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Commentary

IT has hardly been necessary to await the publication of the Electronic Engineering Association's Annual Review and Report for 1959 to establish the fact that the British electronic industry has had a most successful year. The signs are everywhere for one to see and everything points to 1960 being an even more prosperous year.

The bare facts are that the industry which doubled its output between 1951 and 1957, is now growing at an annual rate of £30M. For the year under review, the total production of the Radio and Telecommunication industry was valued at £475M, it gave employment to some 350 000 people and ploughed back ten per cent of its turnover into research and development—the highest proportion in the country with the sole exception of the aeronautical industry.

It is a little more than twelve months ago that the separate existence of the Electronic Engineering Association was formed " to take full responsibility for the capital goods side of the industry " comprising sound and television broadcasting equipment, radar and radio navigational aids for military and civil air and marine applications, computers, industrial electronic control equipment, electronic instruments and industrial television.

Today the Association with a membership which includes practically all the leading electronic firms both large and small, can justly claim to have fulfilled its terms of reference, for its own contribution to the total turnover of the industry amounted to close on £160M with direct exports in the region of £26 to £30M.

The Review highlights many of the more important achievements of the members of the Association but much of the industry's activities have been directed towards defence purposes and consequently little information on this subject is available, but it is well known of course that the present day defence strategy leans very heavily on radar and radio communication systems and it is in this field that some notable contributions have been made.

With the higher speeds of military aircraft and the growing importance of the missile in its various forms, the comparatively leisurely methods of prediction of the past have given way to complex data handling and radar systems designed to process a vast amount of information and then to present a clear and unambiguous picture of events. This latest development representing the best endeavours of industry and the various Government research establishments was successfully demonstrated by the aircraft carrier H.M.S. Victorious during her visit to America last year.

Both at home and abroad the British electronic industry has found an expanding market for its products for military applications and among the more notable contracts secured from abroad is one valued at about £7M for an extensive NATO early warning system. The work for this will be shared between Britain and France although the British company will be carrying the major responsibility.

Simultaneously with the EEA's Review has appeared the Report of its Council which differs from its predecessors in that it is confined solely to the work of the Association itself, its committees and its working parties.

It is by no means a dull uninspiring document for it is apparent that the wind of change, to use the current phrase, is already blowing in this young and expanding industry and in order to meet these changes the Association from now on will be known as the Electronics Industry Council with wider and more appropriate terms of reference.

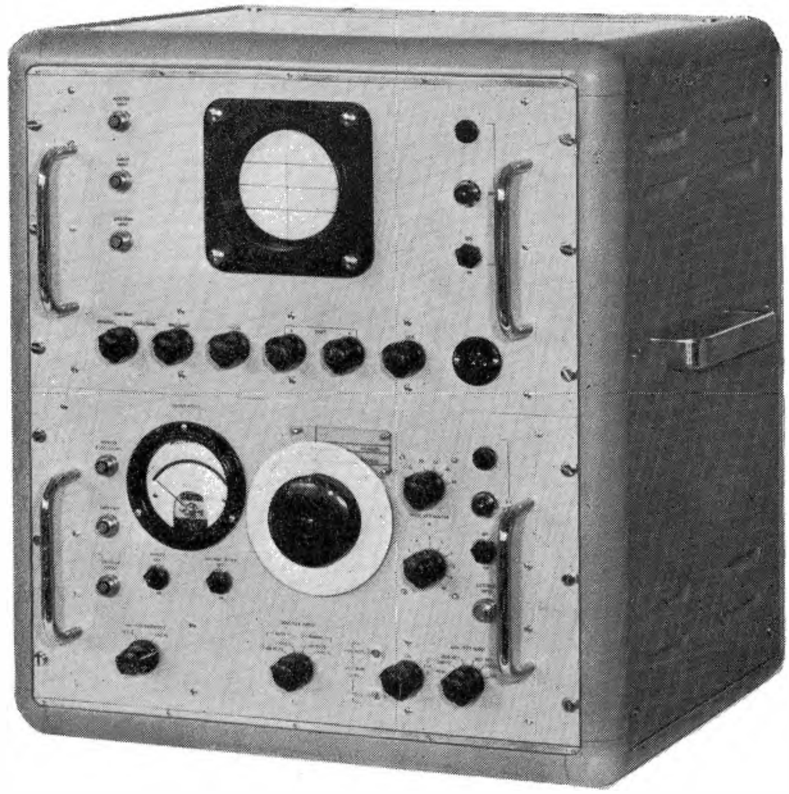
To cover its increasing activities the Association has found it necessary, as a kind of rider to Parkinson's Law, to expand the number of its committees during the current year, the principle ones being those on Digital Data Processing and Analogue Data Processing.

Among the problems with which the Report deals, is one which is engaging the attention of the Association and which still awaits a satisfactory solution, namely the compilation and publication of production and export statistics for the industry. As things stand at present, it is difficult, if not impossible, for the Association to say how well the industry is doing, for the classification of Board of Trade figures are too broad to be of much use, and all it can do is to make an intelligent guess.

There is as yet no Exhibition which contains the activities of the Association's members as a whole and the various sections of the industry continue to hold their more specialized shows. At the end of this month it is the turn of the electronic instrument industry, through the joint sponsorship of the Association and the Scientific Instrument Manufacturers' Association, to show its products at the Instruments, Electronics and Automation Exhibition.

Claiming for itself the title of the greatest exhibition of its kind in the world, the third International Instruments, Electronics and Automation Exhibition will be held at Olympia, London from 23-28 May concurrently with a conference sponsored by the EEA's Electronic Forum for Industry.

A Spectrum Analyser for Video Frequencies



By A. L. Whitwell*, B.Sc., A.M.I.E.E., A.M.Brit. I.R.E.,

and N. Williams*, M.A., A.M.I.E.E.

A general purpose spectrum analyser for the frequency range 500c/s to 95kc/s is described in this article. The principles of design for equipment of this type are not new, but interest in the use of spectrum analysers as general-purpose laboratory and production inspection aids is increasing rapidly. They can be employed for a wide range of measurements and designed for high sensitivity, due to the reduction in the effective noise factor of the input stage, arising from the subsequent filtering action. For example, the instrument described has a threshold sensitivity of about 0·2μV in a 70c/s band and a dynamic range of 50dB.

(Voir page 326 pour le résumé en français: Zusammenfassung in deutscher Sprache Seite 333)

A SPECTRUM (or wave) analyser is basically a four-terminal or band-pass filter whose centre or mean frequency f_0 can be varied in a known manner to any value between two desired frequency limits. Coupled to the output terminals of the network is an indicator, which provides a measure of the output from the filter (see Fig. 1(a)). Such an instrument can be employed to obtain the Fourier series corresponding to a complex harmonic waveform or, alternatively, a noise spectrum, or a combination of both types of signal. The single criterion, which must be satisfied to enable measurements to be made, is that the r.m.s. value of the signal components within the pass-band of the analyser must remain constant when integrated over a sufficiently long period of time. This criterion is obviously satisfied by the nature of harmonic waveforms, however complex, and also by any signal which has the characteristics of electrical noise arising from molecular phenomena. It is not satisfied if the signal contains isolated, discontinuous noise or harmonic components of short duration and random amplitude/frequency distribution.

Any attempt to plot a spectrum by means of the simple system outlined above tends to be a time-wasting, laborious process, involving separate amplitude and frequency measurements at every point on the spectrum at which information is required. Practical analysers are designed to present this information rapidly and in an easily assimilated form.

The effective centre frequency f_0 is frequently changed or 'swept' automatically between the required limits, while the X-plate potentials of an oscilloscope are varied in synchronism. At the same time, the filter output signal is amplified and applied to the Y-plates of the tube, so that an amplitude-frequency graph is plotted on the face of the c.r.o., as indicated in the schematic diagram, Fig. 1(b).

A visual display of this type provides qualitative information about the unknown spectrum, very rapidly and simply. The inherent weakness of such an elementary system is found in its lack of accuracy in the measument of both amplitude and frequency.

The spectrum analyser described here successfully embraces the merits of both systems by combining an oscillographic, or panoramic display, with a power level indicating meter, which can be rapidly switched into operation and adjusted to permit accurate measurements of amplitude and frequency at any part of the spectrum.

Filters employed in wave analysers are generally either 'constant percentage bandwidth' or 'fixed bandwidth' types. Both have their particular advantages and disadvantages, depending on the nature of the required measurement. Specifically, a constant-percentage bandwidth analyser is most useful when the waveform to be analysed has a predominantly 'line' spectrum as, for example, in the examination of acoustic or mechanical vibration phenomena. On the other hand a noise, or combination of noise and line spectra, is difficult to interpret with this type of filter and a fixed bandwidth type is preferable. This is the technique employed in the type 190 instrument, which is illustrated by the schematic diagram, Fig. 2.

* *James Scott & Co. Ltd.*

General Description of System and Measurement Technique

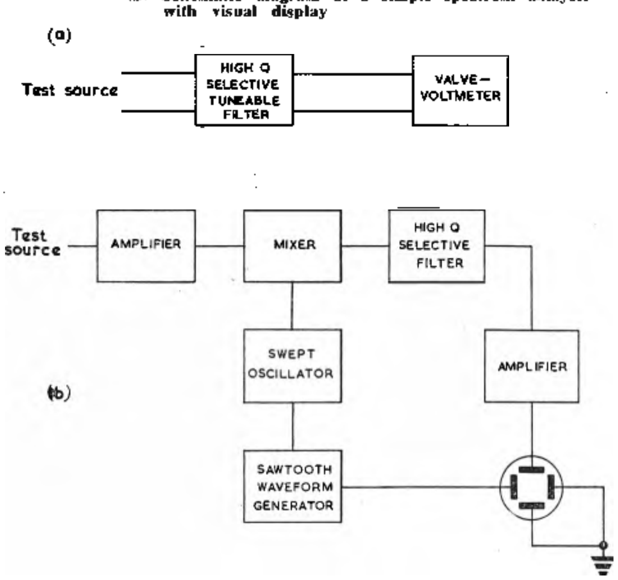
Referring to Fig. 2, the incoming signal is first amplified and then fed to a wide-band mixer, the reference carrier into the mixer being derived from a tuned oscillator with a frequency range from 100·5 to 195kc/s. Sum and difference frequency components are generated in the mixer and appear as output signals which are coupled into a frequency selective amplifier centred on 100kc/s. Thus, only difference signals having frequency components within the pass-band of the tuned amplifier (which is arranged for sharp cut-off beyond its range) are transmitted through to the final stages and measuring circuits of the equipment. For quantitative work, a manually tuned oscillator is used to provide the mixer reference signal, and the output power in a 70c/s band of frequencies is determined by means of a thermocouple microammeter.

In order to permit qualitative visual examination of the spectrum, the manually tracked local oscillator of the spectrum analyser may be switched out of circuit and replaced by a sweeping oscillator covering the required difference frequency in two ranges, 0 to 7kc/s and 0 to 100kc/s. The local oscillator sweep voltage generator is also used to derive the *X*-sweep of an oscilloscope employed for viewing the spectrum.

On the 0 to 7kc/s range, the 70c/s band-pass amplifier is used, but the bandwidth is normally increased to 1kc/s for automatic sweep on the 0 to 100kc/s range in order to produce an easily read spectrum curve. The sweep rate is necessarily rather low, in order to avoid distortion and loss of signal through the narrow-band amplifier and it is, therefore, necessary to use a medium or long-persistence c.r.o.

