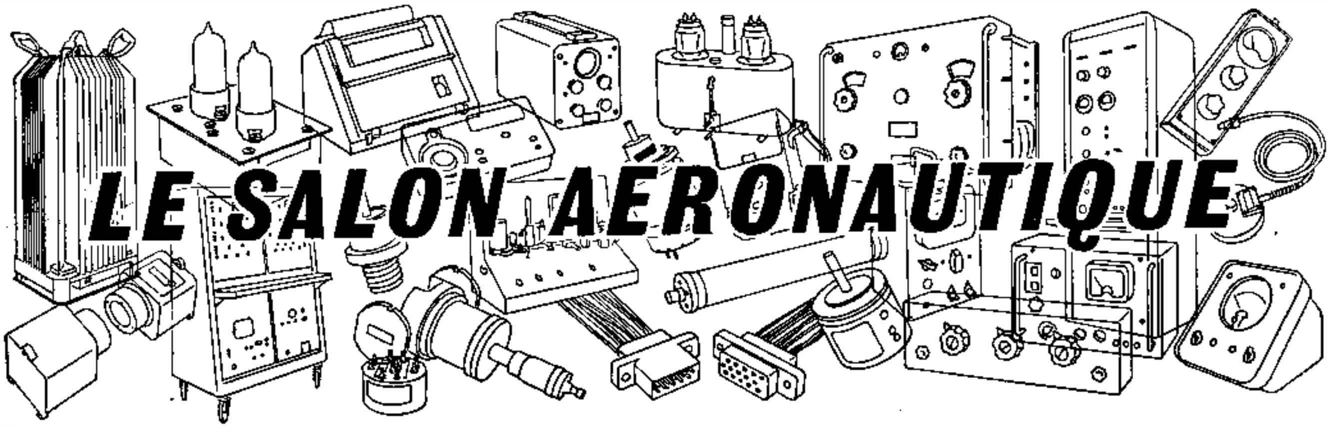
COLD-CURE SILASTOMER is a booklet describing the cold curing silicone rubbers. These rubbers, which are made in three grades, have been developed at the Midland Silicones plant at Barry, Glamorgan and were recently put on a regular production basis. This booklet describes in detail the properties of these rubbers, explains how they are processed, lists established uses and suggests many more. The publication is well illustrated and has several tables on which the properties of these products are clearly set out. Midland Silicones Ltd., 68 Knightsbridge, London S.W.1.

GRAPHICAL SYMBOLS FOR TELECOMMUNICATIONS (B.S. 530:1948), SUPPLEMENT NO. 6:1959. SERVICES PREFERENCES AND ADDITIONAL SYMBOLS is published by the British Standards Institution. This new supplement to B.S. 530 will meet the needs of the many users of that standard who are either in the Services or connected with the supply of equipment to them. The new supplement does not constitute part of the standard proper, but relates entirely to the special requirements of the Services. It gives certain symbols additional to those in B.S. 530 and the first five Supplements; it indicates Services' preferences concerning alternative symbols, as well as depicting a few deviations from B.S. 530 which the services find necessary. British Standards Institution, Sales Branch, 2 Park Street, London W.1. Price 3s. (Postage extra to non-subscribers.)

ABBREVIATIONS OF ASSOCIATIONS, INSTITUTIONS, RESEARCH AND TRADE ASSOCIATIONS has now been reprinted in an enlarged and revised form. In selecting abbreviations to be included in the book, the choice was based on those which are likely to be met in the electrical and allied industry. The British Electrical and Allied Manufacturers' Association, 36 Kingsway, London, W.C.2. Single copies 1s. 6d., 20 for £1 and 100 for £4.

(Pages 631 to 644 French and German Sections)

LE SALON AÉRONAUTIQUE



Une description, basée sur des renseignements fournis par les fabricants, de certains des appareils électroniques exposés au dernier salon de la Société des Constructeurs d'Avions Britanniques.

Traduction des pages 624 à 699

BLACKBURN ELECTRONICS LTD

Brough, Yorkshire

(*Illustration à la page 624*)

VOLTMÈTRES À TROIS ET QUATRE SIGNES

Ces voltmètres digitaliseurs électroniques ont été conçus pour des applications exigant des mesures précises de tension continues. Ces dernières comprennent les sorties et les entrées de calculateurs, les mesures d'alimentations de référence, les mesures d'extensiomètres, le contrôle en cours de fabrication de composantes électroniques et électriques, ainsi que le traitement et la présentation de données (c'est à dire: conversion analogique à numérique).

Le dessin du circuit de ces instruments garantit un facteur élevé de sécurité, de même qu'une précision de ±0,1% pour le modèle B.I.E. 2113 et ±0,5% pour le modèle B.I.E. 2114, sur n'importe laquelle des trois gammes du voltmètre. Ces gammes sont de 0 à ±9,60; 9,60 à ±96,0; ±96,0 à ±999 V.c.c. sur le modèle B.I.E. 2113, et de 0 à ±9,600; ±600 à ±96,00 et ±96,00 à 999,9 V.c.c. sur le modèle B.I.E. 2114. L'impédance d'entrée des deux modèles est de 50 mégohms dans la gamme inférieure, et de 5 mégohms dans les gammes supérieures. Le changement de gamme et le choix du signe s'effectuent automatiquement et tous deux sont présentés sur l'indicateur visuel à tubes numériques au néon.

Les temps de lecture de ces instruments sont de 13 m.sec. pour le modèle à trois signes B.I.E. 2113 et de 103 m.sec. pour le modèle à quatre signes B.I.E. 2114.

Durant le fonctionnement, les instruments comparent la tension d'entrée avec une rampe qui est linéaire à ±0,01% près. Le nombre de cycles produits par l'oscillateur piloté par quartz est compté entre la durée du passage à 0 Volt de la tension de rampe et l'amplitude de la tension d'entrée. Le chiffre de ce compte représente la valeur numérique de la tension et il est indiqué directement sur les tubes numériques au néon. L'étalonnage transmis à une pile Weston courante est incorporé dans les deux instruments.

Des impulsions négatives de 80 Volts actionnent un appareil imprimeur 'en ligne', une machine à écrire électrique, un enregistreur à cartes perforées ou à bandes magnétiques, ou bien encore des téléindicateurs.

Des interrupteurs de balayage à grande vitesse ainsi que des éléments téléindicateurs à mémoire peuvent être fournis également. Ces éléments permettent la présentation continue de grandes quantités de données sous forme numéral, à n'importe quel point éloigné du voltmètre.

Les deux modèles comportent des blocs incorporés d'alimentation stabilisée et ils sont livrables soit pour montage sur support soit dans un coffret à instruments.

EE 14 751 *pour plus amples renseignements*

EQUIPEMENT DE CONTRÔLE ET D'ENREGISTREMENT DE DONNÉES

Cet equipement a été conçu pour l'emploi avec des installations de traitement continu, afin de contrôler automatiquement les informations sur le rendement de l'installation. Il détecte des conditions défectueuses dès qu'elles se produisent et il enregistre les données de recherches et de développement.

Les signaux d'entrée provenant de thermocouples, transducteurs de pression, débitmètres, etc. sont balayés de façon répétée. La tension représentant chaque paramètre d'entrée est comparée, après amplification, avec une limite préréglée supérieure ou inférieure. Au cas où un paramètre sortirait des limites préréglées, ceci serait instantanément détecté par l'équipement qui donnerait alors un avertissement visuel et sonore. Le paramètre erroné est indiqué avec une précision de trois nombres décimaux (soit une précision de ±0,1% de l'échelle totale) et il est enregistré par un appareil imprimeur en ligne ou à bande perforée qui indique également l'identification du paramètre et le moment des erreurs.

L'équipement peut aussi noter toutes les données à des intervalles de temps prédéterminés, sur demande ou continûment.

EE 14 752 *pour plus amples renseignements*

BURGESS PRODUCTS CO. LTD

Dukes Way, Team Valley, Gateshead 11

MICRO-CONTACTS

(*Illustration à la page 624*)

Le modèle M1 est un nouveau micro-contact subminiature bistable et à quatre circuits. Logé dans un boîtier en plastique phénolique, il est muni d'un bouton de fonctionnement en nylon, et de huit bornes à souder en argent pur. Ses dimensions aproximatives, en dehors du bouton de fonctionnement et des bornes, sont: 25,4 mm de largeur par 22,9 mm de hauteur et 10,2 mm d'épaisseur. Sa force de fonctionnement est de 567 grammes au maximum, son différentiel de mouvement est de 0, 51 mm et ses caractéristiques de durée mécanique sont de 10 millions d'opérations au minimum.

Le but principal du micro-contact 71 est de permettre la commutation bipolaire instantanée dans un espace très restreint. Ceci s'effectue avec les advantages que comporte la rupture brusque, la précision de répétition, les petits mouvements et une longue durée. Le micro-contact M1 peut fermer, ouvrir ou commuter au même moment deux circuits complètement séparés et il peut être employé pour toutes les fonctions bipolaires de précision, de même que pour la commande d'équipements triphasés par la commutation de deux phases.

Il se distingue également par son pouvoir de commander des courants continus d'une valeur considérablement plus élevée que ceux qu'on prête normalement aux interrupteurs subminiature de précision. Lorsqu'il est convenablement relié à la charge ou aux charges, il peut être employé soit comme interrupteur à double écartement, avec une puissance de courant continu de 5A à 30 volts, soit comme interrupteur unipolaire à quadruple écartement avec une puissance de courant continu de IA à 250 volts. Ces caractéristiques sont obtenues sans aucun circuit de protection. Le régime de courant alternatif est de 5A à 250 volts.

EE 14 753 *pour plus amples renseignements*

EKCO ELECTRONICS LTD

Southend-on-Sea, Essex

RADAR MÉTÉOROLOGIQUE POUR AVIONS

(Illustration à la page 625)

La Société Ekco Electronics a exposé un nouveau radar météorologique d'avion dont le poids n'est que de 26 kgs., ce qui est moins que la moitié du poids des appareils précédents de performances semblables. Le nouveau radar ne comporte que trois éléments principaux: l'indicateur, l'émetteur/récepteur et l'antenne, permettant ainsi une écomonie considérable dans les frais d'installation, sans compter l'économie de câblage et de poids du support. L'espace nécessaire et la consommation électrique en sont aussi grandement réduits.

Ce nouveau radar rend tous les services des meilleurs systèmes de radars météorologiques: il comporte une antenne tournante de 60 cm, ainsi qu'un indicateur à haute brillance et sa portée est de 150 milles marins. La réduction de poids a été réalisée par l'utilisation étendue de transistors et par une nouvelle disposition du système, éliminant les éléments conventionnels de synchronisation et de commande.

L'indicateur est conforme aux dimensions-étalon de l'A.R.I.N.C. et comprend toutes les commandes de fonctionnement. Un tube cathodique rectangulaire de 127 mm × 76,2 mm, fonctionnant à 18 kV, donne une image de la distance entière pendant tout le balayage azimutal de 180°. La tension élevée de fonctionnement. alliée au filtre polaroïde replié, assure une image d'une qualité exceptionnelle, même dans des conditions de vive lumière ambiante. Des installations à double indicateur ou à indicateur unique peuvent être livrées au choix.

L'émetteur/récepteur met an œuvre les plus récentes techniques de réception d'hyperfréquences et comprend un isolateur de charge en ferrite. C'est un appareil à bande X, d'une puissance de 15 kW, logé dans un coffret de format réduit, se conformant également aux normes de l'A.R.I.N.C.

Le dispositif de balayage comporte une antenne de 60 cm. à faisceau commutable filiforme/compensé. La méthode de stabilisation à ligne de vue est utilisée dans ce nouveau radar avec un servo-système à vitesse corrigée qui maintient la précision de la stabilisation pendant tout le balayage d'un secteur de 180°.

EE 14 754 *pour plus amples renseignements*

FERRANTI LTD

Hollinwood, Manchester

RADAR D'AVION ET CONTRÔLEUR DE TIR

Le Radar d'Avion et Contrôleur de tir Ferranti, designé sous le nom de "AIRPASS," a été présenté au public pour la première fois à Farnborough.

Le nom d'AIRPASS est formé des mots "Airborne Interception Radar and Pilot's Attack Sight System" (voulant dire radar aéroporté d'interception et système de visée d'attaque). Les deux appareils de base qui constituent cet équipement, soit le radar d'interception et le viseur d'attaque, ont été étudiés de manière à former un ensemble complet permettant au pilote de chasseur de dépister et de détruire sa cible, même si elle lui est invisible et qu'elle vole à une vitesse supersonique.

Ce puissant radar détecteur comprend également un intégrateur de route dont les indications orientent le chasseur vers la meilleure position pour l'attaque finale. Ces indications sont présentées au pilote dans son appareil de visée, ce qui contribue à assurer la précision du tir des armes de bord.

Durant la phase de poursuite, le radar balaye un secteur du ciel sur un angle fort étendu, horizontalement et verticalement des deux côtés de la ligne de vol, et ce sur plusieurs kilomètres à l'avant du chasseur. Lorsque le pilote détecte une cible dans ce secteur, il peut "bloquer" le faisceau du radar sur cette cible. Le radar la suivra alors automatiquement et le calculateur d'approche entrera en fonctionnement.

Aussitôt que le radar a commencé à suivre la cible toutes les indications nécessaires pour compléter l'interception sont présentées au pilote qui pourra effectuer des attaques "à l'aveuglette." S'il y a lieu, cependant, il pourra aussi procéder à l'attaque visuelle à l'aide de cet équipement. Afin de prévenir le risque de collision avec la cible, au cas où l'avion s'en approcherait trop, le radar est muni d'un feu d'alarne pour avertir le pilote d'abandonner l'attaque.

Dans le "English Electric Lightning," le radar est logé, comme appareil unique, à l'intérieur de la partie centrale de la prise d'air du moteur. Quoique ceci ait suscité au début certains problèmes de construction mécanique, il en est résulté, cependant, un radar des plus compacts. Le système à coffret unique présente maints avantages, dont l'économie de poids n'est pas le moindre, découlant de l'élimination des divers coffrets séparés ainsi que de leurs prises et câbles d'interconnexion.