The time-base is calibrated in frequency by a 'marker' signal derived from the manual oscillator and coupled into the noise amplifier before the mixer circuit. The manual oscillator tracks over the range from 100 to 195kc/s and it is, therefore, necessary to use a second mixer to produce a lower sideband marker signal in the required range from 0·5 to 95kc/s. A crystal-controlled local oscillator is used as a reference in the marker frequency mixer. This method of deriving a marker signal has the particular advantage that once any particular 'line' on the display has been identified by superimposing the marker 'line' on it, very little further adjustment of the manual oscillator frequency

Fig. 1. (a) Basic elements of a spectrum or wave analyser
(b) Schematic diagram of a simple spectrum analyser with visual display



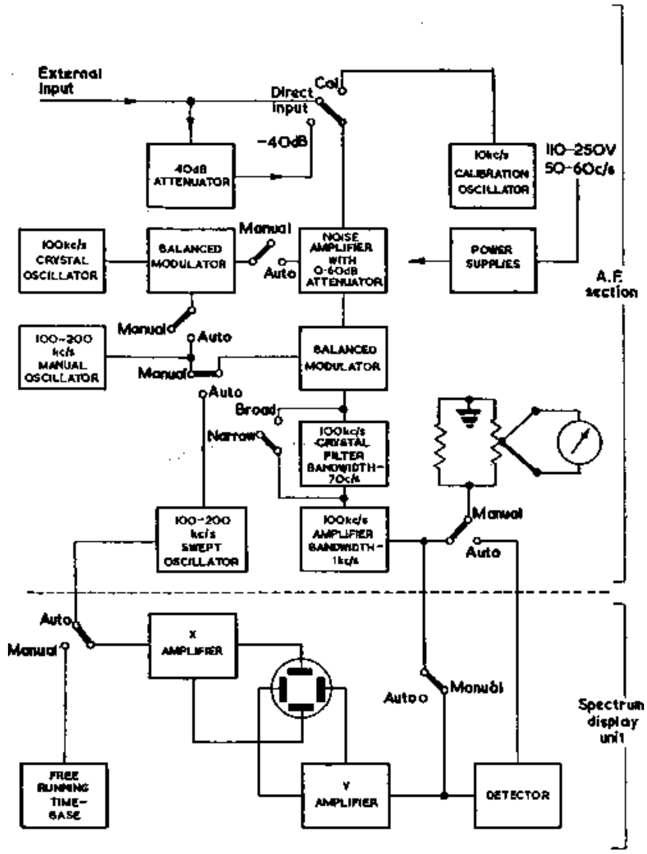


Fig. 2. Schematic diagram of video spectrum analyser Type 190

dial is necessary to peak the meter reading on switching to 'manual', since the same oscillator is employed.

While the instrument is calibrated for a voltage measurement range of 100dB the range of input levels (in terms of the Fourier series components), which can be measured at the same time, lies between 50dB and 60dB, depending on the frequency separation between adjacent large and small signal components. This performance is obtained by the use of a comparison technique which ensures that:

(a) Measurements are basically dependent only on the accuracy of a resistive decade attenuator.

(b) The use of the attenuator in the early stages of the noise amplifier ensures that subsequent circuits are not over-loaded.

In practice, the system sensitivity is adjusted with the decade attenuator switched to some convenient figure (i.e. mid-range position reads '30dB') and, by reference to a known level of input signal, which is obtained from a 10kc/s calibrating oscillator. The gain controls are changed to show a convenient deflexion on the thermocouple meter. All subsequent amplitude measurements are made by increasing or decreasing the decade attenuator setting, so that the output signal as first observed on the oscilloscope and next on the thermocouple meter gives the same reading or deflexion as the calibration signal. For example, suppose in a particular instance the attenuator setting was found to be 16dB at a frequency of 53kc/s. The calibrating voltage level is 40μV, so the amplitude of the signal components at 53kc/s is 14dB down on 40μV, i.e. 8μV.

If the instrument is calibrated, as outlined above, with the decade attenuators reading 30dB, it follows that the range of signals which can be measured directly is $\pm$30dB on 40μV. This range can be extended by 40dB in the

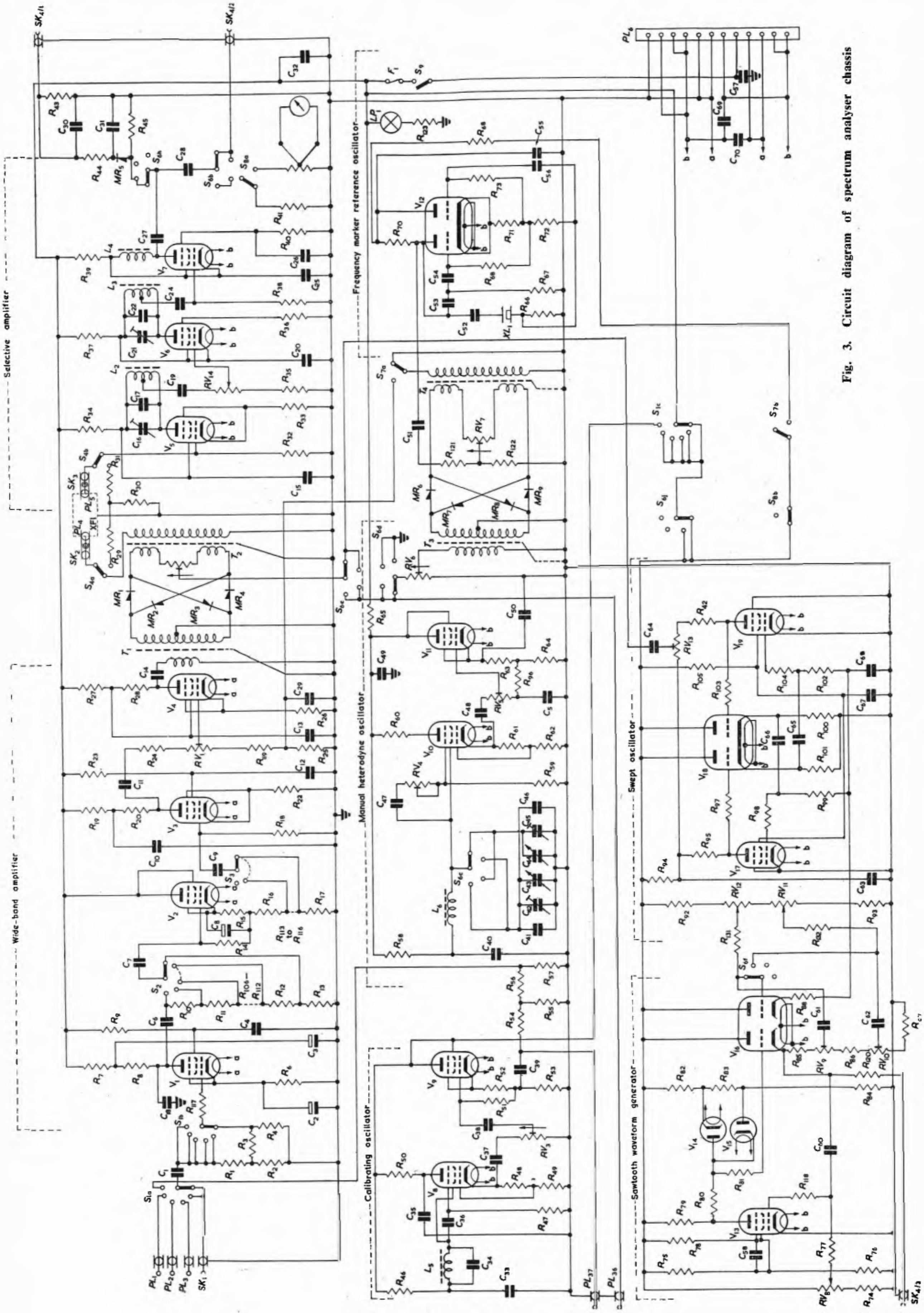


Fig. 3. Circuit diagram of spectrum analyser chassis

positive direction by switching to the '40dB' position on the selector switch and by at least 10dB in the negative direction by calibrating for a lower output, i.e., setting the attenuators to 40dB and noting the new output figure. The range of the instrument will now be -40 to $+60$dB on 40μV, which is 0·4μV to 40mV. Thus, any signal in the range 1 to 90kc/s and of r.m.s. level between 0·4μV and 40mV can be accurately measured by setting the attenuators for the noted output level on the meter or, less accurately, by observing the level on the display unit. In this manner, by adjustment of the sensitivity level, the range may be changed to enable signal levels in the range 0·4μV to 1·0V to be measured.

Signal levels outside the specified frequency range, i.e. 200c/s to 1kc/s and 90 to 95kc/s, may also be checked by referring to the response characteristic of the amplifier. (See Fig. 4.)

Detailed Description

A circuit diagram of the main analyser chassis is shown in Fig. 3, and the operation of the circuits is explained in the following sections which are laid out under headings corresponding to those used in the schematic diagram, Fig. 2.

Input Circuit

The six-way switch S_1 enables the following input signals to be selected, as required.

Position 1.—External input, voltage range approximately 1·0μV to 10·0mV.

Position 2.—External input with 40dB attenuator (resistor network R_1-R_4) inserted. This raises the input level range from 0·1mV to 1V, approximately.

Positions 3, 4 and 5.—Only used for r.f. noise measurements.

Position 6.—Standard 40μV calibrating signal at 10kc/s fixed frequency.

Noise Amplifier

The amplifier stage comprising circuits V_1, V_2, V_3 and V_4 is designed to have a flat response characteristic over the frequency range 500c/s to 90kc/s.

Since the frequency of the i.f. is 100kc/s, it is necessary to ensure sharp cut-off in the amplifier response before this frequency is reached. A typical characteristic is shown in Fig. 4.

It has been mentioned previously that all quantitative measurements are made by employing calibrated attenuators to adjust the measured power out to a standard level, the amplifier having previously been adjusted in sensitivity to indicate the standard level against a known calibrating signal. As with all similar comparative measurements, the accuracy achieved is independent, for practical purposes, of relatively large changes in the absolute sensitivity of the instrument and, therefore, it is possible to make useful measurements at frequencies outside the range of flat amplifier response, i.e. below 500c/s and above 90kc/s, providing a correction is applied to the indicated level to allow for the reduced sensitivity of the amplifier. The required correction at any frequency can be obtained from Fig. 4 and, in practice, it will be found that useful measurements can be made over the frequency range 200c/s to 95kc/s. Beyond these frequencies the use of the instrument is limited not only by the reduction in sensitivity, but also by 'breakthrough' at 100kc/s arising in the signal or in the heterodyne oscillator.

The first two valves in the noise amplifier are chosen for low noise and microphony, and in practice the inherent noise level is approximately 0·2μV, referred to the input terminals out of a source with a resistance of 5kΩ.

The output from the first stage V_1 is connected to an attenuator providing 5 $\times$ 10dB gain steps via S_2 to the grid of the cathode-follower V_2. The load of this stage is a further attenuator providing 10 $\times$ 1dB steps via S_3 to the final amplifier and output stages V_3 and V_4. An uncalibrated variable attenuator RV_1 coupled between these two stages has a range of 6dB and enables the overall sensitivity to be standardized with reference to the calibration signal.

Apart from the usual advantages of employing what is commonly known as a 'substitution' method of measurement, there are even stronger reasons for doing so in this particular case. In particular, the possibility of overloading by line or noise peaks in later stages of the apparatus are strongly reduced. Final measurements are made by means of a thermocouple meter and involve r.m.s. levels of not greater than 1·0V at the output of the selective 70c/s i.f. amplifier, but there is no bandwidth limitation prior to the crystal filter other than that imposed by the noise amplifier response. Thus, it is important to have the means

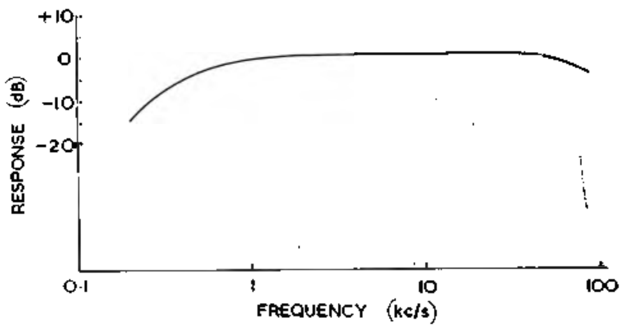


Fig. 4. Noise amplifier response

for limiting the sensitivity to prevent overloading the later stages of the noise amplifier and the following modulator or mixer stage.

Signal Mixer

After amplification, the signal is fed to a mixer circuit supplied with a biasing voltage from a local oscillator with a frequency range from 100 to 195kc/s.

The signal mixer is a double balanced modulator using four crystal diodes in a conventional bridge-ring arrangement. The output stage of the noise amplifier is matched to the input transformer T_1 of the modulator and the output transformer T_2 is matched to the input of the 100kc/s filter XF_1. A double balanced modulator is employed to ensure isolation from 100kc/s noise signal components which would otherwise pass directly through the 100kc/s filter and also to suppress local oscillator signal components which would similarly tend to break through as the local oscillator frequency approached 100kc/s.

Signal Mixer Drive

Depending on whether or not the equipment is being used on 'automatic sweep' (panoramic display) or on 'manual' in conjunction with the modulator milliammeter M_1, the biasing drive to the modulator is derived from either the swept oscillator V_{17}, V_{18}, V_{19} or the manually tuned oscillator V_{10}, V_{11}.

In either case, two frequency ranges are available which, in terms of the difference or lower sideband frequencies, correspond to 0 to 7kc/s and 0 to 100kc/s, and these may be selected by switch S_6, as required.

Manual Oscillator

This comprises a tuned grid, *LC* transitron oscillator V_{10} followed by a buffer stage V_{11}.

With the switch wafer* S_{6b} turned to position 4, the oscillator will tune through the range 100 to 107kc/s and on position 3 the range is 100 to 200kc/s.

In both cases, tuning is capacitive by means of C_{43} and C_{44}. Both are driven from the same shaft and the dial is calibrated in two ranges in terms of the lower sideband or difference frequency of the mixer, i.e. signal frequency, from 0 to 7kc/s and from 0 to 100kc/s.

In addition to the function of supplying the signal mixer drive on 'Manual', the oscillator is used in the 'Auto' condition to generate a frequency marker line (or more correctly, 'blip'), on the panoramic display for frequency calibration purposes. Thus the two frequency ranges are repeated on positions 1 and 2 of S_{6b}.

In order to generate a frequency marker by means of the manual oscillator, it is necessary to employ a further mixer MR_6 to MR_9 which derives its switching frequency from a 100kc/s fixed frequency crystal controlled oscillator V_{12}. The output from this mixer is then coupled into the noise amplifier at the junction of resistors R_{19} and R_{25}. For example, suppose the manual oscillator is tuned to 110kc/s, so that the dial reading is 10kc/s; then, the signal out of the mixer will comprise principal sum and difference components of 210kc/s and 10kc/s together with subsidiary sidebands arising from oscillator harmonics. In practice, these are negligible and the upper sideband is outside the response of the instrument. It follows that the marker signal is a single frequency component which appears as a line superimposed on the signal spectrum of the subsequent panoramic display.

This method of frequency calibration has two very distinct advantages. In the first place, the same oscillator is used both to derive the marker line and subsequently to provide the bias to the signal mixer on the manual tuning. This means that once any particular line of interest in a signal spectrum has been lined up to the frequency marker, then subsequently switched to manual, little or no further adjustment of the tuning control will be necessary in order to obtain the maximum power figure corresponding to the line. Secondly, the accuracy of frequency calibration cannot be affected by factors such as linearity of the frequency base and changes in the spectrum width swept out by the panoramic display circuits.

Swept Oscillator

When the equipment is switched to 'Auto' on positions 1 and 2 of S_6, the manual oscillator drives the marker mixer, as described in the previous section, and the drive to the signal mixer is taken from the 'swept' oscillator which has two tuning ranges, from 100 to 107kc/s and from 100 to 200kc/s.

The swept oscillator comprises a free-running multivibrator, $V_{17}V_{19}$, with cathode-followers V_{18} interposed between the anode and grid feedback circuits in order to maintain the output amplitude constant and fast rise-time over the frequency range. The signal mixer drive is obtained from a variable tap on RV_{13} connected differentially between the anode loads.

Tuning of the multivibrator is affected by varying the grid bias levels symmetrically by means of a sawtooth waveform developed by the sweep generator V_{14} and V_{15}.

V_{13} is an orthodox phantastron time-base with the maximum and minimum voltage excursions determined by limiting action on diodes V_{14} and V_{15}.

The sweep rate is variable by means of RV_3 from (0·5sec/sweep) to (5·0sec/sweep). The spectrum width on the narrow and wide spectrum ranges may be adjusted by means of the potentiometers RV_{10} and RV_9, and the 'zero' adjustments RV_{11} and RV_{12} are preset controls to enable the commencement of the panoramic sweep to be set to 100kc/s on both ranges.

The output from the cathode-follower V_{16} drives the swept oscillator and is also taken through coupling capacitor C_{57} to coaxial connector SK_8 and from this point to the display unit in order to drive the oscilloscope time-base.

100kc/s Amplifier

The output from the signal mixer, which has been described, is connected via switch S_{1a} to the 100kc/s amplifier and output circuits. With the switch in position 1, the input signal is taken through a narrow-band crystal filter XF_1, having a noise bandwidth of 70c/s. The response characteristics of the narrow-band filter determines the selectivity of the whole system. Since one of the functions of the spectrum analyser is to enable noise powers to be measured accurately, the filter has a response curve shaped so that it corresponds very closely to the ideal, rectangular form. A typical curve is shown in Fig. 5.

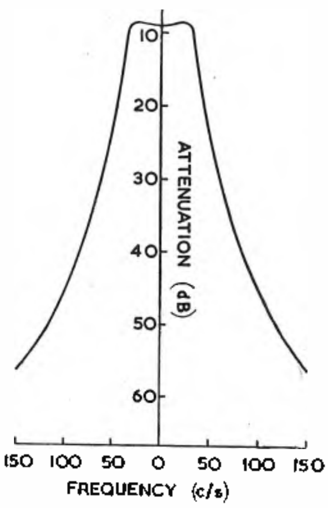


Fig. 5. 70c/s filter characteristic (100kc/s band centre)

For fast spectrum sweep rates, the attenuation introduced by the 70c/s filter is too severe to enable small signal components to be observed. The amplifier, comprising stages V_5, V_6 and V_7 is, therefore, arranged to have a noise power bandwidth of approximately 1kc/s and the 70c/s filter can be switched out of circuit for automatic scan on the 100kc/s range by the switch S_4. The attenuator pad, comprising R_{29}, R_{30} and R_{31}, represents the attenuation which is introduced, by the filter XF_1, and so ensures that the effective gain of the selective amplifier is held constant for both bandwidth conditions.

When the display selector switch S_6 is set to the 'Auto' condition, the output from the 100kc/s amplifier is taken to the *Y* amplifier of the display unit. If, on the other hand, S_6 is turned to the 'Manual' position, the amplifier output is taken to the thermocouple meter M_1. At the same time the signal is connected through to the display unit.

Calibration Oscillator

All amplitude measurements of signal or noise power are made with reference to a 10kc/s signal generated by the calibration oscillator V_3. The output from this oscillator is taken via the calibration level RV_3 through the cathode-follower V_9. The level control is adjusted so that the calibration signal level monitored at socket SK_6 is 400mV which thus ensures that a 40μV signal is developed across R_{57}.

Display Unit

The display unit chassis contains power supplies for the complete equipment, *X* and *Y* amplifiers, the circuits associated with the c.r.t. and a single valve time-base generator for use on 'Monitor'. A circuit diagram is given in Fig. 6.

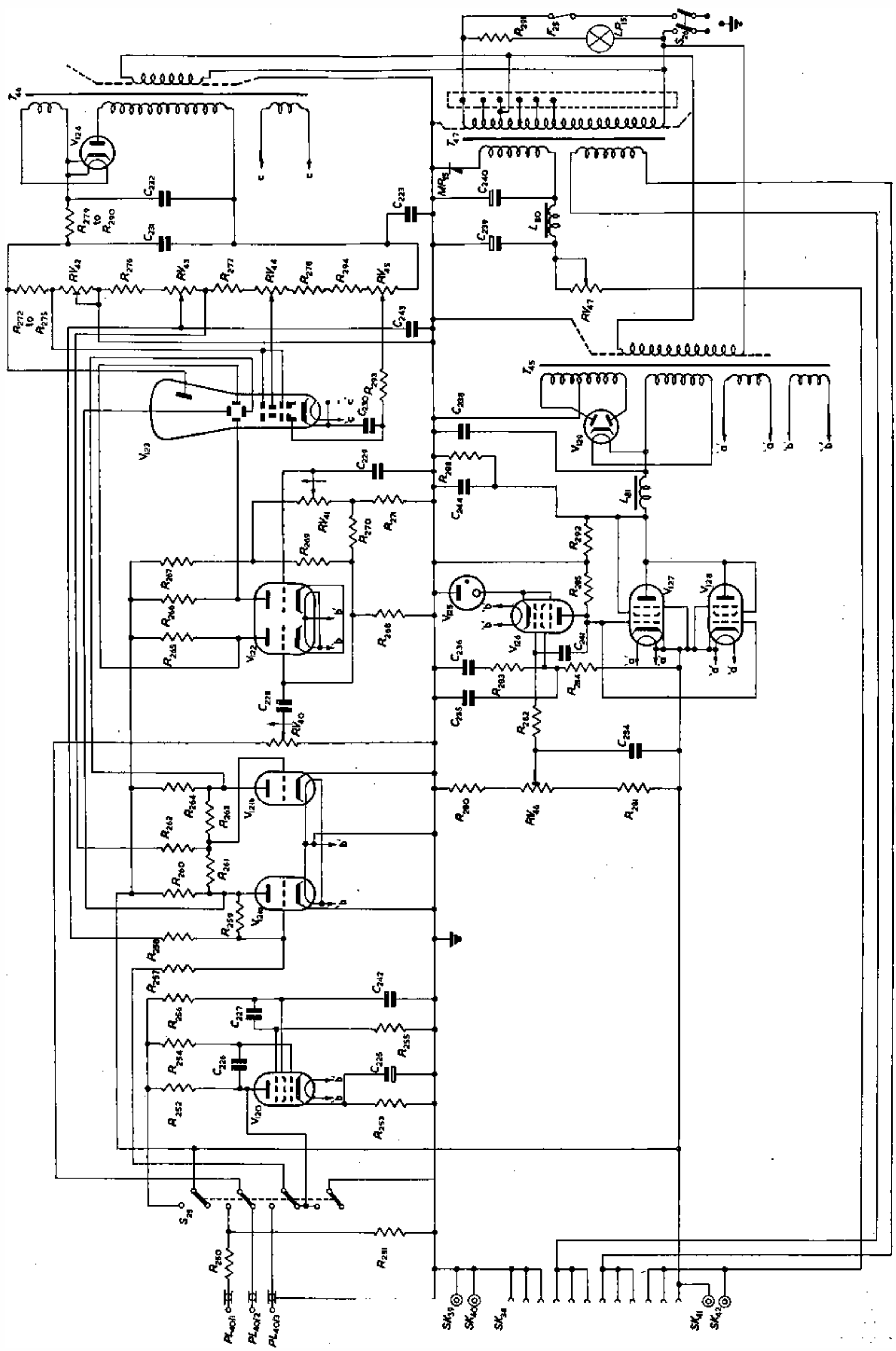


Fig. 6. Circuit diagram of display unit

Operation

SETTING-UP

The instrument, which is illustrated on page 268, is first set-up to operate on the 'Manual' condition and the next operation is to adjust the sensitivity to the required level by turning the input switch to 'Calibrate' and tuning the manual oscillator dial slowly in the region of 10kc/s for maximum output on the oscilloscope trace.

The output level is then adjusted by the decade attenuators to coincide with lines engraved on the viewing graticule, before using the meter, in order to avoid the possibility of damaging the thermocouple. Having thus roughly set up the correct sensitivity, the thermocouple meter is then switched into circuit and set to a reading of 2mA by fine adjustment of the noise amplifier gain. The analyser is then calibrated for an input signal of 40μV, the setting of the decade attenuator is noted and all subsequent measurements are made using this as a basis for calculation.

MEASUREMENT

When the instrument has been set for the required sensitivity, a test signal can be coupled in and the controls altered to the 'Auto' condition. The oscilloscope then provides a spectrum trace which can be adjusted by the attenuators for comfortable visual examination. Frequency 'lines' on the display can be selected, one by one, for evaluation by bringing the frequency 'marker' into operation and tuning the oscillator dial until the marker coincides with the required 'line'. At this point the instrument is switched back to manual operation and the tuning is adjusted to show maximum output on the oscilloscope which, in this condition, functions as a 'visual' voltmeter. The sensitivity is next adjusted to give approximately the required standard output level and then, more accurately, by switching in the meter.