EE 14 755 *pour plus amples renseignements*

THE GENERAL ELECTRIC CO LTD

Magnet House, Kingsway, London W.C.2

TÉLÉCOMMANDE D'AVIONS

(Illustration à la page 625)

Apres avoir réalisé avec plein succès un équipement de 80 MHz pour le contrôle à distance du JINDIVIK et d'autres avions cibles, la Société General Electric a maintenant mis au point un équipement plus compact fonctionnant sur 400 MHz.

L'équipement complet comprend habituellement un récepteur à huit tonalités, un récepteur à deux tonalités, et deux blocs d'alimentation. Le récepteur à huit tonalités sert au contrôle de vol normal, le récepteur à deux tonalités est employé comme dispositif "destructeur" de secours et les deux blocs d'alimentation séparés servent à garantir, qu'en cas de panne d'alimentation de l'avion principal, le dispositif "destructeur" soit toujours disponible.

Le récepteur à huit tonalités comprend un récepteur à gain élevé à modulation de fréquence de 400 MHz, ainsi que des circuits sélecteurs à tonalité acoustique et des relais. Grâce à l'utilisation en paires de huit tonalités acoustiques, 24 combinaisons peuvent être transmises et on peut leur faire fermer 24 circuits de commande différents à la sortie du récepteur.

Le récepteur à deux tonalités est une version simplifiée du récepteur à huit tonalités et il est prévu pour la réception d'une commande unique.

Le bloc d'alimentation fournit la haute tension et la basse tension (totalisant 60 Kw) qui commandent chacun des deux récepteurs. Il est formé d'un régulateur transistorisé et d'un convertisseur et il est actionné à partir de l'alimentation 28 volts de l'avion.

EE 14 756 *pour plus amples renseignements*

INDICATEUR D'ATTITUDE DE PROJECTILES TÉLÉGUIDÉS

(Illustration à la page 625)

L'indicateur d'attitude de projectiles G.E.C. a été réalisé pour les Services des Armes Téléguidées du "Royal Aeronautical Establishment," pour mesurer l'attitude durant le vol en tonneau, le tangage et les mouvements de lacet de la fusée expérimentale "Skylark."

Deux émetteurs à bande X d'ondes entretenues sont installés des deux côtés de la ligne centrale du champ afin d'éclairer la fusée. Une travée spéciale dans la fusée, montée sur un anneau solide, porte les circuits et trois antennes d'hyperfréquences affleurantes, disposés de manière à former deux interféromètres orthogonaux. Le circuit d'hyperfréquences est presque entièrement contenu dans l'anneau solide.

Les angles entre chaque ligne de base d'interféromètre et chaque émetteur sont mesurés par la méthode de comparaison de phases. On peut combiner n'importe quels trois angles de ces quatre angles avec la connaissance de la position de la fusée dans l'espace, pour déterminer le vol en tonneau, le tangage et les mouvements de lacet à quelques minutes d'arc près. L'interféromètre mesure la différence de phase entre les signaux reçus par deux antennes d'une même source.

Le système G.E.C. obvie à la difficulté de maintenir des réponses de phase égales dans deux canaux d'amplification. Une basse fréquence ω est ajoutée, à l'entrée de l'oscillateur local, à un mélangeur, au moyen d'un modulateur à bande latérale unique d'hyperfréquences.

Les différences de phases et de fréquences sont préservées durant le mélange, et l'amplification qui suit, après avoir ajouté les deux signaux, s'effectue dans un amplificateur à fréquence intermédiaire commun. Après la détection et une nouvelle amplification, la basse fréquence résultante est comparée en phase avec une fréquence de référence « dérivant directement du modulateur, afin de produire une tension proportionnelle à la différence de phase existant aux antennes réceptrices. Cette tension est alors renvoyée à terre par télémètre.

EE 14 757 *pour plus amples renseignements*

HONEYWELL CONTROLS LTD

Ruislip Road, East Greenford, Middlesex

ORDINATEUR

(Illustration à la page 626)

L'ordinateur Honeywell peut recevoir un nombre de signaux électriques différents de couples thermoélectriques, de transducteurs de pression, etc. En une seule opération, il mesure et enregistre ces variables et présente l'information sous forme numérique dans des lumières à l'avant de l'équipement, ainsi que sur une feuille d'enregistrement au moyen d'une machine à écrire électrique. En plus de cette simple fonction, il peut aussi, le cas échéant, rendre linéaire des entrées non-linéaires et vérifier des entrées en fonction de repères d'alarme de haut et bas niveau. Le système comporte une méthode fort pratique de changement de niveaux d'alarme, et de durées des transducteurs au moyen de tableaux à pointes.

Le système effectue l'analyse en fonction de niveaux d'alarme hauts et bas à six points par seconde, enregistrant tous les variables sur une machine à écrire électrique au taux d'un point par seconde. La précision de mesure est de $\pm 0{,}1\%$ de lecture soit une position.

Le système est utile lorsqu'il est nécessaire d'enregistrer avec précision et sous une forme arithmétique un grand nombre de variables (40 ou advantage) et lorsqu'il faut vérifier un grand nombre de points en fonction de niveaux d'alarme.

EE 14 758 *pour plus amples renseignements*

ANALYSEUR D'ONDES AUTOMATIQUES

L'analyseur d'ondes automatique est un appareil du type hétérodyne de grande précision, conçu pour produire automatiquement des analyses de fréquences tracées en courbe de signaux de données compliquées, dont les composantes vont de 3 Hz à 2 kHZ. Grâce à sa grande sélectivité, il fournit une méthode pour l'analyse de données audessous de la limite de fréquence basse normale de nombreux analyseurs d'ondes. Les circuits de contrôle automatique rendent possible l'analyse d'importantes quantités de données ne requérant que, de temps en temps, l'attention de l'opérateur.

L'appareil convient à l'emploi avec des cadres à ruban magnétique qui permettent l'analyse de signaux d'une durée de quelques secondes à peine. Pour de telles applications, Honeywell offre des appareils appropriés d'enregistrement sur bande magnétique et de réenregistrement.

On peut analyser l'amplitude par rapport à la fréquence, l'alimentation par rapport à la fréquence et, dans certains cas, on peut effectuer une analyse de l'amplitude par rapport au temps d'un signal ayant des composantes de fréquences à temps variable.

EE 14 759 *pour plus amples renseignements*

MARCONI'S WIRELESS TELEGRAPH CO LTD

Chelmsford, Essex

NAVIGATEUR DOPPLER

(Illustration à la page 626)

Les navigateurs Doppler de la série AD 2300 emploient un système à ondes entretenues et à modulation de fréquences, fonctionnant à une fréquence de 8800 MHz (± 50), à une puissance de 1w. Afin de répondre à un variété étendue d'applications, le navigateur est livrable soit comme appareil sensible comprenant un organe indicateur (donnant les renseignements de base sur la vitesse au sol, l'angle de dérive et la distance parcourue), soit comme appareil sensible avec un intégrateur, au choix de deux modèles pour donner plus de détails. Ces trois variantes sont indiquées ci-après par les suffixes "A," "B" ou "C":

Type AD 2300 A. (Appareil sensible et indicateur) qui donne:

(a) la vitesse au sol et l'angle de dérive
(b) la distance couverte

Type AD 2300 B. (comportant un intégrateur type 4381 au lieu de l'indicateur), et donnant:

(a) la vitesse au sol et l'angle de dérive, comme pour le type AD 2300 A.
(b) le kilométrage à effectuer le long d'un parcours choisi
(c) l'erreur en kilomètres (à gauche ou à droite) à travers le parcours choisi.
(d) le kilométrage total pour arriver à destination.
(e) le plan du prochain parcours de vol.
(f) les indications de conduite à l'indicateur de direction du pilote dans le système de vol (ou au pilote automatique).

Type AD 2300 C. (Comportant un intégrateur type 4382, comme sur la gravure, au lieu de l'indicateur) et donnant:

(a) la vitesse au sol et l'angle de dérive.
(b) la vitesse et la direction du vent.
(c) la position momentanée en latitude et en longitude.
(d) les variations magnétiques.
(e) le kilométrage total pour arriver à destination.
(f) le kilométrage restant sur le parcours de vol en cours d'achèvement.
(g) le plan du prochain parcours de vol.
(h) les indications de conduite à l'indicateur de direction du pilote dans le système de vol (ou au pilote automatique).

Les trois variantes ci-dessus répondent aux besoins de toutes les catégories d'avions. Une présentation sur carte roulante du sillage de l'avion peut être fournie sur demande avec les deux derniers types d'equipements. Une carte de ce genre (fabriquée par la Société Kelvin & Hughes Ltd.) a été exposée au stand conjointement avec le Navigateur Doppler Marconi AD 2300-B.

Le Navigateur AD 2300 entre en fonctionnement au décollage et demeure pleinement effectif jusqu'à un plafond de 1500 mètres et jusqu'à des angles d'inclinaison sur l'aile de 30°. Il garde toute sa précision avec un angle de dérive allant jusqu'a 30° (à babord ou à tribord). La gamme de vitesses au sol varie entre 80 et 900 nœuds.

L'équipement comprend un système de guidage électromécanique qui le rend invulnérable aux pannes, c'est à dire, qu'à condition d'employer un signal Doppler approprié, le traqueur donnera toujours une réponse exacte. En cas de panne du signal, le traqueur retournera automatiquement à la "mémoire," se fixant à nouveau sur le signal dès que le niveau aura été rétabli.

Après des vols d'essai très poussés sur de nombreuses routes aériennes mondiales, il a été établi que même avec des vents de surface de moins de 5 nœuds et une mer calme au-dessous, une ample marge de signal maintient l'équipement en plein fonctionnement, même à des altitudes de plus de 13.500 m.

EE 14 760 *pour plus amples renseignemesnt*

RADIOGONIOMÈTRE À CANAL UNIQUE POUR ONDES ULTRACOURTES

(Illustration à la page 626)

Les radiogoniomètres terrestres pour ondes ultracourtes de la série MARCONI AD 200 sont en service dans le monde entier, plus de deux cents étant en usage quotidien constant. Cette série vient d'être remplacée par la série AD 210. L'indicateur de direction type AD 210 C, montré en fonctionnement à Farnborough, a été spécialement prévu pour les besoins des petits aérodromes, c'est à dire qu'il fournit des renseignements radiogoniométriques de la manière la plus simple et la moins coûteuse. Quoiqu'il s'agisse essentiellement d'un appareil à canal unique, il peut, cependant, faire fonction d'appareil multicanaux, par l'introduction d'éléments locaux reliés en cascade.

On s'est particulièrement attaché dans la construction de cet appareil à réaliser la qualité de rendement la plus élevée. Il fonctionne dans la gamme fréquences de 100 à 156 MHz; n'importe laquelle des cinq fréquences pilotées par quartz peut être choisie par bouton-poussoir sur l'indicateur principal.

Les relèvements se présentent de façon entièrement automatique sur un tableau de paliers de 20,3 cm. de diamètre. Les indicateurs sont normalement montés sur pupitres ou, si l'on préfère, incorporés dans le pupitre de commande principal de l'aérodrome. En plus de l'indicateur principal, on peut utiliser un indicateur asservi, pouvant être situé à 152 m. de celui-là.

Un simple système de télécommande, fonctionnant sur lignes téléphoniques ordinaires, permet le contrôle à distance jusqu'à une distance de 12,8 km.

La précision de relèvement (lors de la

réception d'une onde verticalement polarisée) est extrêmement bonne, l'erreur de fonctionnement étant de l'ordre de 1,5 à 2° sur toute la bande.

EE 14 761 *pour plus amples renseignements*

MICROCELL LTD

56 Kingsway, London W.C.2.

COMMUTATEUR ADDITIF À ACTION RAPIDE

La Société Microcell Electronics, une branche de la Société Microcell Limited, a récemment résolu, en collaboration avec le Ministère des Approvisionnements, le problème du déverrouillage à grande vitesse et sans risques de verrous electromécaniques.

Pour des raisons de praticabilité électrique, les verrous ne peuvent être debloqués simultanément et force fut donc de réaliser un relais secondaire télécommandé à action rapide qui puisse les débloquer à la file, l'intervalle de temps entre chaque déblocage étant de 10 microsecondes.

De rigoureuses restrictions s'imposaient à la réalisation d'un tel appareil: il fallait qu'il soit de dimensions réduites sans être pour cela un petit accessoire sans valeur, mais plutôt un élément capable de subir les essais électroniques courants, ainsi que d'être réparé en cas de panne.

Ce problème a été résolu par la mise au point d'un commutateur électronique entièrement transistorisé, comprenant un nombre d'étages successifs dont chacun déclenche une impulsion de courant de 5 ampères aux verrous électro-mecaniques. Grâce à ce dispositif, aucun étage ne fonctionne avant que le précédent n'ait complété son action. Il a, en outre, le pouvoir de s'arrêter devant un verrou tenace, pendant un bref moment, de manière à ce que des dispositifs de sûreté puissent alors entrer en action et permettre au déblocage de se poursuivre au cas où ce verrou refuserait de se laisser débloquer.

EE 14 762 *pour plus amples renseignements*

MULLARD EQUIPMENT LTD

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

APPAREIL DE PROSPECTION ÉLECTROMAGNÉTIQUE

Un avion "TWIN PIONEER" de la Société Scottish Aviation, spécialement adapté pour les relevés géophysiques aériens, a été muni d'un appareil de prospection électromagnétique fourni par la Société Mullard Equipment Limited.

Cet avion a été construit et équipé pour la Société RIO TINTO, un groupe international de compagnies ayant de gros intérêts miniers, et il a été exposé à Farnborough par la Scottish Aviation.

L'appareil électromagnétique en question a été conçu pour détecter les zones de minerais conducteurs, telles que les dépots massifs de sulfures métalliques, gisant près de la surface de la terre. Il comprend essentiellement un émetteur, un récepteur et un indicateur, ainsi que les organes auxiliaires fournissant les renseignements complémentaires servant à interpréter les données primaires.

La sortie de l'émetteur est appliquée à la production d'un champ magnétique alternatif dans un bobinage accordé, porté par une rallonge du bout de l'aile de tribord de l'avion. Ce champ induit une tension dans le second bobinage, monté pareillement sur l'aile de babord et accouplé au récepteur. Tant le bobinage du récepteur que celui de l'émetteur peuvent être orientés de manière à ce que l'axe de chacun d'eux soit horizontal ou vertical par rapport à la ligne de vol normale de l'avion et en fonction du plan correspondant des gisements de minerai dans la region prospectée.

Lorsque l'avion vole hors de portée de toute influence terrestre, la tension induite dans le bobinage récepteur est contrebalancée par une tension d'ampleur égale mais en opposition de phase, provenant d'un enroulement secondaire sur le bobinage émetteur. Le système est alors équilibré. Lorsque, cependant, l'avion s'approche d'un vaste gisement conducteur, des courants de Foucault sont induits dans le gisement qui modifie le champ magnétique dans le bobinage récepteur. Ceci a pour résultat de déranger l'équilibre des tensions au bobinage récepteur et de produire un signal déséquilibré. Ce dernier est amplifié et transmis au récepteur où, après une nouvelle amplification, il est résolu en composantes en phase et déwattées. Les signaux de sortie qui en résultent sont alors amplifiés pour l'enregistrement sur l'enregistreur multicanaux de bord.

L'enregistrement de signaux de composantes en phase et déwattées, allié au choix d'un émetteur de fréquences appropriées, permet à l'appareil de distinguer entre les gisements de valeur potentielle et ceux de faible conductivité, telles que les couches marécageuses et salines.

En plus des sorties en phase et déwattées, un troisième canal, fonctionnant sur une fréquence voisine, indique le volume de bruit atmosphérique tandis qu'un quatrième canal permet d'enregistrer l'effort dans les supports d'ailes, révélant ainsi la présence de tout coup de vent qui pourrait produire des indications électromagnétiques erronées. Les canaux complémentaires enregistrent les réponses des autres instruments de relèvement dont est muni l'avion, soit un magnétomètre, un radioaltimètre et un scintillomètre pour détecter le rayonnement gamma de sources potentielles d'uranium.

EE 14 763 *pour plus amples renseignements*

REDIFON LTD

Broomhill Road, London S.W.18

ÉMETTEURS HAUTE FRÉQUENCE

(Illustration à la page 627)

La Societe Redifon vient d'annoncer la réalisation d'une gamme entièrement nouvelle d'émetteurs haute-fréquence répondant aux multiples besoins des communications aéronautiques. Cette gamme, qui appartient à la série C. 420, est basée sur un émetteur à deux bandes latérales de 1,5 kW, s'étendant sur la gamme de fréquences de 1,5 à 30 MHz et comprenant les modèles pour A1, A2, A3, A3a, A3b ainsi que F1, F2, F3 et F4 avec entraînement extérieur. Le contrôle peut être manuel, local ou étendu, ou encore entièrement à distance avec servo-accord. La construction à éléments a été adoptée pour ces émetteurs, de sorte qu'un changement dans les besoins de service peut être effectué à n'importe quel moment en ajoutant simplement l'élément de commande approprié. Jusqu'à huit fréquences de point peuvent être obtenues dans un élément haute fréquence ou bien, en cas de nécessité de changement immédiat de fréquence, jusqu'à huit éléments haute fréquence préaccordés peuvent être fournis à partir de circuits d'entraînement commun.

La série des émetteurs G 420 est livrable sous les trois formes de base suivantes:

(a) A commande manuelle pour le fonctionnement à canal unique, nécessitant un réglage manuel pour les changements de fréquence. Un dispositif de télécommande peut être fourni, mais non pour le changement de fréquence.
    On trouvera pratique d'ajouter des éléments à un émetteur existant, pour le convertir de la commande manuelle à la servo-commande.
(b) A servo-commande, pour le fonctionnement à canal unique, et permettant de choisir à distance jusqu'à huit fréquences de point. Un dispositif de télécommande totale peut être fourni.
(c) A multi-baies, pour le fonctionnement à canal unique, et permettant de choisir très rapidement à distance jusqu'à huit éléments haute fréquence. Ce modèle est particulièrement indiqué pour les applications ou la durée des changements de fréquence doit être réduite au minimum et ou un grand nombre de changements est nécessaire journellement.

EE 14 764 *pour plus amples renseignements*

RADIOTÉLÉPHONE POUR HYPERFRÉQUENCIES

(Illustration à la page 628)

Le radiotéléphone modulé en amplitude pour hyperfréquences, type GR 318, est un appareil de conception avancée pour communications de parole entre vehicules mobiles et une station principale appropriée. C'est un organe compact de 4 à 5W, à faible comsommation électrique et des plus adaptables au montage. Ces caractéristiques le rendent particulièrement adéquat pour les services mobile de terre d'aérodromes.

L'utilisation judicieuse des transistors contribue à la sûreté du fonctionnement et à une consommation économique de batterie. Un convertisseur à transistors fort efficace effectue la conversion de l'énergie en haute tension. Quant au modulateur, également transistorisé, il remplit la double fonction d'étage modulateur et d'étage de sortie de récepteur.

Grâce à son encombrement réduit, le radiotéléphone GR 318 peut être convenablement installé dans le tablier ou le haut de la cabine du conducteur du camion.

Le choix des modèles livrables s'étend à sept bandes de fréquences dans la gamme de 72 à 175 MHz.

EE 14 765 *pour plus amples renseignements*

ROYSTON INSTRUMENTS LTD

Canada Road, Oyster Lane, Byfleet, Surrey

ENREGISTREUR DE DONNÉES DE VOL

(Illustration à la page 628)

L'appareil enregistreur de données de vol MIDAS, à bande magnétique, est le fruit de l'expérience considérable acquise par la Société Royston Instruments Ltd. dans la construction d'équipements d'enregistrement et de traitement de données pour les travaux de recherches et d'essais d'avions. Son rôle est de fournir un enregistrement complet, précis et permanent, de chaque paramètre de fonctionnement d'importance, du décollage à l'atterrissage, pendant toute la durée de l'avion. Grâce à un système accéléré de dépouillement de données, l'enregistrement est traduit sous une forme compacte et utilisable.

L'enregistreur à bande magnétique est un appareil spécial à vitesse réduite, basé sur des principes longuement éprouvés dans le système MIDAS, depuis de nombreuses années. Il comprend, au total, sept canaux de signaux fonctionnant en parallèle, dont l'un est réservé à l'heure moyenne de Greenwich, à la date, au numéro de vol, etc. La capacité d'enregistrement des six autres canaux est augmentée par un commutateur à division de temps qui choisit un grand nombre de paramètres en succession. En fait, il y a six ensembles de commutation, actionnés de façon continue par un moteur commun, chacun d'eux alimentant le canal d'enregistrement correspondant et combinant ainsi un nombre de différents paramètres. L'enregistreur est un organe entièrement classique et il est, sauf en cas de spécification contraire, identique pour tous les avions, quelle que soit l'installation et les paramètres à enregistrer.

Quoique de conception et de construction courantes, le commutateur à division de temps est, toutefois, relié aux connexions d'entrée multivoies de façon à s'adapter à l'instrumentation particulière de l'avion. Le commutateur à division de temps peut être monté en un endroit adéquat au centre du câblage du transducteur (c'est à dire, probablement, dans la cabine) et il peut être relié physiquement à l'arrière de l'enregistreur par des prises et des fiches du type " Unitor." Si, d'autre part, l'usager désire un dispositif d'éjection de secours, l'enregistreur seul peut être fixé dans la queue et un câble suffit à relier les deux éléments.

Le commutateur à division de temps détermine le nombre total de paramètres à essayer, ainsi que le taux d'échantillonnage. Le commutateur multivoies effectue une rotation toutes les cinq secondes, de sorte que chaque paramètre est essayé à la cadence minimum d'une fois par cinq secondes. Le taux d'échantillonnage peut être doublé ou triplé par la mise en parallèle des sections du commutateur à de faibles intervalles, de manière à ce que ces signaux soit prélevés deux ou trois fois par révolution du commutateur. Donc, n'importe quel nombre de canaux et de combinaisons de taux de prise d'échantillons peuvent être obtenus dans un minimum de 300 canaux, chacun d'eux étant ensuite essayé toutes les cinq secondes.

Les bobines de ruban magnétique qui forment l'organe d'enregistrement sont préchargées dans une cassette qui peut être insérée ou retirée par un personnel non-qualifié. Ce facteur, joint à la capacité d'enregistrement de 160 heures, signifie qu'il n'existe aucune difficulté à garder un enregistrement complet pendant les circuits de vol les plus longs, même lorsque la bande magnétique n'est remplacée qu'au retour à la base principale. L'enregistreur a été étudié pour une durée infinie et il garantit un niveau de précision très élevé. Il est de format réduit (IATR) et pèse moins de 18,15 Kg.

Un appareil de réenregistrement et de traitement de données peut être livré à l'usage de cet enregistreur.

EE 14 766 *pour plus amples renseignements*

THERMIONIC PRODUCTS (ELECTRONICS) LTD

Hythe, Southampton

ENREGISTREURS À BANDE MAGNÉTIQUE MULTI-CANAUX

(Illustration à la page 628)

Ces enregistreurs sont formés d'un nombre d'éléments séparés et, en choisissant les éléments qui conviennent le mieux, on peut donc fournir un enregistreur pour la plupart des applications.

L'enregistreur figurant dans notre gravure est le modèle IA/IA, prévu pour l'enregistrement simultané de n'importe quel nombre de signaux de parole ou codes, jusqu'à un maximum de 20 signaux. Ce modèle comprend normalement un canal code de temps et un canal de réserve automatique.

Il est également doté d'un élément à fréquence vocale arrêt/marche, permettant un fonctionnement maximum de 24 heures sans qu'on s'en préoccupe et pouvant être prolongé en proportion du temps total d'absence de signaux.

Tous les canaux sont pourvus de contrôle continu, le canal de réserve étant mis en fonctionnement à la première panne, sans que la continuité de l'enregistrement en soit affectée. Toutes les facilités de réenregistrement s'obtiennent dans l'emplacement "de secours."

L'appareil est entièrement transistorisé et, lorsque cela est faisable, des circuits imprimés sont utilisés pour en assurer la sécurité de fonctionnement. L'élément à minuterie et de codage est à commande piézoélectrique, pour plus de précision, et il a quatre étages interchangeables afin de simplifier le dépannage, si nécessaire.

EE 14 767 *pour plus amples renseignements*

THORN ELECTRICAL INDUSTRIES LTD

105-109 Judd Street, London W.C.1

INDICATEUR ÉLECTROLUMINESCENT

(Illustration à la page 629)

L'indicateur électroluminescent à 100 points que construit maintenant la section des accessoires et connecteurs d'avions de la Société Thorn Electrical Industries Limited, applique un nouveau principe aux systèmes à carte de lecture arithmétique et de marche.

Cet accessoire ne mesure que 6,73 mm. de profondeur et n'occupe qu'une fraction de l'espace nécessaire aux indicateurs à lampes. Il n'utilise pas de filaments et ne produit donc virtuellement aucune chaleur. Il comprend 100 points rectangulaires phosphorescents, serrés entre deux surfaces conductrices d'électricité. En appliquant une tension alternative à ces points, on provoque leur luminescence. Les changements dans le dessin formé par les points éclairés permettent à l'indicateur de lire n'importe quel caractère ou chiffre, ou de fournir une indication de marche précise.

L'indice de brillance de l'indicateur est de 15 lambert-pieds. Pour obtenir l'intensité requise, il est recommandé d'employer une alimentation haute fréquence d'une tension d'environ 350 volts.

EE 14 768 *pour plus amples renseignements*

ULTRA ELECTRIC LTD

Western Avenue, London W.3

INDICATEUR DE LIEU D'ATTERRISSAGE BRUTAL

Il s'agit ici d'un dérivé intéressant de l'équipement de recherche et de sauvetage SARAH. Il permet de repérer rapidement et avec certitude l'emplacement de survivants ou de débris, même dans le brouillard ou dans des conditions d'isolement en mer ou sur terre.

L'indicateur en question, désigné sous l'abréviation C.P.I. (Crash Position Indicator), est un émetteur-balise construit sur une antenne à plaque mince. Il est noyé dans une matière plastique cellulaire formant un plan aérodynamique recouvert d'une pellicule feuilletée en nylon. Le C.P.I. est incorporé à l'avion de manière à faire partie intégrante du fuselage.

Un mécanisme sensible de déclenchement, actionné par fils métalliques tendus le long de la structure de l'avion, détecte le premier indice de collision ou d'anomalie dans la structure. En moins de 100 m.sec., c'est à dire en moins de temps qu'il ne faut pour qu'un avion moderne à grande vitesse soit complètement détruit, le C.P.I. sera projeté au loin de ce dernier afin de tomber à distance sûre de l'épave et commencer aussitôt des émissions de balise. Dans des conditions normales et favorables, ces émissions peuvent être détectées dans un rayon de 48 km. Si les conditions sont défavorables, et que, par exemple, le C.P.I. se trouve enterré dans la neige, dans une vallée profonde, on peut quand même compter sur 8 à 18 km. de portée.

L'émetteur fonctionne même submergé dans l'eau et sa durée ininterrompue de fonctionnement dépasse 100 heures.

Le C.P.I. atterrit sans risques sur glace ou rocher à 64 km./h., et à plusieurs fois cette vitesse sur la neige ou sur terrain mou. Sa chute à terre est d'environ 42 km./h. et il flotte hors de l'eau dans une proportion de 80%.

Le signal du C.P.I. s'accorde avec l'équipement de recherche SARAH, qui est l'appareil courant des forces armées du Commonwealth britannique et de la plupart des états membres de l'O.T.A.N. La zone d'action du SARAH s'étend à l'Atlantique Nord, à la Mer du Nord, à la Mer Baltique, a la Mer des Caraïbes, à la Méditerranée, à l'Océan Indien, au Golfe du Bengale, à l'Australasie, à la Mer de Chine du Sud et au Pacifique Occidental.