At this point the reader may be prompted to inquire if the use of the meter is not merely duplicating the function of the oscilloscope in the measurement which has been described. In fact, a spectrum comprising simple harmonic components, with only a limited amplitude range, can be examined in this way but the addition of the thermocouple meter ensures that random noise spectra can be determined as accurately and conveniently as line spectra. Also, the meter is more sensitive at very low signal levels.

Applications

The analyser can be used for three principal types of measurement with a number of possible subsidiary roles.

WAVE ANALYSER

When used as a wave analyser for harmonic analysis, the number of harmonics which can be resolved will obviously depend on the fundamental frequency. Assuming the minimum measurement range is, for practical purposes, up to the third harmonic, the instrument can be used to about 30kc/s as shown in Table 1.

The range of measurement for fundamental components is, obviously, up to 95kc/s.

TABLE 1

FUNDAMENTAL FREQUENCY (kc/s)	MAXIMUM MEASURABLE HARMONIC FREQUENCY
0.5	180th
1	90th
5	18th
30	3rd

NOISE ANALYSER

Since the analyser has a constant and accurately known bandwidth over the entire range of frequencies, it can be used for the analysis of random noise spectra, for the determination of noise factor at video frequencies and for distinguishing between harmonic signal components and random noise co-existing in the signal spectra.

AS A SIDEBAND ANALYSER AT R.F.

The modulation spectrum on an r.f. carrier can be measured and analysed if the spectrum analyser is used with a suitable superheterodyne receiver and a linear type of envelope detector at the output for analysing amplitude modulations—or with any suitable f.m. discriminator for f.m. noise analysis. This technique enables harmonic modulation components to be isolated and measured in the presence of noise of considerably greater r.m.s. magnitude.

Acknowledgments

The authors wish to make acknowledgments to James Scott & Co. Ltd for permission to publish this article, also to the Royal Radar Establishment of the Ministry of Supply and to Ferranti Ltd for sponsoring a development programme from which the present system has evolved.

Precision Flow Measurement

De Havilland Propellers Ltd have recently commenced to manufacture, under licence, a precision turbine-type flowmeter, known as the Pottermeter.

The complete Pottermeter system consists of a sensing element and a suitable indicating or recording unit, which can be placed any distance from the sensing element, to which it is connected by a suitable electric cable.

The flow sensing element is based on a novel principle using the kinetic energy of the measured fluid itself to suspend the turbine-type rotor in an axial 'floating' position. Thrust friction is thus eliminated, and the rotor spins freely on plain guides without slippage at a rate governed by and directly proportional to the velocity of the flow through the unit.

Mounted within the body of the standard flowmeter rotor is a small permanent magnet, polarized at right angles to the axis of rotation, and, as the rotor spins, this magnet sets up an alternating current in a coil of wire held within its field. The total number of pulses generated by the sensing element is directly proportional to the total quantity of liquid passing through it.

The frequency of the alternating current produced is directly proportional to the rate of flow of the liquid through the sensing element.

Used in conjunction with an electronic totalizer, the sensing element measures flow with extreme precision. Accuracies of ±one-tenth of 1 per cent, over long periods of operation are not uncommon.

In order to measure rate of flow, the electrical output of the sensing element is applied to an electronic frequency convertor, which produces a d.c. signal proportional to frequency. This signal can be measured as rate of flow by either a milliammeter, an electronic potentiometer, or a high speed recorder. Although the accuracy of a rate measuring system must necessarily depend on the degree of precision of the instrument used to read the output of the frequency convertor, systems with an over-all accuracy of $\pm\frac{1}{4}$ per cent or better of instantaneous flow are available.

The sensing element rotor, spinning freely in the measured fluid, is instantaneously sensitive to flow rate changes. Tests have shown that the rotor responds to a flow rate change in the time required for it to make one revolution. Average response time is about 10msec.

A standard range of indicating equipment is available for use with the sensing elements: these range from simple digital counters, to direct reading counters which read out direct in any required unit of measure. Frequency convertors, leak detection units, and flow proportioning and batch controllers are also available.

A Graphical Analysis of the Blocking Oscillator

(Part 1)

By D. Chambers*, B.Sc.

In this article the blocking oscillator action is described, and the choice of each component in the circuit is studied in detail. Voltage and current waveforms are constructed by a step-by-step method, and the most suitable valve, blocking capacitor, transformer core and the turns ratio are determined for a particular application.

(Voir page 326 pour le résumé en français: Zusammenfassung in deutscher Sprache Seite 333)

BLOCKING oscillators using thermionic valves may be defined as transformer-coupled feedback oscillators in which the valve draws large grid and anode currents for relatively short periods during each cycle. This type of circuit is widely used in radar, television, and other branches of electronics to obtain short duration pulses of relatively high peak power from small receiving type valves. The mode of operation is not, however, so widely understood, partly because of the lack of information about the behaviour of valves in the positive grid region. As a result, design has been largely empirical.

A special valve analyser (Cossor Instruments Model 1070) has been developed by the author for the measurement of characteristics in the positive grid region, and equipment has also been produced to measure the pulse permeability of magnetic materials used in small pulse transformers. With the information thus made available, it has been possible to devise a graphical method of analysing the circuit behaviour, and a design procedure that gives more exact results than the empirical method with a considerable saving of design and development time.

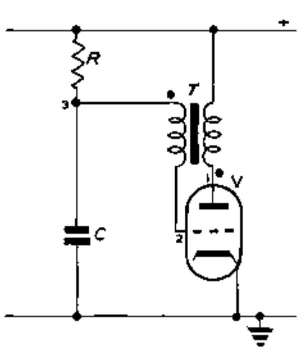


Fig. 1. Basic circuit

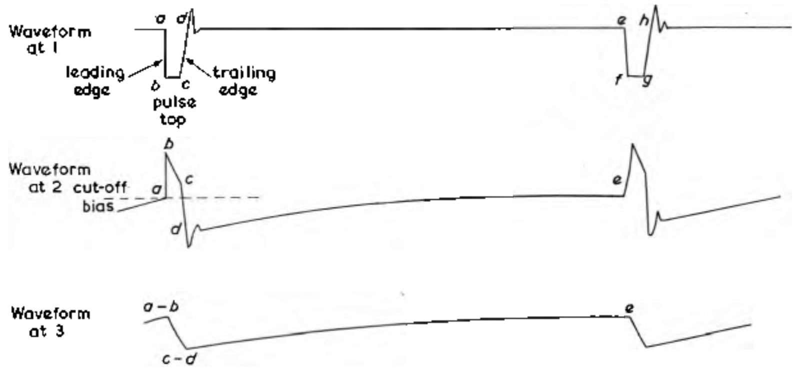


Fig. 2. Oscillograms showing complete cycle of operation

Basic Circuit Operation

The basic circuit is shown in Fig. 1 and comprises four components: valve V, transformer T, capacitor C and resistor R. In addition, there are the anode and grid capacitances of the valve, and the shunt capacitance and leakage inductance between the transformer windings. These elements play an important part in the transient behaviour of the circuit during the rise and fall times of the pulse. As shown, the circuit is free-running, but for triggered operation cut-off bias would be applied to R and the circuit would be triggered by a positive pulse at the grid. The only effect of this would be a speeding up of the start of the pulse.

Fig. 2 shows a complete cycle of operation at various points in the circuit. Only the portion between a and d need be considered in detail here since the overshoot and exponential discharging of the capacitor C between d and e are well understood. The pulse may conveniently be divided into three parts, the leading edge (a to b), the pulse top (b to c), and the trailing edge (c to d).

Certain conditions are assumed at this stage in order that the description of the basic circuit operation may be clarified. The transformer leakage inductance and winding resistance are taken to be zero, the turns ratio is fixed at 1:1 and the rise and fall times are considered to be very short compared with the pulse duration. Later, the circuit operation under practical conditions will be studied.

The Leading Edge

Consider the action as commencing when the grid voltage is rising towards the cut-off level at point a (Fig. 2). As soon as anode current starts to flow, a voltage is produced across the transformer windings which tends to increase the grid voltage. This effect is regenerative and both grid voltage and anode current increase rapidly.

The manner in which the anode and grid voltages and currents build up can be followed from Fig. 3 which shows the average positive-grid characteristics for a commercial type 6J6 triode. By considering the operating conditions of the valve step-by-step during the build-up, it is possible to construct the paths of operation for the anode and grid on the characteristic curves. The assumed conditions imply that C does not have time to charge appreciably during the leading edge so that, for every volt change at the anode, there is an equal and opposite change at the grid.

The path of operation for the anode starts at point A, with $i_a = 0$ and v_a equal to the supply voltage, taken to be 200V. The start of the grid path (A′) occurs at cut-off bias (−7V) and, since the grid current here is also zero, A′ and A are coincident. As the grid voltage rises from cut-

* *Formerly, Cossor Radar & Electronics Ltd.*

off (-7V) to 0V, there is an equal 7V fall at the anode to point B, due to the 1 : 1 transformer coupling between the anode and the grid. The grid current here at B′ is still virtually zero, and the anode current i_a is all flowing into the shunt stray capacitance. (For a transformer with zero leakage inductance, the anode, grid, and transformer shunt capacitances can be lumped together, and the total capacitance (C_s) can be assumed to exist between anode and earth.) The anode voltage is therefore changing at the rate i_a/C_s. These conditions are unstable and lead to a continued increase of grid voltage and anode current, with a consequent fall of anode voltage along the path of operation to the next point C (the 10V rise in grid voltage from B′ to C′ corresponds to an equal drop at the anode).

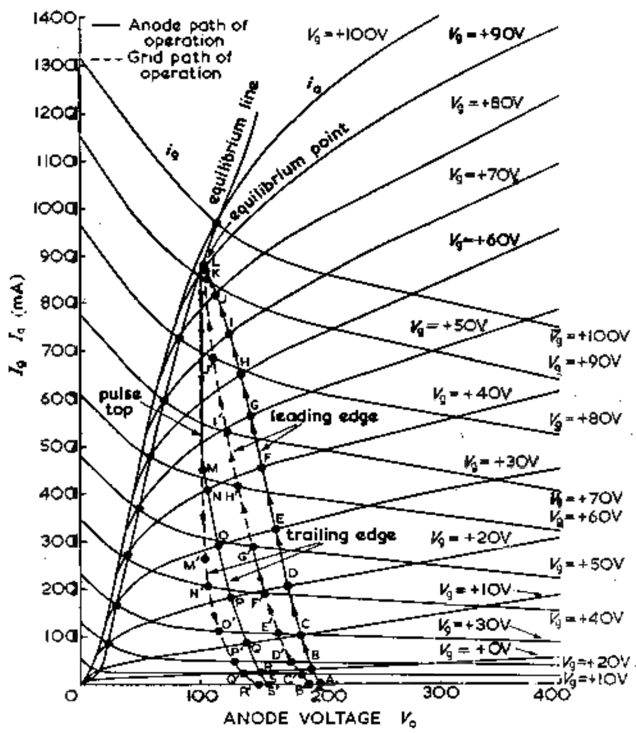


Fig. 3. Positive grid characteristics for average ½ 6J6

At this stage grid current is flowing, and this current is supplied from the anode circuit. Because of the increased grid voltage, the anode current available at C is now able to supply the grid current at C′ and still have a large excess for charging C_s.

The anode and grid paths during the leading edge are constructed step-by-step in the above manner to the point L where the grid current is equal to the available anode current. It will be evident when constructing the paths that, so long as there is an excess of anode current over the grid current, the valve is in an unstable transient state, and the anode voltage must continue to fall. An 'equilibrium line' connecting points of equal anode and grid currents, can be drawn on the valve curves, and the operating conditions of the valve at the end of the leading edge must then be at a point on this line. It should be noted that the grid path of operation A′ to L′ is constructed by dropping a vertical line from each point plotted on the anode path to intersect the corresponding grid current curve. The lengths of the vertical lines represent the available excess current, i_e, and the variation in excess current with grid voltage is plotted (A to L) in Fig. 4. The paths A to L and A′ to L′ in Fig. 3 are independent of circuit parameters and only the value of C_s need now be known to enable the waveforms for the leading edge to be drawn.