EE 14 769 *pour plus amples renseignements*

Analyseur de l'état du moteur

(Illustration à la page 629)

Il s'agit ici d'un nouvel instrument donnant des indications sur la température et les vibrations du moteur, présentées sous une forme intelligible au coup d'œil. C'est un système des plus souples, pouvant être employé pour les vérifications de pré-vol, le contrôle de vol et l'entretien courant.

Des niveaux de température allant jusqu'à 40 thermocouples sont indiqués sur le tube cathodique supérieur de l'indicateur. Ces indications de température sont divisées en quatre groupes, c'est à dire un groupe par moteur. Au moyen d'une commande du panneau avant, les dix points de température de n'importe lequel des moteurs, peuvent être presentés sur toute la largeur de l'écran. Deux lignes droites horizontales sont tracées à travers l'indicateur de température: l'une d'elles est la ligne de temperature de base, l'autre est la ligne d'étalonnage. Le côté gauche de l'indicateur porte une échelle avec des marqueurs à intervalles de 25°C. La sensibilité est de 300°C par pouce et l'instrument est réglé avant l'installation de façon à ce que la température nominale de service du moteur soit au milieu de l'écran.

La ligne de base demeure fixe dans l'une des quatre positions, choisie par une commande du panneau avant. Les quatre positions sont "Décollage," "Montée," "Croisière" et "Descente." Lorsque le moteur fonctionne correctement, la ligne de base passe par la moyenne des indications de température, de sorte que tout glissement de température, à n'importe lequel des points de contrôle, est immédiatement apparent. La position de la ligne pour chacun des quatre états est préréglée pour le type particulier du moteur, avant l'installation de l'analyseur.

La ligne d'étalonnage de température peut être déplacée vers le haut ou vers le bas par une commande du panneau avant: on peut lire la température de n'importe quel point à 0,5% près de l'échelle totale, en ajustant la ligne d'étalonnage jusqu'à ce qu'elle coïncide avec le point, et en lisant l'échelle sur la commande d'ajustage. Si on n'a besoin que d'une lecture approximative, la temperature peut être calculée simplement de l'échelle du tube cathodique.

Les indications des deux transducteurs de vibrations se trouvant sur chacun des moteurs sont présentées de façon continue sur le tube cathodique inférieur de l'indicateur. Une ligne de danger de vibrations est tracée à travers l'indicateur. La hauteur de cette ligne est ajustée selon le type particulier de moteur auquel elle se rapporte, avant l'installation de l'analyseur, et elle est réglée de manière à ce que les indications restent au dessous de la ligne si les moteurs fonctionnement correctement. Les points de l'indicateur correspondent à des amplitudes de vibration de première grandeur. Les vibrations à des fréquences autres que celles correspondant à la vitesse de rotation du moteur ne sont pas indiquées.

EE 14 770 *pour plus amples renseignements*

WAYNE KERR LABORATORIES LTD

Roebuck Road. Chessington. Surrey

Mesureur de distance

(Illustration à la page 629)

Ce appareil, qui fonctionne conjointement avec des palpeurs non-contacteurs, constitue une méthode fondamentalement linéaire pour mesurer des distances de 50 micropouces à un demi pouce. La distance est lue directement sur un appareil de mesure à cadre mobile, dont la portée totale dépend du palpeur utilisé. Il existe quatre palpeurs, pour des portées totales de 1, 10, 100 et 500 millièmes de pouce respectivement.

La précision de lecture est de ±2% de déviation totale et la discrimination est supérieure à 0·5% de la déviation totale.

La longueur du câble du palpeur est de 3m ce qui permet de placer l'instrument à bonne distance de la structure d'essai. Une ouverture sur le panneau avant a été prévue pour brancher des enregistreurs ou des systèmes de servo-commande.

L'instrument représente un moyen simple et rapide de mesurer la distance dans les procédés de meulage de précision et il peut être adapté facilement à des applications telles que l'équilibrage dynamique de pales d'hélices d'hélicoptère et la mesure de la verticalité des barres de contrôle dans les centrales nucléaires.

On obtient une indication directe de la dilatation sans imposer une charge quelconque à l'objet sous observation et les changements de dimensions peuvent être contrôlés ou enregistrés continûment en fonction du temps ou de la température.

Parmi les autres applications, il faut inclure toutes les mesures de distance jusqu'à 12,7 mm, lorsque la lecture directe et le contrôle continu sont requis, ainsi que les vérifications courantes de pièces et les mesures sur place durant les opérations d'usinage. L'instrument peut être réglé pour le rejet automatique de pièces hors tolérance.

EE 14 771 *pour plus amples renseignements*

Résumés des Principaux Articles

Un appareil de mesure de distribution d'amplitude par M. Drayson

Résumé de l'article aux pages 578 à 584

Cet article souligne l'importance des mesures de distribution d'amplitude de signaux, donnant un bref aperçu de l'expression mathématique de la distribution d'amplitude en ce qui a trait particulièrement aux approximations qu'entraîne tout système de mesure pratique. Un nouveau genre de dispositif d'échantillonnage de formes d'ondes a été produit qui permet des mesures jusqu'à une résolution d'au moins 1%. Après avoir dépeint ce dispositif, l'auteur donne une description détaillée des circuits qui s'y rapportent. Ces derniers forment un appareil capable de mesurer des tracés de distribution dans des gammes de 0,3 à 30 volts d'amplitude, et de 0,1 à 10 MHz de fréquence, indiquant également la distribution d'amplitude de formes d'ondes d'une fréquence minimum de plus 1 Hz. Il reproduit, enfin, des photographies de tracés de distribution sur un tube cathodique.

L'étude de convertisseurs push-pull de courant continu à transistors par W. L. Stephenson, L. P. Morgan et T. H. Brown

Résumé de l'article aux pages 585 à 589

Il existe de nombreuses méthodes au moyen desquelles on peut convertir le courant continu d'une basse à une haute tension en utilisant quelque dispositif oscillant, mais l'une des méthodes les plus efficaces emploie un oscillateur à ondes carrées à transistors, actionné par un transformateur à saturation. Les formules d'étude de ce système s'obtiennent en fonction des paramètres de fonctionnement.

Un aperçu est donné des principales difficultés et limites d'ordre pratique, mais elles ne sont pas examinées en détail car des solutions individuelles sont habituellement nécessaires.

Quelques aspects du dessin logique et des circuits d'un calculateur arithmétique de champ par I. F. Brown et B. Meltzer

Résumé de l'article aux pages 590 à 592

Les principes d'un nouveau type de calculateur arithmétique pour la solution de problèmes de champ sont décrits dans cet article. En rendant le calcul simultané à tous les points du réseau, ce calculateur raccourcit fort sensiblement la durée du calcul. L'arithmétique unitaire (abaque), plutôt que binaire ou décimale, est employée. Le calcul s'effectue par éléments de réseau, chacun d'eux étant relié à un point de réseau du champ. Les auteurs décrivent la synthèse des éléments du réseau à partir d'éléments de base plus simples et présentent le dessin expérimental d'un élément de base pour problèmes de potentiel et autres. Un bref aperçu de certains des nombreux problèmes ardus que résoud le calculateur conclut l'article.

Diagrammes pour trouver les paramètres-r de transistors à partir de paramètres-h par G. W. E. Stark, B.A., A.M.I.E.E.

Résumé de l'article aux pages 592 à 593

L'article traite de diagrammes simplifiant le processus d'induction des valeurs de paramètres-r d'un transistor, de ses paramètres-h de base à la terre mesurés.

Etude de certains aspects des circuits de numération à diode par W. Alexander et H. V. Bell

Résumé de l'article aux pages 594 à 598

Les auteurs définissent la numération et examinent certains travaux réalisés à ce jour sur les numérateurs à diode. Après avoir comparé trois circuits, tant du point de vue théorique que par la mesure, ils présentent les résultats obtenus. Une application possible de ces circuits est décrite en conclusion.

Un simple tachymètre électronique pour automobiles par M. J. Wright

Résumé de l'article aux pages 599 à 602

On obtient du système d'allumage d'une automobile un signal électrique ayant une fréquence fondamentale proportionnelle à la vitesse du moteur. Un nouveau circuit, qui unit les fonctions de filtrage et d'écrêtage, prévient les oscillations parasites et fournit une onde carrée d'amplitude fixe à un circuit de mesure de fréquence de pompe à diode, calibré directement en fonction de la vitesse du moteur.

Le circuit complet est simple, n'a pas d'éléments actifs et il est actionné par la batterie de l'automobile. L'étalonnage n'est que faiblement affecté par les changements de tension de la batterie ou de la température ambiante.

Un simple contrôleur de transistors par G. G. Yates

Résumé de l'article aux pages 602 à 603

L'auteur décrit un instrument simple et peu coûteux pour mesurer les paramètres les plus utiles (I_{co}', β, h_{11}, f_a) *des transistors à basse et haute fréquence de faible puissance. On détermine* I_{co}' *et* β *par une méthode à courant continu, sans qu'il soit besoin d'appareil auxiliaire. Pour déterminer* h_{11} *et* f_a, *il faut en plus un générateur de signaux à gamme de fréquences de moins de 1 kHz à environ 200 kHz, ainsi qu'un voltmètre électronique ou un oscilloscope. Ceci n'est pas considéré comme un désavantage vu que même un laboratoire modestement équipé devrait pouvoir posséder ces instruments.*

Un circuit démultiplicateur à transistors, à temps de résolution réduit par B. Collinge

Résumé de l'article aux pages 604 à 605

Les circuits démultiplicateurs de rapport 2/1, à action rapide, sont nécessaires à diverses applications. L'auteur examine les avantages que comporte le couplage de deux circuits démultiplicateurs de rapport 2/1 au moyen d'un circuit à porte.

Il présente un schéma des circuits et décrit l'élément complet comprenant un circuit à porte et un circuit bistable.

Un indicateur de signes et de grandeurs de chiffres à l'usage des compteurs inversibles par D. L. A. Barber

Résumé de l'article aux pages 605 à 606

L'auteur traite d'un circuit pouvant être utilisé avec n'importe quel compteur inversible afin de fournir une indication des signes et de l'ordre de grandeur de chiffres.

Solutions approximatives de guides d'ondes pour diodes en circuits d'impulsions par D. C. Dillistone

Résumé de l'article aux pages 607 à 610

En prenant une représentation approximative d'une diode, on obtient des solutions de formes d'ondes pour un groupe de circuits simples auxquels on applique une forme d'onde rectangulaire ou (dans un cas particulier) triangulaire. Certaines applications des résultats sont discutées par l'auteur en termes généraux.

Diviseur de fréquences d'oscillateur surcouplé de transistor déclenché par générateur de formes d'ondes à plots par F. Butler

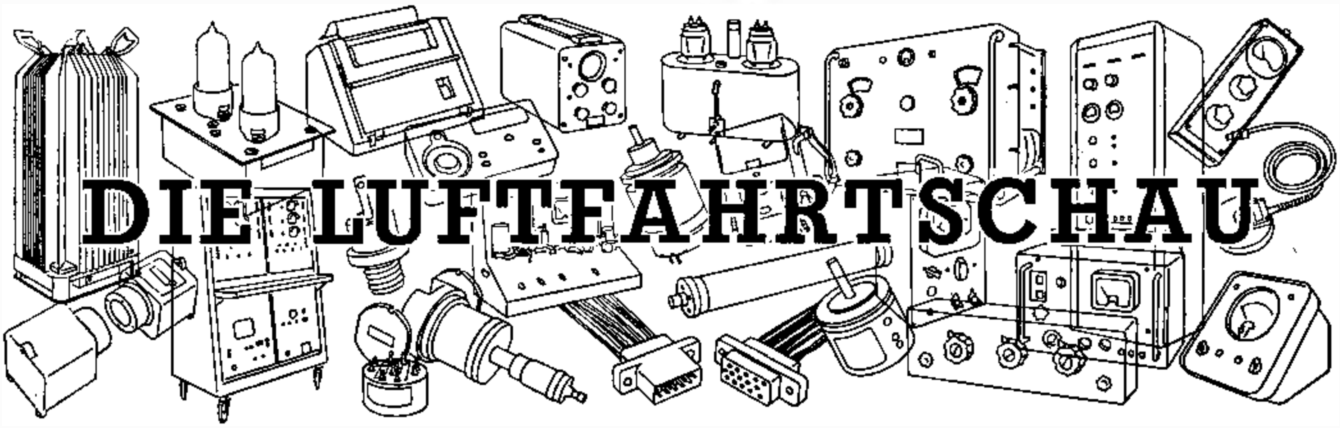
Résumé de l'article aux pages 611 à 612

Ce diviseur de fréquence comprend un générateur de formes d'ondes "à escalier", prévu pour produire une forme d'ondes multi-plots dont tous les accroissements de tension sont d'égale amplitude. Après un nombre prescrit de plots, on emploie la sortie pour déclencher un oscillateur de blocage donnant une sortie d'impulsions périodiques à un sous-multiple fixe de la fréquence des plots d'entrée. Le diviseur de fréquence fonctionne à partir d'une source donnant une forme d'onde à sortie sinusoïdale carrée et limitée, afin de rendre le mode de fonctionnement indépendant de l'amplitude du signal d'entrée. Un amplificateur de sortie suit l'oscillateur de blocage. L'appareil est entièrement muni de transistors et de diodes de modèles courants.

Détecteur sensible aux phases à diodes, avec charge par R. Chidambaram et S. Krishnan

Résumé de l'article aux pages 613 à 616

Les auteurs s'attachent à un examen théorique du fonctionnement d'un détecteur push-pull à diodes sensibles aux phases, avec une charge. Il s'avère que les rapports de transfert de deux diodes varient considérablement avec le signal. La non-linéarité de la sortie, due à ces variations, est évaluée et un tableau est donné qui permet de juger immédiatement de l'applicabilité d'un détecteur donné. Des expériences confirment quantitativement les résultats théoriques.



DIE LUFTFAHRTSCHAU

Beschreibung einiger auf der Luftfahrtschau des Vereins Britischer Flugzeugkonstrukteure (SBAC) gezeigten elektronischen Geräte, die nach Angaben der Hersteller zusammengestellt wurde.

(Übersetzung der Seiten 624 bis 629)

BLACKBURN ELECTRONICS LTD

Brough, Yorkshire

DREI- UND VIERSTELLIGES VOLTMETER

(Abbildung Seite 624)

Diese elektronischen Zahlenvoltmeter wurden für Anwendungsgebiete entwickelt, die genaue Messungen von Gleichspannungen erfordern. Dazu gehören Rechnereingaben und -ausgaben, Bezugsquellenmessungen, Dehnungsmessungen, laufende Prüfung elektronischer und elektrischer Elemente, Datenverarbeitung und -anzeige (z.B. Analog/Digital Umwandlung).

Die Schaltungen dieser Instrumente garantieren hohe Betriebssicherheit. Die Genauigkeit für jeden der drei Bereiche ist ±0,1% für Modell B.I.E. 2113 und ±0,5% für Modell B.I.E. 2114. Für Modell B.I.E. 2113 sind die Bereiche 0 .. ±9,60, ±9,60 ... ±96,0, ±96,0 ... ±999 V Gleichstrom und für Modell B.I.E. 2114 0...±9,600, ±9,600...±96,00, ±96,00 ... ±999,9 V Gleichstrom. Für beide Instrumente ist die Eingangsimpedanz 50 MΩ für den niedrigeren Bereich und 5 MΩ für die höheren Bereiche. Bereichschaltung und Vorzeichenwahl sind automatisch und erscheinen auf der Sichtanzeige, für die Neonzahlröhren verwandt werden. Die Anzeigezeiten sind 13 ms für das dreistellige Modell B.I.E. 2113 und 103 ms für das vierstellige Modell B.I.E. 2114.

Im Betrieb vergleicht das Gerät die Eingangsspannung mit einer ansteigenden Spannung, deren Verlauf innerhalb ±0,01% geradlinig ist. Zwischen dem Zeitpunkt, in dem die ansteigende Spannung die Nullinie kreuzt und demjenigen, in dem sie die Amplitude der Eingangsspannung erreicht, werden die Schwingungen eines quarzgesteuerten Oszillators gezählt. Diese Zählung ist der Digitalwert der Spannung, der direkt durch die Neonzahlenröhren angezeigt wird. Eine Eichung mit Bezug auf eine Weston-Normalzelle ist in beiden Modellen eingebaut. 80 V negativgehende Impulse der Instrumente können Drucker, elektrische Schreibmaschinen, Lochstreifenstanzer oder Magnetbandgeräte sowie Fernanzeigen treiben. Schnellüberstreichschalter und Fernanzeigen mit Speicher sind lieferbar. Diese Geräte können grosse Datenmengen fortlaufend in Zahlendarstellung an irgendwelchen, vom Voltmeter entfernten Plätzen sichtbar machen.

Beide Instrumente haben eingebaute, stabilisierte Netzgeräte und können entweder für Gestellmontage oder im Instrumentgehäuse geliefert werden.

EE 14 751 *für weitere Einzelheiten*

DATENÜBERWACHUNGS- UND AUFZEICHNUNGSGERÄT

Das Gerät wurde für Anlagen mit fortlaufenden Herstellungsverfahren entwickelt, deren Leistung es automatisch überwacht. Es entdeckt Misstände, wenn sie auftreten, oder zeichnet Daten für Forschungs- oder Entwicklungszwecke auf.

Die Eingangssignale entstehen in Thermoelementen, Druckumwandlern, Strömungsuhren usw. und werden wiederholt abgetastet. Nach Verstärkung wird jede einen Eingang darstellende Spannung mit eingestellten oberen und unteren Grenzwerten verglichen, der Zustand festgestellt und ein hörbarer oder sichtbarer Alarm gegeben. Der fehlerhafte Parameter wird mit einer Präzision von drei Dezimalstellen in Zahlen umgewandelt (±0,1% Genauigkeit der Vollskala) und durch Drucker oder Streifenstanzer aufgezeichnet, wobei gleichzeitig der Parameter und Zeitpunkt des Fehlers angegeben wird.

Das Gerät kann auch alle Daten in vorherbestimmten Zeitintervallen, oder auf Befehl, oder dauernd aufzeichnen.

EE 14 752 *für weitere Einzelheiten*

BURGESS PRODUCTS CO. LTD

Dukes Way, Team Valley, Gateshead 11

MIKROSCHALTER

(Abbildung Seite 624)

Der M 1 ist ein neuer, vierkreisiger, bistabiler Kleinst-Mikroschalter. Das Modell hat ein Phenolharz-Gehäuse mit Nylon-Schaltknopf und acht reinsilbernen Lötanschlüssen. Einschliesslich Schaltknopf und Anschlüssen sind die Abmessungen ungefähr 25,4 mm breit, 22,9 mm hoch und 10,2 mm tief. Die Höchstschaltkraft beträgt 567 g, das Bewegungsdifferential 0,51 mm und die mechanische Lebensdauer mindestens 10 Millionen Schaltungen.

Der M 1-Schalter wird hauptsächlich für räumlich sehr beschränkte doppelpolige Momentanumschaltungen eingesetzt. Das Modell führt diese Aufgaben mit Schnellstwirkung, kürzestem Weg und langer Lebensdauer aus. Der Schalter kann zwei vollständig getrennte Kreise zum selben Zeitpunkt ein-, aus- oder umschalten. Er entspricht allen doppelpoligen Präzisionsanforderungen, sogar der Steuerung von Drehstromgeräten durch Zweiphasenschalten.

Der M 1 kann weiterhin bedeutend höhere Gleichströme schalten, als die bisher für Präzisions-Kleinstschalter bekannten Werte annehmen lassen. Bei geeigneter Verbindung zum Verbraucher oder den Verbrauchern kann er mit entweder 5 A 30 V Gleichstrom als Doppelspalt-Schalter, oder mit 1 A 250 V Gleichstrom als einpoliger Schalter mit Vierfach-Spalt belastet werden. Diese Nennwerte werden ohne Schutzkreise erreicht. Für Wechselstrom sind die Nennwerte 1 A bei 250 V.

EE 14 753 *für weitere Einzelheiten*

EKCO ELECTRONICS LTD

Southend-on-Sea, Essex

BORDGERÄT FÜR WETTER-RADAR

(Abbildung Seite 625)

Ekco Electronics stellten ein neues Transistor-Bordgerät für Wetter-Radar aus, das nur 24,5 kg und damit die Hälfte früherer Geräte gleicher Leistung wiegt. Da das Gerät nur aus drei Hauptteilen besteht - Anzeige, Sender-Empfänger und Abtaster - wird wesentlich an Anlagekosten, sowie Kabel und Gestellgewicht gespart. Der beanspruchte Platz und Stromverbrauch sind viel geringer.

Dieses neue Radar weist alle Einrichtungen der besten Wetter-Radarsysteme auf, einschliesslich einer stabilisierten 60,96 cm Antenne, 240 km Reichweite und sehr heller Sichtanzeige. Die Gewichtsherabsetzung wurde durch weitgehende Verwendung von Transistoren und eine Neuanordnung des Systems erreicht, die die üblichen Synchronisierungs- und Steuergeräte überflüssig macht.

Das Anzeigegerät entspricht in seinen Abmessungen den ARINC-Normen und enthält alle Bedienungselemente. Eine rechteckige Bildröhre (127 mm × 76,2 mm) arbeitet mit 18 kV und gibt eine Sichtanzeige für die volle Reichweite und den überstrichenen Azimutwinkel von 180°. Die hohe Betriebsspannung ergibt zusammen mit einem gewinkelten Polarisationsfilter selbst in heller Umgebung eine ungewöhnliche Anzeigesichtbarkeit. Anlagen mit einfacher oder doppelter Anzeige sind lieferbar.

Der Sender-Empfänger ist entsprechend der neuesten Mikrowellentechnik gebaut und Ferrit-Belastungsisolation ist vorgesehen. Das Gerät arbeitet im X-Band mit einer Leistung von 15 kW und ist, entsprechend ARINC-Normen, in einem halben, kurzen ATR Gehäuse untergebracht.

Das normale Abtastgerät hat eine 60,96 mm Antenne und ist von scharfer Bündelung auf cosec^2 Strahl umschaltbar. Die Visierlinie ist stabilisiert, und diese Stabilisation wird durch ein geschwindigkeitskorrigiertes Servo-System über den ganzen Abtastsektor von 180° genau eingehalten.

EE 14 754 *für weitere Einzelheiten*

FERRANTI LTD

Hollinwood, Manchester

BORDRADAR- UND FEUERLEITGERÄT

Ferranti zeigte in Farnborough zum ersten Mal öffentlich das als AIRPASS bekannte Bordradar- und Feuerleitgerät.

AIRPASS bedeutet „Bordabfangradar und Angriffsvisier des Piloten". Abfangradar und Angriffsvisier sind die beiden Hauptbausteine des Gesamtgerätes, das entwickelt wurde, um dem Jagdflieger zu ermöglichen, das Ziel zu finden und zu zerstören, trotzdem er es nicht sehen kann und es mit Überschallgeschwindigkeit fliegt.

Das Radar ist ein leistungsfähiges Such- und Verfolgungsgerät mit einem Rechner, der den Kurs festlegt, dem der Jäger folgen muss, um eine für den Endangriff geeignete Position zu erreichen. Diese Information wird dem Piloten in seinem Visier angezeigt, wodurch gleichzeitig alle Waffen des Jägers richtig avisiert werden.

Während der Suchzeit überstreicht das Radar in einem Himmelsabschnitt einen weiten Horizontal- und Vertikalwinkel auf jeder Seite des Flugzeuges und viele Kilometer voraus. Wenn der Pilot in diesem Abschnitt ein Ziel findet, kann er auf Zielverfolgung umschalten. Das Radar wird dann automatisch das Ziel verfolgen und der Anflugrechner fängt an zu arbeiten.

Sowie das Radar die Zielverfolgung übernimmt, werden dem Piloten alle zum erfolgreichen Abfangen notwendigen Daten angezeigt, so dass er mit Hilfe dieses Gerätes einen Blindangriff ausführen kann. Jedoch ist auch ein Sichtangriff möglich, wenn es wünchenswert erscheint. Als Vorsichtsmassnahme ist ein Warnlicht vorgesehen, das anzeigt, wann der Pilot den Angriff abbrechen muss, um nicht mit dem Ziel zusammenzustossen.

Im „Lightning" der English Electric ist das Radar als Einzelgerät im Einlaufdiffusor des Triebwerkes untergebracht. Trotzdem im Anfang einige mechanische Konstruktionsschwierigkeiten enstanden, wurde dadurch ein sehr kompaktes Gerät geschaffen. Der Zusammenbau in ein Gerät hat viele andere Vorteile, unter denen die Gewichtsersparnis durch Fortfall der verschiedenen Gehäuse, Stecker und Kabel eine wichtige Rolle spielt.

EE 14 755 *für weitere Einzelheiten*

THE GENERAL ELECTRIC CO LTD

Magnet House, Kingsway, London W.C.2

FLUGZEUG-FERNLENKUNG

(Abbildung Seite 625)

Nach dem Erfolg des 80 MHz Fernlenkgerätes für das „Jindivik" und andere Zielflugzeuge hat die GEC nunmehr ein sehr kompaktes Gerät für 400 MHz Betrieb entwickelt.

Die Gesamtanlage besteht aus einem Achtton-Empfänger, einem Zweiton-Empfänger und zwei Netzgeräten. Der Achtton-Empfänger dient der normalen Fluglenkung, der Zweiton-Empfänger steuert die „Zerstör"-Einrichtung, und die zwei Netzgeräte stellen sicher, dass die „Zerstör"-Einrichtung noch zur Verfügung steht, wenn die Flugzeugversorgung ausfällt.

Der Achtton-Empfänger besteht aus einem frequenzmodulierten 400 MHz Empfänger mit hoher Verstärkung, niederfrequenten Tonwählerkreisen und -relais. Als Paare eingesetzt, können die acht Töne 24 Kombinationen übermitteln und damit 24 verschiedene Befehlskreise am Empfängerausgang steuern.

Der Zweiton-Empfänger ist eine vereinfachte Ausführung des Achtton-Empfängers, der nur für den Empfang eines einzigen Befehls bestimmt ist.

Das Netzgerät liefert Hoch- und Niederspannung für jeden Empfänger (insgesamt 60 W). Es besteht aus einem stabilisierten Transistor-Umwandler, der von der Flugzeugversorgung mit 28 V gespeist wird.

In der Abbildung ist der Achtton-Empfänger links, der Zweiton-Empfänger dahinter, und die beiden Netzgeräte rechts zu sehen.

EE 14 756 *für weitere Einzelheiten*

FLUGLAGENANZEIGER FÜR FERNLENKKÖRPER

(Abbildung Seite 625)

Der GEC Fluglagenanzeiger für Fernlenkkörper wurde für die Fernlenkwaffenabteilung des Royal Air Force Establishment entwickelt, um die Fluglage einer „Skylark" Versuchsrakete beim Rollen, Stampfen und Gieren im Flug zu messen.

Zwei CW X-Band Hochleistungssender werden auf jeder Seite der Feuerstellungsmittellinie aufgestellt, um die Rakete zu „beleuchten". In einem besonderen Abteil der Rakete befinden sich, auf einen kräftigen Ring montiert, Schaltkreise und drei eingebaute Mikrowellenantennen, deren Anordnung zwei rechtwinklige Interferometer bildet. Die Mikrowellenschaltung ist fast vollständig auf diesen kräftigen Ring beschränkt.

Die Winkel zwischen jeder Interferometerbasis und jedem Sender werden mittels Phasenvergleich gemessen. Durch Verknüpfung der bekannten Raketenposition im Weltraum mit je drei der vier Winkel können Rollen, Stampfen und Gieren innerhalb weniger Bogenminuten bestimmt werden. Das Interferometer misst die Phasendifferenz zweier Signale, die von derselben Quelle durch zwei Antennen empfangen werden.