The Pulse Top

The pulse top starts at the equilibrium point (L) and extends to the point where the valve can no longer supply the transformer magnetizing current.

Two simultaneous actions control the paths of operation: the grid voltage falls due to grid current i_g charging the blocking capacitor C, while magnetizing current i_m builds up due to the voltage pulse across the transformer. From the valve curves it will be seen that a fall in grid voltage from the equilibrium point initially reduces the grid current more than the resulting reduction in anode current. This produces an excess of anode current which is available for supplying the transformer magnetizing current and keeps the circuit in a stable condition during the pulse top. Fig. 4 shows how the excess current builds up as the grid voltage falls between L and M.

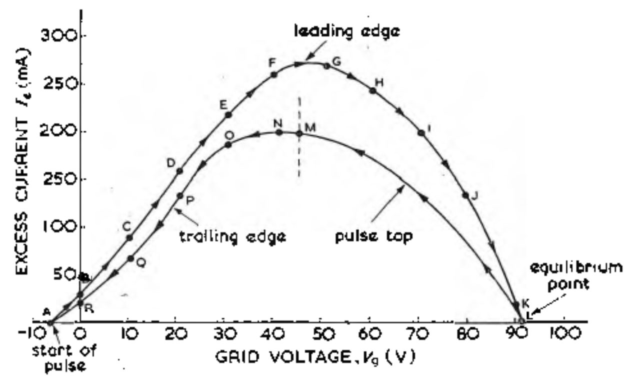


Fig. 4. Variation in excess current during blocking oscillator pulse

When the excess current reaches near maximum value (at M) the valve is unable to supply the increasing magnetizing current at the rate necessary to maintain the voltage across the transformer, and a fall in voltage must then occur across the windings. This point can be regarded as the end of the pulse top and the start of the trailing edge.

The paths of operation L to M and L′ to M′ in Fig. 3 are drawn for the case when the transformer pulse inductance (L_p) and the blocking capacitance (C) have been chosen to produce a flat-topped anode voltage pulse. These conditions exist if the magnetizing current rising in L_p reaches the maximum value of i_e in the same time as the voltage across C changes by an amount equal to the grid voltage drop (between L′ and M′).

A perfectly flat-topped pulse is not always realizable in practice since the anode voltage is constantly adjusting itself during the pulse top to maintain the condition $i_a = i_g + i_m$, and the waveform of rising magnetizing current may not match the excess current obtained along a path dropped vertically from the equilibrium point.

The Trailing Edge

The trailing edge starts when the valve excess current supplying the transformer magnetizing current approaches maximum value and can no longer increase at the rate necessary to maintain the voltage pulse across the transformer. The resulting fall in voltage across the windings however, tends to increase the available excess current from the valve, and also produces capacitive current from the effective shunt capacitance (C_s) of the transformer. The action is therefore influence by degenerative factors that cause the start of the trailing edge to be relatively slow.

The excess current does not begin to fall off until the grid has dropped appreciably (point O in Fig. 4), but when the point is reached on the valve curves where a further reduction in voltage across the windings causes less excess current to be available, the action becomes regenerative. The rate of fall then increases rapidly and the magnetizing current previously supplied by the excess current is replaced by shunt capacitance current.

The paths of operation during the trailing edge M to S and M′ to S′ are drawn in Fig. 3 point by point as described earlier for the leading edge. The paths are again independent of circuit parameters and enable the trailing edge waveform to be plotted for a particular value of C_s.

The Construction of the Waveforms

The paths of operation of Fig. 3 will now be used to construct, point by point, the anode and grid waveforms produced in the basic circuit (Fig. 1) during a blocking oscillator pulse. Although this procedure would not normally be necessary in a practical design, it is useful

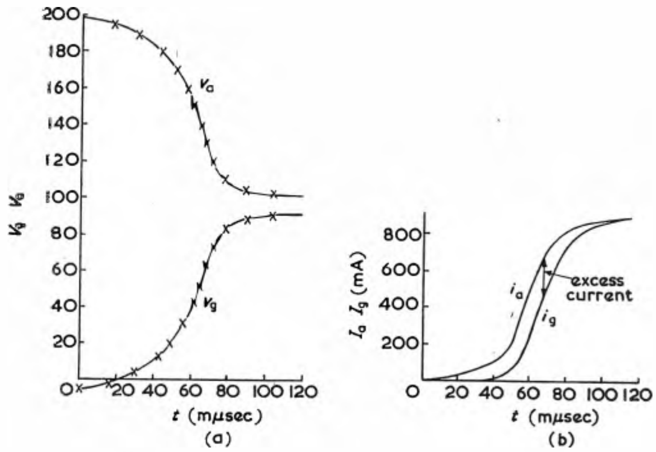


Fig. 5. (a) Anode and grid voltage waveforms during rise-time
(b) Anode and grid current waveforms during rise-time

when first considering the quantitative operation of the blocking oscillator circuit. The values of transformer pulse inductance (L_p) and blocking capacitance (C) will be chosen to give a nominally flat topped 1μsec pulse across the transformer, and one half of a 6J6 with a supply voltage of +200V will be used to generate the pulse.

The Leading Edge

The leading edge of the anode voltage pulse starts at the supply voltage and finishes at the equilibrium point. With the assumptions made earlier, the whole of the excess current ($i_e = i_a - i_g$) plotted in Fig. 4 is used to charge the lumped shunt capacitance C_s. Since the instantaneous rate of fall of the anode voltage is proportional to the instantaneous value of i_e, the waveform on a time scale can be constructed by considering small increments of voltage and current.

Under free-running conditions, the anode voltage starts to fall extremely slowly owing to the excess current being initially zero. In order to construct the pulse shape on a reasonable time scale, it will be assumed that the pulse starts from the point where v_a has already fallen by 2V to 198V giving an excess current of 10mA. While v_a falls a further 3V, i_e increases from 10mA to 25mA and the mean excess current $i_{e(mn)}$ can be taken to be 17·5mA. The time taken for this step is given by $t = (C_s/i_{e(mn)}) \cdot \delta V$, where C_s is the effective shunt capacitance of the trans-

TABLE 1

ANODE VOLTAGE DROP (V)	MEAN EXCESS CURRENT (mA)	TIME TAKEN DURING INCREMENT (mμsec)	TOTAL TIME (mμsec)
198–195	17·5	17·1	17·1
195–190	37·5	13·3	30·4
190–180	75·0	13·3	43·7
180–170	140·0	7·1	50·8
170–160	205·0	4·9	55·7
160–150	250·0	4·0	59·7
150–140	270·0	3·7	63·4
140–130	250·0	4·0	67·4
130–120	210·0	4·8	72·2
120–110	150·0	6·7	78·9
110–105	80·0	6·3	85·2
105–102	20·0	15·0	100·2

former windings plus the valve and wiring capacitances. Typical values would be 80pF for the transformer and 20pF for the strays

$$\text{Substituting values} \quad t = \frac{100 \times 10^{-12} \times 3}{17{\cdot}5 \times 10^{-3}} \text{ sec} = 17{\cdot}1\text{m}\mu\text{sec}$$

Similarly, the time taken for the next drop of 5V is 13·3mμsec. (See Table 1.)

From this point, the curve is constructed by taking 10V steps up to the point where $V_a = 110$V near the pulse top. The last portion of the leading edge requires smaller increments of voltage, because the rate of fall becomes smaller as the equilibrium point is approached. Fig. 5(a) shows the anode and grid voltage waveforms constructed by this method. It will be seen that the rise-time, from 10 per cent to 90 per cent of the peak value, is about 50mμsec. The operating conditions of the valve at the equilibrium point are:

$$v_a = 102\text{V} \quad i_a = 870\text{mA} \quad v_g = +91\text{V} \quad i_g = 870\text{mA}$$

The anode and grid current waveforms shown in Fig. 5(b) are constructed by reading off values directly from the characteristic curves. In a practical circuit, the equilibrium point would be slightly lower because transformer

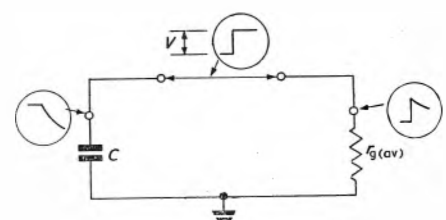


Fig. 6. Equivalent grid circuit during pulse top

leakage inductance and the charging of the blocking capacitor C must be considered. These factors tend to reduce the values of V_a and V_g at the equilibrium point.

When the step-by-step construction is not justified in an actual design, a simple estimation of the rise-time, t_r, can be made by taking the average value of excess current during the leading edge and substituting in the expression $t_r = (C_s/i_{e(mn)})\delta V$. This approximate method gives a rise-time of 72mμsec in the example under consideration, taking $i_{e(mn)}$ as half the maximum value during the rise-time.

The Pulse Top

For a nominally flat-topped pulse, the voltage at the beginning and end of the pulse top are equal and the point of termination occurs where a perpendicular dropped from the equilibrium point L crosses the region of maximum excess current at M to M′ (Figs. 3 and 4, $i_{e(max)}=195$mA at $V_g=+45$V, $V_a=102$V).

The actual shape of the anode waveform during the pulse top is not necessarily flat, but depends upon the valve characteristics and transformer core material. As previously mentioned, the anode voltage is constantly adjusting itself so that the excess current is equal to the magnetizing current, and a flat topped pulse would therefore only be realized if excess current were released along the vertical L to M at the same rate as required by the transformer magnetizing current to maintain the pulse. In most cases, there is some bowing of the pulse top, but the construction of this waveform is tedious and requires a knowledge of the shape of the rise in magnetizing current. It is normally satisfactory to assume that the pulse top at the anode is straight between the extreme points.

The required value of pulse inductance can be found from the fundamental expression:—

$$E = L_p\,(\delta i/\delta t)$$

where E is the voltage pulse across the transformer primary

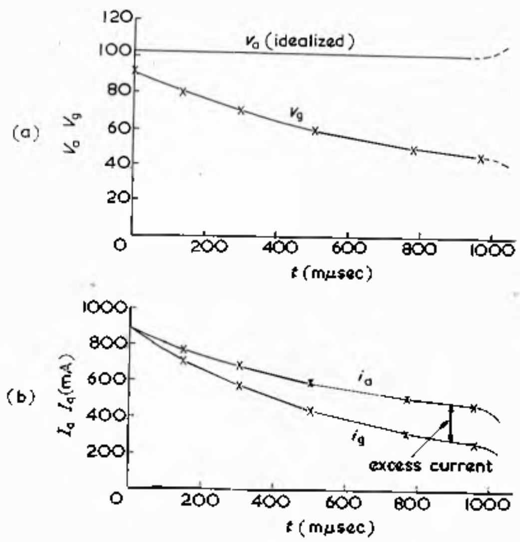


Fig. 7. (a) Anode and grid voltage waveforms during pulse top (b) Anode and grid current waveforms during pulse top

TABLE 2

GRID VOLTAGE DROP (V)	MEAN GRID CURRENT (mA)	TIME INCREMENT (mμsec)	TOTAL TIME (mμsec)
91–80	780	141	141
80–70	620	161	302
70–60	490	204	506
60–50	370	270	776
50–45	280	179	955

and δi is the current rise in time δt. Substituting values derived from Fig. 3, for a 1μsec pulse,

$$L_p = \frac{98 \times 10^{-6}}{0{\cdot}195}\text{H} = 0{\cdot}50\text{mH}$$

The grid voltage waveform falls during the pulse top, and the rate of fall is controlled by the grid current and the value of C. From the paths of operation constructed on the valve curves, the required voltage drop δv at the grid is known (L′ to M′ in Fig. 3) and it is possible to estimate the mean grid current $i_{g(mn)}$. For a flat topped anode pulse any voltage change across C appears directly at the grid, so that C can now be found from the expression ($C = (i_{g(mn)})t_d/\delta v$), where t_d is the required pulse duration.