Das GEC-System vermeidet die Schwierigkeit, in zwei Verstärkern den gleichen Phasengang aufrecht zu erhalten. Durch einen Mikrowellen-Einseitenbandmodulator wird dem örtlichen Überlagerereingang zu einer Mischstufe eine niedrige Frequenz ω zuaddiert.

Während des Mischens werden die Frequenz- und Phasendifferenzen aufrecht erhalten, und nachdem die zwei Signale addiert sind, erfolgt die weitere Verstärkung in einem gemeinsamen ZF-Verstärker. Nach Demodulation und weiterer Verstärkung wird die Phase der niedrigen Frequenz mit einer Bezugsfrequenz ω verglichen, die direkt vom Modulator kommt. Die erhaltene Spannung ist der Phasendifferenz proportional, die an den beiden Empfangsantennen besteht. Das Resultat wird für Fernmessung zum Boden zurückgestrahlt.

EE 14 757 *für weitere Einzelheiten*

HONEYWELL CONTROLS LTD

Ruislip Road, East Greenford, Middx.

DATENVERARBEITUNG

(Abbildung Seite 626)

Das Honeywell Datenverarbeitungssystem fragt eine Anzahl verschiedener elektrischer Signale von Thermoelementen, Druckumwandlern usw. ab. In einem Arbeitsgang werden diese Variablen gemessen, in Ziffern umgewandelt, die Information in Form beleuchteter Zahlen auf der Vordertafel angezeigt und über eine elektrische Schreibmaschine in einem Logblatt eingetragen. Ausser diesem einfachen Arbeitsvorgang kann das Gerät, wenn nötig, auch ungeradlinige Eingänge in geradlinige umwandeln und Eingänge mit vorgewählten oberen und unteren Alarmpegeln vergleichen. Alarmpegel und Umwandlerbereiche können in dieser Anlage mittels Schalttafeln und Verbindungsschnüren sehr bequem geändert werden.

Die Abfragegeschwindigkeit des Systems ist sechs Punkte per Sekunde für obere und untere Alarmpegel, und die elektrische Schreibmaschine kann Variable mit einem Punkt per Sekunde eintragen. Die Messgenauigkeit is ±0, 1% der Anzeige oder eine Ziffer.

Das Gerät ist für die genaue Aufzeichnung einer grösseren Anzahl von Veränderlichen in Ziffern (ungefähr 40 oder mehr), sowie die Überprüfung vieler Punkte gegen obere oder untere Alarmpegel geeignet.

EE 14 758 *für weitere Einzelheiten*

Automatischer Wellenanalysator

Der automatische Wellenanalysator ist ein sehr genau arbeitendes Gerät der Überlagerungstype zur automatischen Aufzeichnung der Frequenzanalyse komplexer Datensignale, deren Komponenten zwischen 30 Hz und 2 KHz liegen. Durch die hohe Selektivität der Anlage können Daten analysiert werden, deren untere Grenzfrequenz unter der für viele Wellenanalysatoren üblichen liegt. Durch automatische Steuerschaltungen können grosse Datenmengen mit nur gelegentlicher Wartung durch Bedienungspersonal analysiert werden.

Das Gerät kann mit endlosen Magnetbändern zur Analyse von Signalen Verwendung finden, die nur wenige Sekunden anhalten. Für diesen Einsatz bietet Honeywell geeignete Magnetband-Aufnahme- und Wiedergabegeräte an.

Die Analyse kann für Amplitude über Frequenz oder Leistung über Frequenz erfolgen. In einigen Fällen kann das Gerät auch für eine Amplitude-gegen-Zeit von Signalen Verwendung finden, deren Frequenzkomponenten sich mit der Zeit ändern.

EE 14 759 *für weitere Einzelheiten*

MARCONI'S WIRELESS TELEGRAPH CO LTD

Chelmsford, Essex

Doppler-Navigationsgerät

(Abbildung Seite 626)

Doppler-Navigationsgeräte AD 2300 arbeiten in verschiedenen Ausführungen nach einem frequenzmodulierten CW System auf 8800 MHz mit 1 W Leistung. Um den verschiedensten Ansprüchen gerecht zu werden, wird die Anlage entweder als Richtungsbestimmer mit eingebautem Anzeigetreiber (Angabe der Hauptdaten für Grundgeschwindigkeit, Abtriftwinkel und geflogener Entfernung) oder als Richtungsbestimmer mit einem der zwei geeigneten Rechner geliefert. Die der Typennummer folgenden Buchstaben kennzeichnen die verschiedenen Modelle:

Type AD 2300 *A (Richtungsbestimmer mit Anzeigetreiber) zeigt an:*

(a) Grundgeschwindigkeit und Abtriftwinkel

(b) Geflogene Entfernung

Type AD 2300 *B (mit Rechner Type* 4381 *anstelle des Anzeigetreibers)*

(a) Grundgeschwindigkeit und Abtriftwinkel Type AD 2300 A

(b) Verbleibende Entfernung auf einer gewählten Teilstrecke

(c) Fehler quer zur gewählten Teilstrecke in km (links oder rechts)

(d) Verbleibende Flugstrecke zum Bestimmungsort in km (gesamt)

(e) Plan für die nächste gewählte Teilstrecke

(f) Steuerdaten für den Kursanzeiger des Piloten (oder für automatische Kurssteuerung)

Type AD 2300 *C (mit dem abgebildeten Rechner Type* 4382 *anstelle des Anzeigetreibers)*

(a) Grundgeschwindigkeit und Abtriftwinkel

(b) Windgeschwindigkeit und -richtung

(c) Geographische Länge und Breite des Messpunktes

(d) Magnetische Abweichung

(e) Verbleibende Flugstrecke zum Bestimmungsort in km (gesamt)

(f) Verbleibende Entfernung auf einer gewählten Teilstrecke in km

(g) Plan für die nächste gewählte Teilstrecke

(h) Steuerdaten für den Kursanzeiger des Piloten (oder für automatische Kurssteuerung.

Die angeführten drei Ausführungen genügen den Anforderungen aller Flugzeugkategorien. Mit den beiden letzteren Typen kann auf Wunsch eine Rollkartenanzeige des Flugzeugkurses geliefert werden. Auf dem Stand wurde eine Anzeige dieser Art mit dem Marconi Doppel-Navigationsgerät AD 2000 B gezeigt, die von Kelvin and Hughes Ltd. hergestellt wird.

Gerät AD 2300 kommt beim Abflug in Betrieb und ist bis zu 15.250 m Höhe und 30° Querneigungswinkel voll wirksam. Die Genauigkeit wird für Abtriftwinkel bis zu 30° (entweder Backbord oder Steuerbord) vollständig aufrecht erhalten. Grundgeschwindigkeitsbereich von 148,25 km/h...1667,78 km/h.

Ein elektromechanisches Kursfolgesystem ist eingebaut, das beim Versagen der Anlage ausfällt, d.h. solange ein brauchbares Dopplersignal vorhanden ist, muss das Kursfolgegerät richtig ansprechen. Wenn das Signal ausfällt, schaltet die Kursfolge automatisch auf „Speicher" und verriegelt sich wieder mit dem Signal, wenn der Pegel hoch genug ist. Erschöpfende Flugerprobung über viele Flugstrecken der Welt hat gezeigt, dass das Gerät selbst bei Bodenwind unter 9,25 km/h und glatter See genügend Spielraum hat, um selbst über 13,700 m Höhe voll betriebsfähig zu bleiben.

EE 14 760 *für weitere Einzelheiten*

Einkanal-UKW-Peilgerät

(Abbildung Seite 626)

Über 200 Marconi UKW-Bodenpeilstellen Type AD 200 sind täglich in der ganzen Welt in Gebrauch. Diese Typen werden nunmehr durch die AD 210 Anlagen ersetzt.

Die in Farnborough in Betrieb vorgeführte Type AD 210 C ist in ihrer einfachsten und preisgünstigsten Gestaltung besonders für Bodenpeilung auf kleineren Flughäfen geeignet. Trotzdem es sich grundsätzlich um eine Einkanalanlage handelt, kann sie durch Kaskadenschaltung vorhandener Geräte für Mehrkanalarbeiten eingesetzt werden.

Bei der Entwicklung des Gerätes wurde grosser Wert auf Erzielung höchster Betriebsbereitschaft gelegt. Der Frequenzbereich der Anlage liegt zwischen 100...156 MHz, und fünf quarzgesteuerte Frequenzen können durch Druckknopfschaltung von der Hauptsichtanzeige aus eingesetzt werden.

Die Richtungsanzeige erfolgt automatisch auf einem Bildschirm von 203 mm ϕ. Normalerweise ist die Sichtanzeige in ein Pult eingebaut, auf Wunsch kann sie aber auch in das Hauptkontrollpult verlegt werden. Zusätzlich zur Hauptsichtanzeige kann auch ein Nebensichtgerät eingeschaltet werden, das bis 152 m entfernt aufgestellt werden kann.

Eine einfache Fernsteuerung kann Anlagenbedienung bis zu 12,8 km weit über Fernsprechleitungen übernehmen.

Die Peilgenauigkeit ist für vertikalpolarisierte Wellen äusserst gut, und der Betriebsfehler ist im ganzen Bereich in der Grössenordnung von ½° . . . . 2°.

EE 14 761 *für weitere Einzelheiten*

MICROCELL LTD

56 Kingsway, London W.C.2.

Schnellfolgeschalter

Microcell Electronics, eine Abteilung der Microcell Ltd. hat kürzlich im Zusammenhang mit dem Versorgungsministerium die Aufgabe gelöst, gefahrenlos in schneller Folge elektromagnetische Verriegelungen auszuklingen.

Aus elektrischen Gründen konnte die Entriegelung nicht gleichzeitig erfolgen. Dadurch bestand Bedarf für einen Schnellfolgeschalter, der die Verriegelungen hintereinander mit einem Zeitabstand von 10 ms ausklingen würde.

Die Konstruktion unterlag schweren Einschränkungen, da die Abmessungen klein gehalten werden mussten. Trotzdem sollte der Schalter normale elektronische Prüfmethoden bestehen und musste, wenn nötig, reparaturfähig sein, statt weggeworfen zu werden.

Die Aufgabe wurde durch Entwicklung eines transistorbestückten, elektronischen Schalters gelöst. Der Schalter enhält eine Anzahl unmittelbar aufeinander folgender Stufen. Jede Stufe schaltet einen 5 A Impuls zu den elektromechanischen

Verriegelungen. Keine Stufe kann schalten, bevor die vorhergehende ihre Funktion beendet hat. Der Schalter wartet eine begrenzte Zeit, wenn eine Verriegelung klebt. Eine besondere Schutzvorrichtung wird wirksam, wenn die Verriegelung sich nicht ausklingen lässt und ermöglicht die Wiederaufnahme der Folge.

Der elektronische Schalter ist einem ähnlichen elektromechanischen Schalter überlegen, da er ohne bewegliche Teile arbeitet, die sich festklemmen oder abnutzen können.

EE 14 762 *für weitere Einzelheiten*

MULLARD EQUIPMENT LTD

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

ELEKTROMAGNETISCHES VERMESSUNGSGERÄT

Ein besonders für geophysikalische Vermessung geeignetes „Twin Pioneer" Flugzeug der Scottish Aviation wurde mit einem von Mullard Equipment gelieferten elektromagnetischen Vermessungsgerät ausgerüstet.

Das Flugzeug wurde für die internationale Rio Tinto Gruppe gebaut und ausgerüstet, die weitreichende Bergbauinteressen hat, und wurde in Farnborough durch Scottish Aviation ausgestellt.

Das elektromagnetische Gerät wurde zur Erfassung von Zonen leitender Minerale, z.B. massiver, nahe der Erdoberfläche liegender Metallsulfidlager entwickelt. Es besteht im wesentlichen aus einem Sender, einem Empfänger und einer Sichtanzeige, die mit Hilfsgeräten für zusätzliche Information, die zur Auswertung der Hauptdaten wertvoll sind zusammenarbeiten.

Die Senderleistung wird einer abgestimmten Spule in einer Verlängerung des Steuerbord-Tragflächenendes zugeführt und ruft in ihr ein magnetisches Wechselfeld und damit eine Spannung in einer zweiten Spule hervor, die in gleicher Weise in der Backbord-Tragfläche montiert und zum Empfänger gekoppelt ist. Sowohl Sender- als Empfängerspule können so ausgerichtet werden, dass ihre Achsen entweder horizontal oder vertikal zur normalen Flugbahn liegen, entsprechend der wahrscheinlichen Ebene der Erzkörper in der zu vermessenden Gegend.

Fliegt das Flugzeug ausserhalb des Bereiches möglicher Bodenbeeinflussung, so wird die in der Empfängerspule hervorgerufene Spannung durch eine Spannung gleicher Grösse, aber entgegengesetzter Phase kompensiert, die von einer zweiten Wicklung der Sendespule zugeleitet wird. Das System ist dann ausgeglichen. Wenn das Flugzeug sich jedoch einem grossen, leitenden Körper nähert, werden Wirbelströme, die das magnetische Feld der Empfangsspule ändern, im Körper hervorgerufen. Dadurch wird der Spannungsausgleich in der Empfängerspule gestört und ein unsymmetrisches Signal hervorgerufen. Nach Vorverstärkung wird dieses Signal dem Empfänger zugeführt, weiter verstärkt und in phasengleiche und Quadraturkomponente zerlegt. Die entstehenden Ausgangssignale werden dann zur Aufzeichnung auf einem mitgeführten Mehrkanalschreiber verstärkt.

Die Aufzeichnung der phasengleichen als auch der Quadratursignale, in Verbindung mit der Wahl einer geeigneten Sendefrequenz, ermöglicht es dem Gerät, zwischen abbaufähigen Lagern und Lagern geringer Leitfähigkeit wie z.B. Sumpf und salzhaltigen Oberschichten zu unterscheiden.

Ausser den phasengleichen und Quadraturausgängen zeigt ein dritter Kanal auf einer benachbarten Frequenz die Grösse des atmosphärischen Rauschens an, und ein vierter zeichnet die Beanspruchung der Tragflächenstreben auf, die Gegenwart von Böen und dadurch mögliche falsche elektromagnetische Ablesungen anzeigen. Weitere Kanäle stehen zur Kurvenaufzeichnung für andere im Flugzeug vorhandene Vermessungsgeräte zur Verfügung, z.B. Magnetometer, Funkecholot und Scintillometer zur Feststellung der Gammastrahlungen möglicher Uranium- und Thoriumlager.

EE 14 763 *für weitere Einzelheiten*

REDIFON LTD

Broomhill Road, London S.W.18

HF-SENDER

(*Abbildung Seite 627*)

Redifon haben eine völlig neue HF-Senderserie für die verschiedenen Anforderungen des Luftfahrverkehrs angekündigt. Das Hauptgerät der als G.420 bezeichneten Serie ist ein Einseitenbandsender mit 1,5 kW Höchstnutzleistung im Frequenzbereich 1,5...30 MHz. Sie umfasst Modelle für A1, A2, A3, A3a, A3b und F1, F2, F3, F4, und F6 mit Fremdtreiber. Bedienung erfolgt entweder direkt, oder mit Verlängerungen von Hand, oder durch volle Fernsteuerung mit Servo-Abstimmung. Bausteinkonstruktion ermöglicht durch Zusatz eines geeigneten Treibers Anpassung an wechselnde Verkehrsanforderungen. Bis zu acht gerastete Frequenzen sind in einem HF-Gerät vorgesehen, oder wenn der Frequenzwechsel sehr schnell erfolgen muss, können bis zu acht fest abgestimmte HF-Geräte von gemeinsamen Treiberkreisen gespeist werden.

Die Senderserie G.420 ist in folgenden 3 Hauptausführungen lieferbar:

(a) Handbedienung, Einseitenbandbetrieb mit Handeinstellung des Frequenzwechsels. Fernsteuerung, mit Ausnahme des Frequenzwechsels, kann vorgesehen werden. Ein vorhandener Sender kann durch Zusatzgeräte von Handbedienung auf Servo-Steuerung umgestellt werden.

(b) Servo-Steuerung, für Einseitenbandbetrieb und Fernsteuerung der Frequenzumstellung für acht gerastete Frequenzen. Volle Fernsteuerung ist lieferbar.

(c) Mehrfachgestell, Einseitenbandbetrieb und sehr schnelle Fernwahl für acht HF Geräte. Diese Ausführung ist besonders für Verwendung in Anlagen bestimmt, in denen die Umschaltzeit reduziert werden muss und täglich eine grosse Anzahl von Umschaltungen erforderlich sind.

EE 14 764 *für weitere Einzelheiten*

UKW-SPRECHFUNK

(*Abbildung Seite 628*)

Das amplitudenmodulierte UKW Sprechfunkgerät GR.318 moderner Konstruktion für den Sprachverkehr zwischen Fahrzeugen und einer geeigneten Hauptstation. Es ist ein kompaktes Gerät mit 4...5 W Leistung und geringem Verbrauch und kann den verschiedensten Einbaubedingungen leicht angepasst werden. Durch diese Eigenschaften ist es für bewegliche Bodendienste eines Flughafens besonders geeignet.

Erhöhte Betriebssicherheit und Ersparnisse in Batterieverbrauch werden durch vorteilhaften Transistoreneinsazt erreicht. Die Umwandlung für Hochspannung geschieht in einem leistungsfähigen Transistor-Umformer; ausserdem erfüllt der transistorbestückte Modulator einen Doppelzweck, da er auch als Empfängerendstufe verwandt wird.

Infolge kleiner Abmessungen kann das GR.318 bequem unter dem Instrumentenbrett oder Dach des Fahrzeugführerabteils untergebracht werden.

Zur Wahrung der vollständigen Anpassungsfähigkeit beim Einbau kann ein Nebensteuergerät geliefert werden, in dem alle Bedienungselemente, einschliesslich Frequenzumschalter für das Doppelkanalmodell, ein zweites Mal vorhanden sind. Montagebretter für beide Einbauausführungen, sowie ein widerstandsfähiges Handmikrofon, werden mitgeliefert. Ein Fernsprechhörer oder Mikrofon und Kopfhörer können auch wahlweise benutzt werden.

Modelle sind für sieben Frequenzbänder im Bereich 72...175 MHz lieferbar.

EE 14 765 *für weitere Einzelheiten*

ROYSTON INSTRUMENTS LTD

Canada Road, Oyster Lane, Byfleet, Surrey

FLUGDATEN-AUFZEICHNUNGSGERÄT

(*Abbildung Seite 628*)

Die Entwicklung des umfassenden MIDAS Betriebsflugdaten-Aufzeichnungsgerätes mit Magnetbandverwendung stützte sich auf die Erfahrung dieser Firma in der Konstruktion von Aufzeichnungs- und Verarbeitungsanlagen für Flugzeugforschung und -prüfung. Aufgabe des Gerätes ist die Beschaffung einer vollständigen, genauen Daueraufzeichnung jedes wichtigen Betriebsparameters vom Hochfahren bis zum Landen für die Lebensdauer des Flugzeuges. Beschleunigte automatische Verarbeitung der Daten ermöglicht es, die Aufzeichnung in eine kompakte, bearbeitete und nützliche Form umzuschreiben.

Die Anlage besteht aus einem Magnetband-Aufzeichnungsgerät mit besonders geringer Bandgeschwindigkeit, dessen Grundzüge sich seit Jahren im MIDAS-System bewährt haben. Das Gerät hat sieben unabhängige parallelgeschaltete Signalspuren. Eine Spur ist für Zeit, Datum, Flugnummer usw. vorbehalten. Die Aufnahmefähigkeit der übrigen sechs Spuren wird durch einen Zeitfolgeschalter erhöht, der eine grosse Anzahl von Parametern hintereinander einschalten kann. Tatsächlich sind sechs Schalterreihen vorhanden, die dauernd von einem gemeinsamen Motor angetrieben werden. Jede Schalterreihe speist eine entsprechende Aufzeichnungsspur und fasst dadurch eine Anzahl verschiedener Parameter zusammen. Dieses Aufzeichnungsgerät wird, unabhängig von der Montage und den zu messenden Parametern, in genormter Ausführung für alle Flugzeuge geliefert, soweit nicht Sonderausführungen bestellt werden. Konstruktion und Einzelheiten des Zeitfolgeschalters sind genormt, jedoch wird er mit dem mehrwegigen Eingang entsprechend der Instrumentausrüstung jedes Flugzeuges verdrahtet. Er wird in einem geeigneten Mittelpunkt des Umwandlernetzes eingebaut (wahrscheinlich in der Kabine) und kann mittels Mehrfachsockel und -stecker auf die Rückseite des Aufzeichnungsgerätes montiert werden. Wo jedoch für den Notfall ein Auswerfen vorgesehen werden soll, kann das Gerät im Heck eingebaut werden, und ein zehnadriges Kabel genügt für die Verbindung zum Schalter.

Die Gesamtzahl der abgefragten Parameter und die Abfragegeschwindigkeit hängt vom Zeitfolgeschalter ab. Der Mehrwegschalter macht eine Umdrehung in fünf Sekunden, so dass jeder Parameter mindestens einmal jede fünf Sekunden abgefragt wird. Die Abfragegeschwindigkeit kann verdoppelt oder verdreifacht werden, wenn Abschnitte des Schalters in diskreten Abständen parallelgeschaltet werden, so dass diese Signale zwei- oder dreimal für jede Schalterumdrehung abgefragt werden. Jede Spurenzahl und Kombination von Abfragegeschwindigkeiten kann daher bis zu einem Maximum von 300 Kanälen eingestellt werden, wobei dann jeder einmal in 5 Sekunden abgefragt wird.

Die Magnetbandspulen werden in eine Kassette geladen und können dann von ungelerntem Personal eingesetzt oder entfernt werden. Das bedeutet in Verbindung mit einem Aufzeichnungsvermögen von 160 Stunden, dass eine vollständige Aufzeichnung für die längsten Rundflüge selbst dann keine Schwierigkeit bietet, wenn das Magnetband nur bei Rückkehr zum Hauptstützpunkt ausgewechselt wird. Das Aufzeichnungsgerät hat eine unbeschränkte Lebensdauer, ermöglicht sehr hohe Genauigkeit, ist klein (1 ATR) und wiegt unter 18,1 kg.

Das Gerät hat eine sehr hohe Eingangsimpedanz und eingebaute Anordnungen für die Signalumwandlung von Dreidraht Synchron- zu Analogspannungen. Dadurch sind in fast allen Fällen nur einfache Verbindungen zur bestehenden Instrumentausrüstung für den Einbau erforderlich.

Wiedergabe- und Datenverarbeitungsgeräte für Einsatz mit diesem Gerät sind lieferbar.

EE 14 766 *für weitere Einzelheiten*

THERMIONIC PRODUCTS (ELECTRONICS) LTD

Hythe, Southampton

MEHRSPUR-MAGNETBANDGERÄT

(Abbildung Seite 628)

Diese Bandgeräte werden aus einer Anzahl getrennter Bausteine aufgebaut. Durch Wahl geeigneter Bausteine können Aufzeichnungsgeräte zusammengestellt werden, die den meisten Anforderungen entsprechen.

Das abgebildete Gerät 1A/1A kann gleichzeitig bis zu 20 verschiedene verschlüsselte oder Sprechsignale aufnehmen. Normalerweise wird eine Spur zur Zeitaufzeichnung und eine zweite Spur als automatische Reserve eingesetzt.

Eine durch Sprache oder Träger betätigte Anlauf/Halt Vorrichtung ist auch vorgesehen und ermöglicht bis zu 24 Stunden Betrieb ohne Wartungspersonal. Diese Betriebsdauer wird proportional zu der Gesamtzeit, während der keine Signale aufgenommen werden verlängert.

Alle Spuren werden dauernd überwacht, und beim ersten Ausfall wird automatisch auf die Reserve umgeschaltet, ohne dass die Aufzeichnung unterbrochen wird. Volle Wiedergabeeinrichtungen sind im Bereitschaftsgestell vorhanden.

Die Anlage ist durchgehend transistorbestückt, und gedruckte Schaltungen werden in allen geeigneten Bausteinen verwandt, um die Betriebssicherheit zu erhöhen. Die Uhr und Verschlüsselung sind quarzgesteuert und haben zur Vereinfachung der Wartung Einsteckstufen.

EE 14 767 *für weitere Einzelheiten*

THORN ELECTRICAL INDUSTRIES LTD

105-109 Judd Street, London W.C.1

ELEKTROLEUCHTANZEIGER

(Abbildung Seite 629)

Die Abteilung für Flugzeugbauelemente und Kupplungen der Thorn Electrical Industries Ltd stellt jetzt einen Elektroleuchtanzeiger-Baustein mit 100 Punkten her, der für Ziffernanzeigen und Überwachungskartensysteme ein neues Prinzip anwendet.

Der Baustein ist nur 6,73 mm tief und beansprucht nur einen Bruchteil des von Lampenanzeigern eingenommenen Raumes. Da keine Glühfäden verwandt werden, wird praktisch keine Hitze erzeugt. Der Baustein besteht aus 100 rechteckigen Phosphorpunkten, die zwischen zwei leitende Flächen geschichtet sind. Wenn eine Wechselspannung an diese Punkte gelegt wird, leuchten sie auf. Durch Änderung des Leuchtpunktmusters kann der Anzeiger jeden gewünschten Buchstaben, oder jede Zahl, oder ein genaues Bild des Fortschrittes anzeigen.

Der Nennwert für Anzeigehelligkeit ist 15 Fuss/Lambert. Es ist wünschenswert, HF-Erregung mit ungefähr 350 V Spannung vorzusehen, um die erforderliche Helligkeit zu erreichen.

EE 14 768 *für weitere Einzelheiten*

ULTRA ELECTRIC LTD

Western Avenue, London W.3

ABSTURZ-POSITIONSANZEIGER

Dieses Gerät ist eine interessante Erweiterung des SARAH Such- und Rettungsgerätes. Sein Einsatz würde schnelle und genaue Feststellung des Standortes der Überlebenden und des Wracks selbst bei dichtem Nebel, niedrigen Wolken oder in abgelegenen Land- oder Seegegenden sicherstellen.

Der Absturz-Positionsanzeiger (APA) ist ein Richtfunksender, der auf einer dünnen Blechantenne aufgebaut ist. Er ist in einen zähen Schaumstoff eingebettet, der als Tragfläche ausgebildet und mit einer geschichteten Nylonhaut versehen ist.

Für den Gebrauch ist der APA in einen festverbundenen Teil des Flugzeugrumpfes eingebaut. Eine empfindliche Auslösevorrichtung wird durch Drähte betätigt, die entlang der Flugzeugstruktur gespannt sind und die erste Phase eines Zusammenstosses oder ungewöhnlichen Verhaltens entdecken. Ein modernes Schnellflugzeug kann in weniger als 100 ms vollständig zerstört werden; der APA wird dann freigeworfen, landet in sicherer Entfernung vom Wrack und beginnt die Richtfunksendung.

Unter normalen und günstigen Bedingungen kann er 48 km weit empfangen werden, seine Höchstreichweite ist 112 km. Unter schwierigen Bedingungen, z.B. wenn der APA in einem tiefen Tal unter Schnee begraben ist, ist die Reichweite immer noch 8...16 km.

Der Sender kann unter Wasser arbeiten und hat eine ununterbrochene Betriebslebensdauer von mehr als 100 Stunden.

Der APA landet auf Felsen oder Eis sicher bis zu 64 km/h und auf Schnee oder weichem Boden mit einem Mehrfachen dieser Geschwindigkeit. Er taumelt mit ungefähr 42 km/h aus dem Flugzeug und schwimmt mit 80% ausserhalb des Wassers.