There are, however, difficulties in accurately estimating the value of $i_{g(mn)}$ since the grid current changes very considerably, and in a non-linear manner, during the pulse top. In practice, $i_{g(mn)}$ can be taken to be the mean of the maximum and minimum values of grid current during the pulse top, since the inaccuracies introduced are still relatively small compared with the total tolerances which must be allowed for in the circuit (i.e.: variations in valve characteristics and transformer pulse inductance). In the example under consideration, the value of C would then be

$$\frac{0{\cdot}56 \times 10^{-6}}{46}\text{F} = 0{\cdot}012\mu\text{F}.$$

A more accurate calculation can be made if C is assumed to charge exponentially. The equivalent grid circuit during the pulse top is shown in Fig. 6. V is the voltage across the transformer minus the valve grid-base (98V − 7V) and $r_{g(av)}$ is the average value of the valve grid resistance. This value, which is far more constant during the pulse top than i_g, may be derived with sufficient accuracy from the valve

TABLE 3

GRID VOLTAGE INCREMENT (V)	MEAN VOLTAGE ACROSS L_p (V)	INCREASE IN i_m DURING INCREMENT (A)	MEAN VALUE OF i_m (TAKEN FROM START OF TRAILING EDGE) (A)	MEAN VALUE OF i_g DURING INCREMENT (A)	CURRENT SUPPLIED BY C_s (A)	MEAN CURRENT IN C_s ($C_s=$ 100pF) (A)	INCREMENTAL TIME (δt) (mμsec)	TOTAL TIME (FROM START OF TRAILING EDGE) (mμsec)	INCREASE IN i_m (FROM START OF TRAILING EDGE) (A)
45–40	95·5	$191\times10^3\times\delta t_1$	$95{\cdot}5\times10^3\times\delta t_1$	195×10^{-3}	$95{\cdot}5\times10^{-3}\times\delta t_1$	$\frac{5\times10^{-10}}{\delta t_1}$	$\delta t_1=72$	72	$13{\cdot}8\times10^{-3}$
40–30	88	$176\times10^3\times\delta t_2$	$(13{\cdot}8\times10^{-3})$ $+(88\times10^3\times\delta t_2)$	192×10^{-3}	$(13{\cdot}8\times10^{-3})$ $+(88\times10^3\times\delta t_2)$ $+(3\times10^{-3})$	$\frac{10\times10^{-10}}{\delta t_2}$	$\delta t_2=48$	120	$22{\cdot}3\times10^{-3}$
30–20	78	$156\times10^3\times\delta t_3$	$(22{\cdot}3\times10^{-3})$ $+(78\times10\times\delta t_3)$	165×10^{-3}	$(22{\cdot}3\times10^{-3})$ $+(78\times10^3\times\delta t_3)$ $+(30\times10^{-3})$	$\frac{10\times10^{-10}}{\delta t_3}$	$\delta t_3=19$	139	$25{\cdot}2\times10^{-3}$
20–10	68	Negligible	$25{\cdot}2\times10^{-3}$	102×10^{-3}	$(25{\cdot}2\times10^{-3})$ $+(93\times10^{-3})$	$\frac{10\times10^{-10}}{\delta t_4}$	$\delta t_4=8$	147	$26{\cdot}3\times10^{-3}$
10–0	58	Negligible	$26{\cdot}3\times10^{-3}$	40×10^{-3}	$26{\cdot}3\times10^{-3}$ $+(155\times10^{-3})$	$\frac{10\times10^{-10}}{\delta t_5}$	$\delta t_5=6$	153	27×10^{-3}

curves by taking the mean of the v_g/i_g ratios at the beginning and end of the pulse top. The voltage change V_t across C due to an applied pulse of amplitude V is given

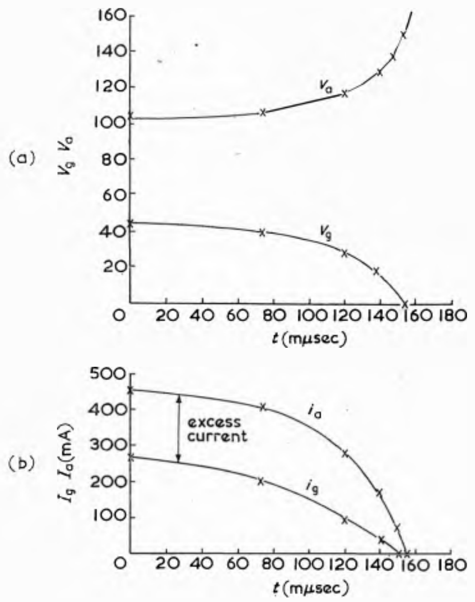


Fig. 8. (a) Anode and grid voltage waveforms during trailing edge (b) Anode and grid current waveforms during trailing edge

by:

$$V_t = V[1 - \exp(-t/r_{g(av)}C)]$$

substituting

$$V_t = 46\text{V} \quad V = 91\text{V} \quad t = 10^{-6}\text{sec and } r_{g(av)} = 139\Omega$$

$$C = 0{\cdot}01\mu\text{F}$$

Using this value of C, the voltage and current waveforms at the anode and grid can be plotted step by step from the relation $\delta t = (C/i_{g(mn)})\,\delta v_g$, where $i_{g(mn)}$ is the mean grid current during a voltage increment δv_g. The figures for each step are given in Table 2 and the voltage and current waveforms are plotted in Figs. 7(a) and 7(b).

The Trailing Edge

After the end of the pulse top at M and M′ (Figs. 3 and 4), the excess current supplying the magnetizing current can no longer increase appreciably. The voltage across the transformer then falls, but the collapse is not instantaneous owing to the shunt capacitance across the windings. The rate of fall at any instant is determined by the value of the capacitive current from C_s required to supply the magnetizing current.

In the first increment shown in Table 3 the excess current is assumed to remain constant, and the incremental time is found by equating the mean increase in magnetizing current during the increment with the current in C_s, ($i_{cs} = C_s(\delta v/\delta t)$). During the second and third increments, the excess current drops, and the average value of the drop

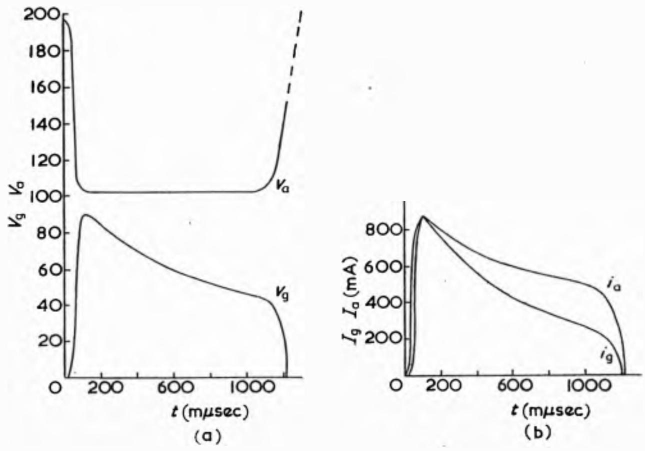


Fig. 9. (a) Anode and grid voltage waveforms during complete pulse (b) Anode and grid current waveforms during complete pulse

is added to the value of magnetizing current before equating with i_{cs}.

The last two increments are of comparatively short duration and the incremental time is calculated, to the nearest millimicrosecond, on the assumption that the increase in magnetizing current is negligible. The current in C_s again supplies the difference between the excess current and the magnetizing current.

When the valve cuts off (s to s′ in Fig. 3), the energy stored in the primary inductance L_p and shunt capacitance C_s produces a damped oscillation with a decrement mainly controlled by the core losses. Figs. 8(a) and 8(b) show the trailing edge waveforms constructed from Table 3, and the composite waveforms of the complete pulse are drawn to a suitable scale in Fig. 9.

In most practical circuits, the fall time of a blocking oscillator pulse is between 100 per cent and 200 per cent longer than the time of rise.

(To be continued)

A Digital Differential Analyser

Corsair is a fully transistorized, transportable digital computer of a type known as a Differential Analyser. Computers falling within this generic title deal with information of a continuously variable nature as normally dealt with by analogue machines, but have the greater inherent accuracy and versatility of digital techniques. The use of digital methods also allows a considerable reduction in volume when compared with analogue machines of equivalent capacity.

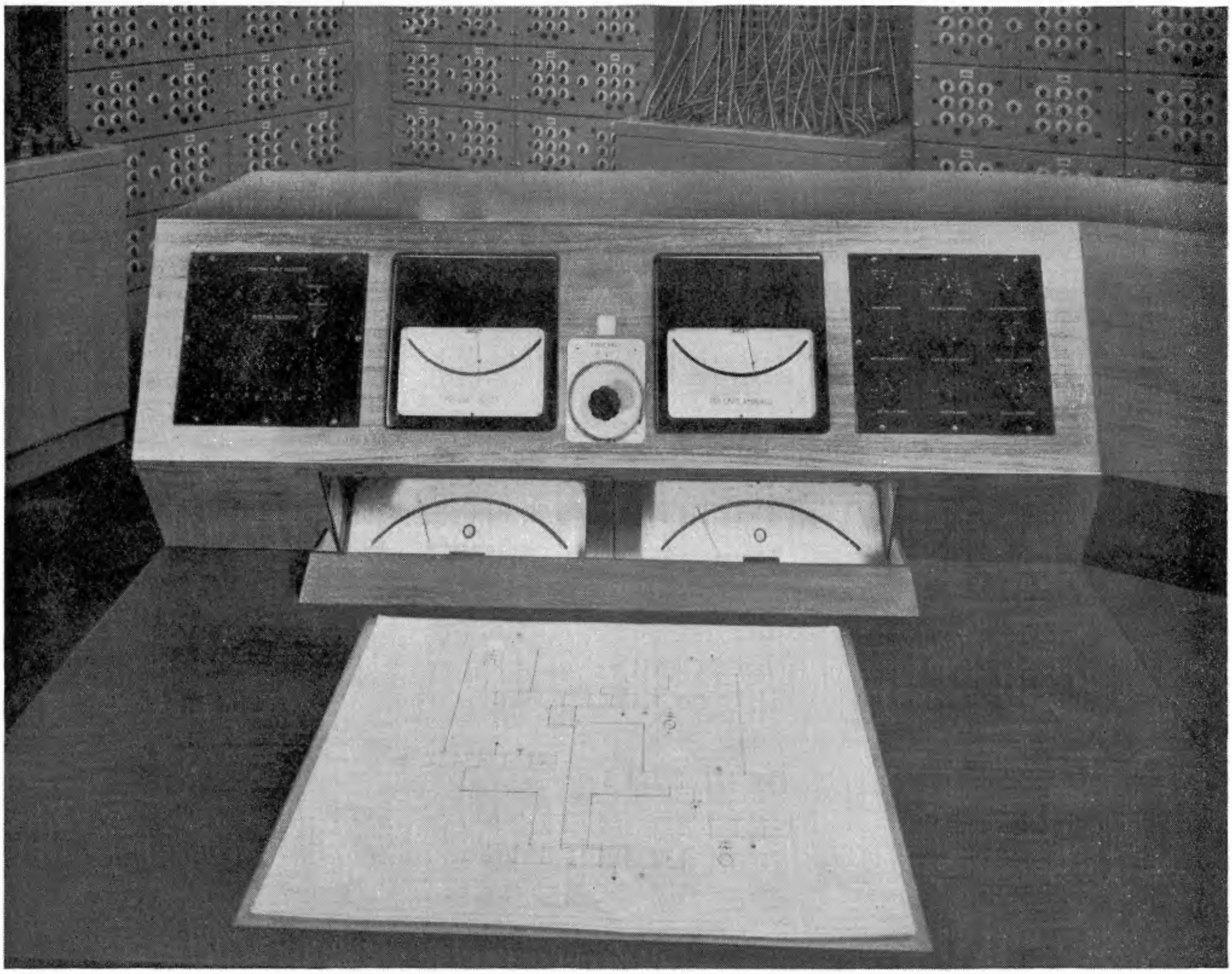
The high-speed section of the machine employs Semiconductors Limited surface barrier transistors in directly-coupled circuits, which do not require the usual coupling and limiting components, and operate from a single 3V supply.