Das APA Signal verträgt sich mit dem SARAH Suchgerät, das zur Ausrüstung der Streitkräfte des britischen Commonwealth und der meisten NATO Länder gehört. SARAH überdeckt vollständig Nordatlantik, Nordsee, Ostsee, Karibisches Meer, Mittelmeer, Indischen

Ozean, Bengalische Bucht, Australasien, Südchinesische See und den westlichen Stillen Ozean.

EE 14 769 *für weitere Einzelheiten*

TRIEBWERK-BETRIEBSZUSTANDS-ANALYSATOR

(*Abbildung Seite 629*)

Dieses neue Gerät wurde für die Datensichtanzeige von Triebwerktemperatur und -schwingung in einer Form entwickelt, die Beurteilung auf einen Blick ermöglicht. Es handelt sich um eine vielseitige Anlage, die für Prüfung auf Flugklarheit, Flugüberwachung und laufende Wartung Verwendung findet.

Temperaturanzeigen von 40 Thermoelementen können hintereinander auf dem oberen Bildschirm der Sichtanzeige erscheinen. Die Temperaturanzeigen sind in vier Gruppen unterteilt, d.h. eine Gruppe per Triebwerk. Die zehn Temperaturpunkte für jedes einzelne Triebwerk können durch ein Frontplatten-Bedienungselement auf der vollen Bildschirmbreite erscheinen. Zwei horizontale Linien sind während der Temperaturanzeige auf dem Schirm sichtbar: eine ist die Temperaturdatumlinie und die andere eine Eichlinie. An der linken Seite der Temperaturanzeige ist eine Skala mit Marken in 25°C Abstand. Die Empfindlichkeit ist 300°/Zoll (118°/cm) und das Gerät wird vor Einbau so eingestellt, dass der Nennwert der Triebwerksbetriebstemperatur auf der Bildschirmmitte erscheint. Ein Frontplattenschalter stellt eine von vier möglichen Stellungen der Datumlinie ein. Die vier Stellungen entsprechen Abflug, Steigen, Fliegen, Sinkflug. Die Datumlinie geht durch Triebwerk-Mitteltemperaturanzeige, wenn das Triebwerk einwandfrei läuft: daher fällt die Temperaturdrift an einem der Messpunkte sofort auf. Die Lage der Linie für jeden der vier Zustände wird für die entsprechende Triebwerktype vor Einbau des Analysators eingestellt.

Die Temperatur-Eichlinie kann durch ein Bedienungselement auf der Frontplatte verschoben werden: wenn sie sich mit dem zu messenden Punkt deckt, kann die Temperatur dieses Messpunktes mit einer Genauigkeit von ½% der Vollanzeige auf der Skala des Bedienungsknopfes abgelesen werden. Die ungefähre Temperatur kann von der Bildschirmskala abgelesen werden.

Auf dem unteren Bildschirm erscheinen laufend die Anzeigen zweier Schwingungsumwandler per Triebwerk. Eine Schwingungswarnlinie erscheint quer über dem Bildschirm. Die Höhe dieser Linie wird vor Einbau des Analysators für diese besondere Triebwerktype so eingestellt, dass die Anzeigen unter der Linie bleiben, solange die Triebwerke einwandfrei laufen. Die Anzeigepunkte entsprechen Schwingungsamplituden erster Ordnung; Schwingungen, die nicht der Triebwerkdrehgeschwindigkeit entsprechen, werden nicht angezeigt.

EE 14 770 *für weitere Einzelheiten*

WAYNE KERR LABORATORIES LTD

Roebuck Road, Chessington, Surrey

ABSTANDSMESSER

(*Abbildung Seite 629*)

Dieses Gerät arbeitet mit nichtberührenden Fühlern und ermöglicht Abstandsmessungen von 50μ Zoll ... 0,5 Zoll (0,127μ ... 12,7 mm) mittels einer an sich geradlinigen Methode. Abstände werden direkt auf einem Drehspulinstrument abgelesen, dessen Messbereich von den in Gebrauch befindlichen Fühlern abhängt. Die vier lieferbaren Fühler geben Vollausschlagsbereiche von 0,025 mm, 0,254 mm, 2,54 mm und 12,7 mm.

Die Ablesegenauigkeit liegt innerhalb ±2% des Vollausschlages, und das Unterscheidungsvermögen ist besser als 0,5% des Vollausschlages.

Das 3 m lange Fühlerkabel ermöglicht es, das Gerät räumlichfrei vom Prüfling aufzustellen. Ein Anschluss für Schreiber oder Servo-Steuerung ist auf der Frontplatte vorgesehen.

Das Gerät macht schnelle und einfache Abstandsmessungen beim Präzisionsschleifen möglich und kann auch leicht anderen Verwendungszwecken angepasst werden, z.B. Auswuchten von Hubschraubenblättern und Messung der Senkrechtstellung von Regelstäben in Reaktor-Kraftwerken.

Direkte Ablesung der Ausdehnung kann ohne Belastung des Prüflings erfolgen, und Abmessungsänderungen können dauernd überwacht oder unter Bezugnahme auf Zeit und Temperatur aufgezeichnet werden.

Einsatzmöglichkeiten bestehen überall, wo Direktablesung der Abstandsmessungen bis 12,7 mm sowie Dauerüberwachung erforderlich sind, z.B. zur laufenden Prüfung von Einzelteilen und allen Messungen von Werkstücken während der Maschinenbearbeitung. Das Gerät kann für die automatische Ausscheidung von Ausschuss eingerichtet werden.

Die Konstruktion stützt sich auf einen Hochleistungsverstärker, dessen negativer Gegenkopplungskreis durch die Kapazität zwischen dem Fühler und dem Prüfling geschlossen wird. Der Verstärkereingang ist ein Bezugsstrom, der von einem eingebauten Oszillator abgegeben wird. Der Verstärkerausgang speist eine Gegenkopplungsschaltung mit Voltmeter, dessen Anzeige dem Abstand zwischen Fühler und Prüfling direkt proportional ist.

EE 14 771 *für weitere Einzelheiten*

Zusammenfassung der wichtigsten Beiträge

Ein Amplituden Pegelmesser von M. Drayson

Zusammenfassung des Beitrages auf Seite 578-584

Die Bedeutung des Messens der Signalamplitudenverteilung wird behandelt. Die mathematischen Ausdrücke für die Amplitudenverteilung, mit besonderer Berücksichtigung der für eine praktische Messanordnung erforderlichen Annäherungen, werden beschrieben. Ein neuartiges System der Probenentnahme ermöglicht Messungen mit einer Zerlegung von mindestens 1%. Der Beschreibung folgen eingehende Angaben für die Schaltungen. Das Gerät ist für Amplitudenmessungen von 0,3 . . . 30 V bis 0,1 . . . 10 MHz geeignet, wobei Amplitudenverteilungskurven mit einer Mindestfrequenz von 1 KHz angezeigt werden. Abbildungen zeigen die Verteilungskurven auf dem Bildschirm einer Elektronenstrahlröhre.

Transistor-Gegentakt-Gleichstromumwandler von W. L. Stephenson, L. P. Morgan und T. H. Brown

Zusammenfassung des Beitrages auf Seite 585-589

Für die Umwandlung niedriger Gleichspannungen in hohe sind viele Methoden bekannt, die Generatoren verwenden. Eine der wirksamsten Anordnungen benutzt einen Transistor-Rechteckwellengenerator, dessen Steuerung durch einen Transformator erfolgt, der gesättigt werden kann. Formeln für die Berechnung solcher Anordnungen sind als Betriebsparameter gegeben.

Die praktischen Grenzen und Schwierigkeiten werden aufgezeigt, aber nicht eingehend behandelt, da jeder Fall seine eigene Lösung erfordert.

Gesichtspunkte der Logik und des Schaltungsaufbaus eines Digitalfeldrechners von I. F. Brown und B. Meltzer

Zusammenfassung des Beitrages auf Seite 590-592

Die Grundbegriffe einer neuen Digitalrechnertype zur Lösung von Aufgaben der Feldlehre werden beschrieben. Der Rechner verkürzt die Berechnungszeit beträchtlich durch gleichzeitiges Ausrechnen für alle Gitterpunkte des Feldes. Statt der binären oder Dezimalarithmetik findet eine unäre Arithmetik (Rechenrahmen) Verwendung. Die Berechnung erfolgt mittels Gittereinheiten, wobei jedem Gitterpunkt des Feldes eine Gittereinheit zugeordnet wird. Die Synthese der Gittereinheiten von einfacheren Grundeinheiten wird beschrieben. Der versuchsmässige Aufbau einer für Spannungs- und andere Probleme geeigneten Grundeinheit ist dargestellt. Einige der vielen ausstehenden Probleme dieses Rechners werden kurz gestreift.

Diagramme zur Ableitung der Transistor-r-Parameter von h-Parametern von G. W. E. Stark

Zusammenfassung des Beitrages auf Seite 592-593

Eine Beschreibung von Diagrammen, die auf Grund der in Basisschaltung gemessenen h-Parameter eines Transistors Rückschlüsse auf die Grössen der r-Parameter vereinfachen.

Untersuchung einiger Gesichtspunkte von Dioden Quantisierungs Schaltungen von W. Alexander und H. V. Bell

Zusammenfassung des Beitrages auf Seite 594-598

Quantisierung wird beschrieben und einige frühere Arbeiten über Dioden Quantisierung sind besprochen. Drei Schaltungen werden theoretisch und durch Messung verglichen und die Resultate dargestellt. Anschliessend wird eine Anwendungsmöglichkeit dieser Schaltungen beschrieben.

Ein einfaches elektronisches Tachometer für Kraftfahrzeuge von M. J. Wright

Zusammenfassung des Beitrages auf Seite 599-602

Die Kraftwagenzündung steuert ein elektrisches Signal, dessen Grundfrequenz proportional zur Motorendrehzahl ist. In einer neuen Schaltung werden durch Sieb- und Begrenzungselemente Störschwingungen unterdrückt und eine Rechteckwelle konstanter Amplitude einem Dioden-Frequenzmesskreis zugeführt, der direkt für Motorendrehzahlen geeicht ist.

Die Gesamtschaltung ist einfach, hat keine beweglichen Teile und wird von der Batterie des Kraftfahrzeuges gespeist. Die Eichung wird durch Schwankung der Batteriespannung oder der Umgebungstemperatur nur geringfügig beeinflusst.

Ein einfacher Transistor-Tester von G. G. Yates

Zusammenfassung des Beitrafes auf Seite 602-603

Ein einfaches und preisgünstiges Messgerät für die Parameter der NF und HF Transistoren geringerer Leistung, die mehr Verwendung finden (Ico', β, h_{11}, $f\alpha$), wird beschrieben. Ico' und β werden mittels Gleichstrommethoden ohne Zusatzgeräte gemessen. Für die Ermittlung von h_{11} und $f\alpha$ werden zusätzlich ein Messender mit einem Bereich von mindestens 1 kHz bis ungefähr 200 kHz, sowie ein Röhrenvoltmeter oder Oszillograf benötigt. Da angenommen werden kann, dass selbst anspruchslos eingerichtete elektronische Labors diese Geräte haben, wird das nicht als Nachteil betrachtet.

Ein Transistor-Untersetzer mit kurzer Auflösezeit von B. Collinge

Zusammenfassung des Beitrages auf Seite 604-605

Schnelle Dualstufen werden für verschiedene Anwendungsgebiete benötigt. Die Vorteile einer Schaltung, in der eine Dualstufe durch ein Stromtor mit einer anderen gekoppelt ist, werden beschrieben.

Der Beitrag bringt Beschreibung nebst Schaltbild eines aus einem Stromtor und einem bistabilen Kreis bestehenden Gerätes.

Vorzeichen- und Grössenanzeige für umkehrbare Zähler von D. L. A. Barber

Zusammenfassung des Beitrages auf Seite 605-606

Die beschriebene Schaltung kann mit jedem umkehrbaren Zähler verwandt werden, um Vorzeichen und Grössenordnung der Zahlen anzuzeigen.

Näherungslösung für Diodenwellenformen in Impulsschaltungen von D. C. Dillistone

Zusammenfassung des Beitrages auf Seite 607-610

Von einer Näherungsdarstellung einer Diode ausgehend, werden Wellenformlösungen für einfache Schaltungen mit aufgedrückten Rechtecks-, oder in einem Falle Dreiecksimpulsen entwickelt. Anwendungen der Resultate in gewissen Fällen werden umrissen.

Durch einen Treppenwellenformgeber getriggerter Transistor-Sperrschwinger-Frequenzteiler von F. Butler

Zusammenfassung des Beitrages auf Seite 611-612

Eingebaut in den Frequenzteiler ist ein Treppenwellenformgeber, der eine mehrstufige Wellenform abgibt, für die alle Spannungsstufen gleiche Amplituden haben. Nach einer vorgewählten Stufenzahl triggert die Ausgangsspannung einen Sperrschwinger, dessen periodische Ausgangsimpulse einen vorherbestimmten Teil der Eingangsstufenfrequenz darstellt. Der Frequenzteiler wird von einer Quelle mit sinusförmiger Ausgangswellenform gespeist, die in eine Rechteckwelle umgeformt und begrenzt wird, um das Arbeiten des Gerätes von der Amplitude des Eingangssignals unabhängig zu machen. Ein Ausgangsverstärker ist dem Sperrschwinger nachgeschaltet. Handelsübliche Transistoren und Dioden finden durchgehend Verwendung.

Phasenempfindlicher Diodendetektor unter Belastung von R. Chidambaram und S. Krishnan

Zusammenfassung des Beitrages auf Seite 613-616

Eine theoretische Untersuchung der Wirkungsweise des einfachen phasenempfindlichen Dioden-Gegentaktdetektors unter Belastung wird durchgeführt. Die Charakteristiken der beiden Dioden ändern sich wesentlich in Abhängigkeit vom Signal. Die durch diese Schwankungen hervorgerufene Ungeradlinigkeit des Ausganges wird bewertet; von einer Tabelle kann die Eignung eines gegebenen Detektors beurteilt werden. Die theoretischen Resultate wurden durch Versuche mengenmässig bestätigt.