Achieving the ½ to 2Mc/s switching speeds used in these circuits requires the particularly tight control of saturation voltage, input characteristic and high-frequency performance offered by the standard SB240 surface barrier transistor; this degree of control is made possible by the automatic electro-chemical production techniques used by Semiconductors which independently determine base resistivity, base width and emitter and collector diameters. This allows transistor design to be optimized so that the SB240 can operate in these simplified, and therefore highly reliable, directly-coupled circuits at switching speeds as high as 4Mc/s.

Other sections of the computer employ Standard Telephones and Cables Limited TK20 bidirectional transistors in conventional applications. Information is stored in a core matrix which is programmed on a simple patch panel, and in this unit the bidirectional characteristics of the S.T.C. transistors allow them to perform both 'reading' and 'writing' functions.

Corsair was built at the Royal Aircraft Establishment at Farnborough as the result of a small research project. Full use has been made of printed wiring and a packaged form of construction.

This computer and its mode of operation will be fully described in a series of articles to be published in Electronic Engineering in the near future.



Direct Metering for a Transformer Analogue Computer

By J. F. Young*, A.M.I.E.E., A.M.Brit.I.R.E.

Computer voltages are phase shifted and added together by amplifiers and the resultants are supplied to adjustable gain amplifiers. These feed power amplifiers with current feedback which control ammeters and wattmeters. Constant resistance phase shift networks are used with a wattmeter to measure reactive volt-amperes. A calibrated phase shifter is used for phase measurement.

(Voir page 327 pour le résumé en français: Zusammenfassung in deutscher Sprache Seite 333)

THE Witton Network Analyser (WINA)[1] is a transformer analogue type of computer[2,3]. When used for network analysis, the in-phase and quadrature components of all voltages and currents in the system being analysed are represented by 75c/s voltages. All voltages in the actual computer are in phase, and these must be phase shifted and combined in the direct metering section of the computer. They are then used to operate ammeters, voltmeters, wattmeters and VAr-meters. A block diagram of the direct metering arrangements is shown in Fig. 1. Here V_p and V_q are the in-phase and quadrature components of the voltage to be measured, and I_p and I_q are the in-phase and quadrature components of the voltage which represents the current to be measured. Since V_p, V_q, I_p and I_q are all in phase on the computer, it is necessary to phase shift V_q and I_q by 90° before vectorial addition to V_p and I_p in order to obtain the correct phase relationships. After addition, the voltages are amplified and by the use of current feedback they are made to determine the currents in the meter windings.

Meter Phase-Shift Networks

In order to read reactive volt-amperes, the phase of the

The above photograph shows the direct metering section of the computer desk.

* *Formerly General Electric Co. Ltd., Witton.*

currents in the VAr-meter coils must differ by 90° when the currents in the wattmeter coils are in phase. It is convenient if the current in one VAr-meter coil is phase shifted by 45° leading and that in the other is phase shifted by 45° lagging on the current in the corresponding wattmeter coil.

Various arrangements are possible to obtain this phase shift, but many of them resonate at some frequency. Since the meters are supplied from a feedback amplifier, any resonance is undesirable because it might cause the amplifier to oscillate. The circuits described are designed to have no resonances at any frequency, and, in fact, they make the VAr-meter circuit appear to be resistive at all frequencies.

In the circuit diagram of Fig. 2 one VAr-meter coil is shown as an equivalent inductance and series resistance. A capacitor in series with a resistor is connected across the VAr-meter coil. The total impedance of this circuit is:—

$$Z = \frac{R + j\omega CR^2 + j\omega L - \omega^2 LCR}{1 + 2j\omega CR - \omega^2 LC}$$

From this it can be seen that, if

$$L = CR^2$$

then $Z = R$ at all frequencies, and the VAr-meter coil appears in the circuit as a resistance. The current in the coil will lag on the main current by an amount depending on the ratio of reactance (at the operation frequency) to the resistance, so that it will normally be necessary to add sufficient inductance or resistance to make $X = R$ in order to obtain a 45° lag. The value required for an added inductance is shown in Fig. 3.

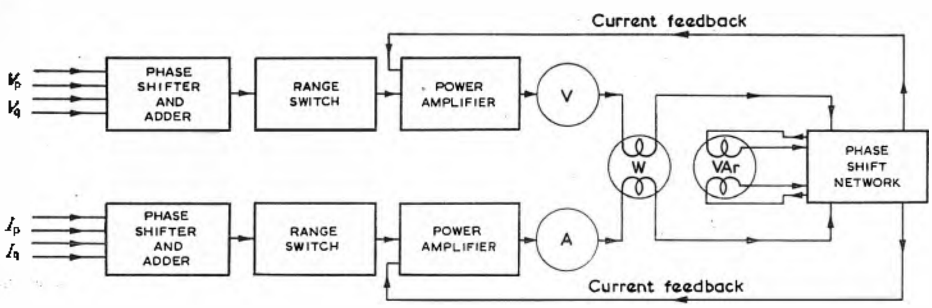


Fig. 1. Arrangement of metering system

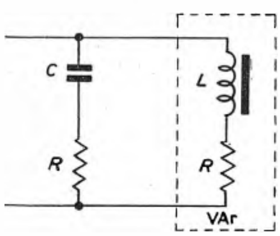


Fig. 2. Basic lagging phase shift circuit

This circuit arrangement is quite well known[4], and has been used in the past to compensate for inductance[5] and, as a constant impedance frequency crossover network, in communications circuits[6,7]. An alternative form of the circuit is shown in Fig. 4(a). This form is derived from the bridge circuit (of Fig. 4(b)), which is balanced at all frequencies[4].

The circuit of Fig. 5 has been derived in a similar way to that given above. The conditions to be fulfilled in order to obtain an impedance which is resistive and independent of frequency are:—

$$L_1C_1 = L_2C_2$$

$$L_2 = C_1R_3^2$$

$$C_2 = \frac{L_1}{R_3^2}$$

$$R_3 + R_3^2/R_2 = R_1$$

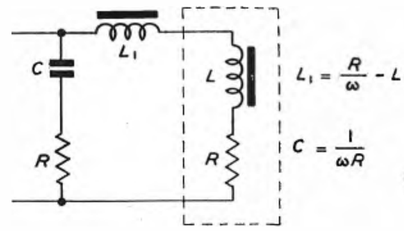


Fig. 3. Added inductance

In addition, at the operating frequency, the admittance of capacitor C_1 must be:—

$$\omega C_1 = \frac{1}{R_2 + \omega L_1}$$

in order to obtain a 45° leading phase shift in the coil current. The effective resistance of the circuit is R_3 at all frequencies. It should be noted that if R_2 is very much greater than R_3, then $R_3 = R_1$ is the correct value to use. This will be true when the Q of the coil is high.

In most cases when these circuits are applied to the compensation of wattmeter coils, the impedances of the two coils will be dissimilar. As an example, the values specified by the makers of the VAr-meter on the Witton Analyser were 150mH and 160Ω for one coil and 3·2mH and 40Ω for the other coil. It is desirable to avoid the necessity for large capacitors in the phase shift circuits since these are bulky and expensive.

In the case of the lag circuit, it is necessary to add inductance as shown in Fig. 3. The capacitance required is

$$C = (L/R^2) = (1/\omega R)$$

where L is the total inductance. Thus the largest possible resistance should be used in such a circuit in order to minimize the size of the capacitor.

In the lead circuit of Fig. 5, one capacitor is required to be

$$C_1 = \frac{1}{\omega R_2 + \omega^2 L_1},$$

and it will be minimized by using the largest possible values of R_2 and L_1. In the same circuit, if R_2 is large

$$C_2 = (L_1/R_3^2 \simeq (L_1/R_1^2)$$

Thus to minimize C_2, inductance L_1 should be small and R_1 as large as possible. If C_2 is expressed as

$$C_2 = \frac{1}{\omega^2 R_1^2 C_1} - \frac{1}{\omega R_1}$$

it is seen that C_1 must be large to keep C_2 small.

With the coils mentioned above the use of the 3·2mH 40Ω coil in the lag circuit would necessitate the use of a 532μF capacitor. The best solution in this case is probably to add a 120Ω series resistor to this coil and use it in the lead circuit. In this way the value of extra inductance required in the lag circuit is minimized and reasonably small values of capacitance are required in all positions. In addition, the effective impedances of the two circuits are equalized. This is of importance since they are supplied from identical amplifiers designed to work into a certain optimum load.

The network analyser is supplied from a 75c/s resistance capacitance oscillator which has a good frequency stability. However long term frequency drifts can be expected and the effects of frequency variations on the phase shift circuits must be considered. Provided the correct values of components are used, both circuits appear resistive at all frequencies, but the phase and amplitude of the currents in the branches containing the meter coils will vary with frequency.

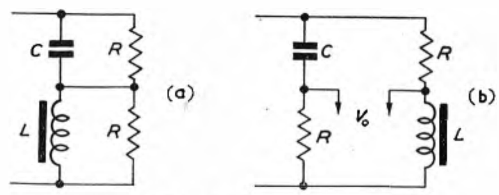


Fig. 4. Alternative connexion

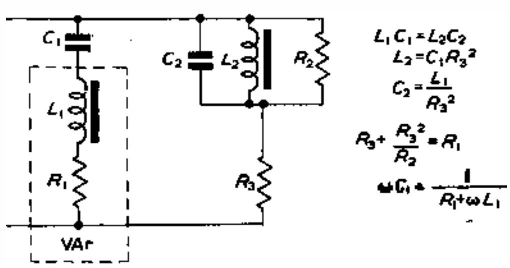


Fig. 5. Leading phase shift circuit

In the case of the lag circuit, it is found that:

$$I_L/I_r = \frac{1}{1 + j(f/f_0)}$$

where I_L is the current in the meter coil

I_r is the total supply current

f is the operating frequency

f_0 is the design frequency.

From this it can be calculated that a 1 per cent frequency error will give a 0·7 per cent error of amplitude and an 18 minute error of phase, and a 0·1 per cent frequency error will give about one-tenth of these errors.

In the case of the lead circuit, the error depends on the magnitude of the inductance of the meter coil. If this is very small, the errors are similar to those obtained with the lag circuit. If L_1 is not small the current ratio is:

$$I_L/I_r = \frac{1}{1 - j(\omega_0/\omega)\left(\frac{\omega^2 LC - 1}{\omega_0^2 LC - 1}\right)}$$

From economic considerations it was decided to use the 3·2mH 40Ω coil with a 120Ω series resistor in the lead circuit. A capacitor of 13μF is required so that LC is 42 × 10^{-9} and $\omega_0^2\, LC$ is 0·009. Hence the errors are very similar to those in the lag circuit.

The lag circuit gives a simple exponential delay of the meter coil current over the main current, with a time-constant of 0·159 of the supply period. There is therefore little delay from this cause in reading the meter when the supply current changes. The response of the lead circuit is similar provided the coil inductance is small, though here there is some initial forcing of the meter which can cause overshooting if the meter is not well damped.

The complete meter circuit is shown in Fig. 6. The 100Ω resistors in each side are used to supply a feedback signal to the current control amplifiers. All resistors shown are wire-wound and have good temperature stability. The capacitors used are of the plastic film type, though these are rather bulky. If a small variation of accuracy with temperature could be accepted, normal capacitors could be used.

The inductors must have the highest possible ratio of reactance to effective resistance. Values of 10 to 12 can be obtained for this ratio with iron cored components. The

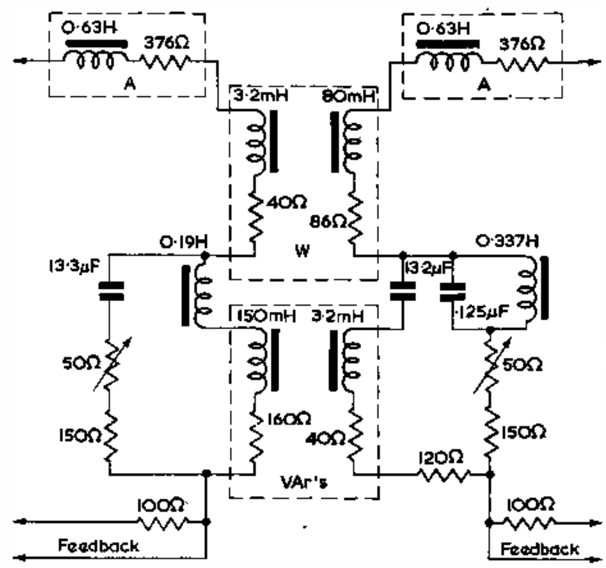


Fig. 6. Complete meter circuits

chokes must not be mounted too closely to the meter since this might cause errors due to inductive pick-up.

Specification

The specification for the complete metering system was as follows:—

Load—Meters having f.s.d. of 43mA.

Accuracy—0·1 per cent of f.s.d. with very small phase shift.

Input—Voltages of 0·1, 0·316, 1·0, 3·16, 10 and 31·6V 75c/s to give full scale deflexion. The actual input is the sum of two voltages V_p and V_q, the latter to be shifted by 90° before addition.

Supplies + 250V and − 100V stable to approximately 0·1 per cent, 6·3V 50c/s.

Input Impedance—Of the order of 1MΩ. Little interaction between the two inputs.

The accuracy requirement involves the use of higher loop gains than is usual. This fact, combined with the relatively low operating frequency of 75c/s, makes the design of the amplifier difficult. Because of this, it was decided to split the problem into two parts:—

(1) An accurate power amplifier giving 43mA output for 10V input.

(2) A pre-amplifier and a mixer to provide range switching and accurate mixing of the two inputs, and to supply 10V f.s.d. to the power amplifier.

The load of one amplifier as described previously is effectively 676Ω in series with 0·63H and that of the other is 722Ω in series with 0·71H. At 75c/s these figures give a power factor of over 0·9. Thus the effective amplifier load is nearly resistive and load inductance causes little interference with the operation of the output stage. The output power required for full scale deflexion is less than 2W. Since an extremely rapid transient response is not required, it is not necessary to use large power valves in the output stage.

An accurate transfer from input voltage to output current is needed, so it is preferable to feed back from the output current rather than from the output voltage as is done in most feedback amplifiers. In this way it is possible to control the output current accurately in amplitude and in phase without the necessity for the extremely accurate matching of load impedances which would otherwise be required. The accuracy of the equipment is unaffected by temperature variations or by random variations of the load parameters provided the resistor used to derive a voltage proportional to load current does not change in value.

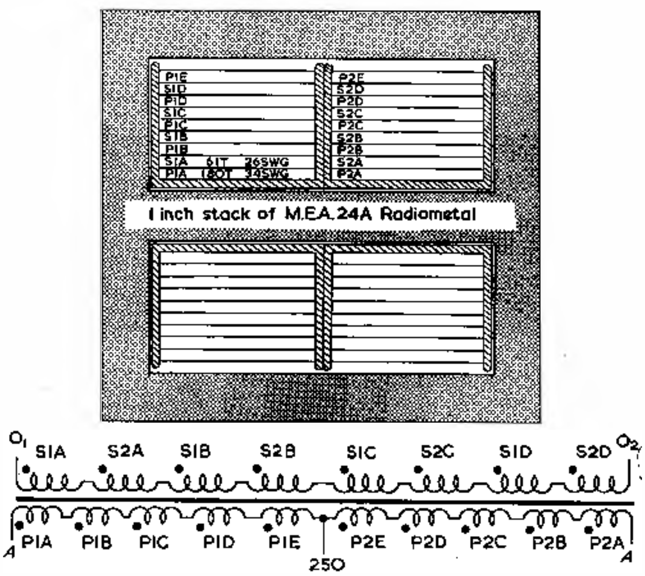


Fig. 7. Output transformer

The meter circuits should preferably be earthed at some point in order to avoid troubles due to pick up of stray fields, and the meters should in any case not be mounted too far from the amplifiers.

It is necessary to design the amplifiers to be reliable and not to require frequent setting up. The circuit used should have all valves and components working well within their ratings, and the accuracy of the equipment should be affected as little as possible by variations of the valve constants. The overall phase shift of the circuit should be small, of the order of a few minutes, and as constant as possible, since two separate channels are compared at the wattmeters. While the supply frequency is fairly constant, the operation of the equipment should not be affected by reasonable changes.

At the same time the circuits should be as simple and inexpensive as possible. Any controls required for initial setting up should be easy to adjust and completely independent in action (i.e. the operation of one control should not require resetting of another control.) In order to achieve this, special circuits are required in the amplifiers, and a fixed gain power amplifier is preferable to one with gain range switching.

The equipment should be free of oscillation or of any tendency to 'ringing'. This is the main problem in the design, since although the amplifiers operate normally at 75c/s, the frequencies at which troubles from oscillation can be expected are less than 1c/s and greater than 10kc/s. Not only must the equipment not oscillate when it is first constructed, but it must have a reasonable tolerance to valve characteristics, so that it does not oscillate after some months of operation. While oscillation would normally be caused by the circuit of the amplifier itself, it can also be caused by poor regulation of the power supply. It is usual to avoid this by decoupling, but this is not possible in the present case because the low operating frequency of 75c/s means that excessive values of capacitance would be required in order to prevent the production of very low frequency oscillations in the main amplifier loop. Where decoupling is absolutely essential, cold-cathode stabilizer tubes must be used instead of capacitors. Elsewhere, reliance must be placed on the power supply, which should have a low output impedance over a wide frequency range.

The equipment should be as little affected as possible by stray pick up, mains hum from heater supplies and by microphonic effects in valves, all of which can cause trouble in fairly high gain amplifiers. These troubles and the possibility of oscillation can be minimized by using only a small number of amplifying stages.

Power Amplifiers

Various output stage circuits in which an output transformer is not required were considered. None of these was considered to be suitable for use with a high gain feedback amplifier for one or more of the following reasons:—

(1) Lack of balance between successive half cycles, producing harmonic distortion and non-linearity.

(2) Presence of a resonant circuit in the feedback amplifier.

(3) Some direct current in the load in addition to the required alternating current.

(4) High voltage power supplies required.

(5) Large capacitors required.

(6) Absence of a suitable point for current feedback.

If an output transformer is used all of these disadvantages are avoided, but the transformer itself must be rather complicated.

The transformer must be designed to have low leakage inductance and low self-capacitance in order to avoid troubles at high frequencies. At the same time, the primary inductance must be as high as possible. In the present case high primary inductance is not required in order to obtain a flat closed loop low frequency response, but rather to obtain a predictable and constant open loop response at fairly low frequencies. There must be good balance between the two halves of a push-pull output transformer in all respects.

The output stage is a simple push-pull pentode stage running in Class-AB_1. The construction of the output transformer is shown in Fig. 7. Both primary and secondary are sectionalized and interwound, the two halves being wound on separate formers. While the construction is complicated, this is unavoidable in view of the wide frequency range which must be considered in order to achieve the required accuracy.

Since the input stage of the amplifier must operate on signals of a millivolt or less, it is essential that it is little affected by mains hum, noise or microphony. It should have a high gain so that there will be little trouble from similar causes in succeeding circuits. The gain should not fall or rise until an extremely low frequency is reached, and if possible the stage should be direct coupled to the following stage. The input should be at earth level and no direct current should be injected via the feedback to the meter circuit. At the same time the circuit should be as simple as possible.

Various circuit arrangements were considered, and the one decided upon as the best compromise is shown in the complete circuit diagram of the power amplifier given in Fig. 8. The Z729 valve is specially designed for this type of service, having low hum and noise levels. Care is required in mounting to avoid microphony, though there are no internal resonances below 1kc/s. A gain of over 90 (39dB) is obtainable with a supply voltage of only 150V and a load of 100kΩ. A 150V cold-cathode stabilizer can, therefore, be used to avoid the necessity for a large decoupling capacitor. In order to eliminate the use of large value cathode by-pass capacitors, bias is obtained from a Zener diode in the cathode circuit. Since direct coupling to the following stage is used, resistive biasing would be preferable. This was in fact tried but trouble was encountered due to charging of the cathode by-pass capacitors on transient overloads.

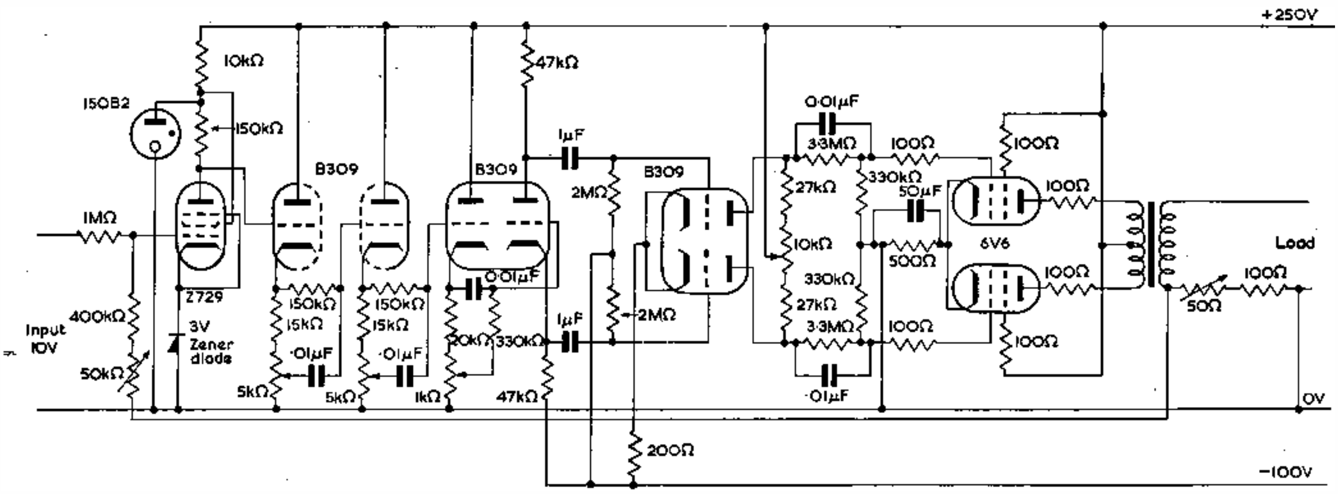


Fig. 8. Power amplifier

The valve pin arrangement is convenient since the grid circuit can be completely screened from the rest of the amplifier, so avoiding the risk of oscillation caused by radiation from the output end.

The stages following the input pentode are simple cathode-followers, their only purpose in the circuit being to provide convenient points for stabilizing adjustments. While the circuit values which affect the stability of the amplifier were known to a reasonable accuracy at the design stage, this accuracy was not great enough to justify the use of fixed stabilizing circuits. Any adjustments provided for stabilizing had to be arranged so that they did not affect the response at 75c/s.

While it is not difficult to vary the low frequency cut-off frequency of a low frequency boost circuit without affecting the high frequency response, it is not usually possible to vary the high frequency cut-off alone. The first two cathode-followers make this adjustment quite easy. In the present case they were adjusted on test to suit the effective time-constant of the load circuit. Some readjustments might be required during the life of the equipment if there are extreme variations of the valve characteristics, but in such a case it would be much better to replace the valves.

The third cathode-follower enables adjustment of the low frequency response of a high frequency boost circuit to be made, without affecting the high frequency cut off frequency. This was adjusted on test to suit the low frequency cut-off of the output transformer, which was inevitably rather dependent on the actual construction of the transformer.

The d.c. level of the output of the third stabilizing stage is rather variable, firstly because of the stabilizing adjustments and secondly because of changes of the characteristics of the previous valves. The operation of the phase-splitter stage must be little affected by this, and it should give a well balanced push-pull output without the use of a complicated arrangement or the need for accurate adjustments. The phase-splitter shown meets these requirements and the accuracy of balance at low frequencies is not affected by variations of the valve characteristics. The phase-splitter is *CR* coupled to the push-pull voltage amplifier. This is the only coupling apart from the output transformer which does not carry d.c., and it is essential that it cuts off at a very low frequency. Consequently, large values of capacitance and resistance are used in the coupling network. Direct coupling would be difficult here, and it has been avoided because the lack of direct negative feedback might cause biasing problems. The valve specified for use as a push-pull voltage amplifier is a medium gain double triode. The circuit used helps to maintain the balance of the system, but it is necessary to adjust the potentiometer in the anode circuit on test in order to compensate for differences between the two anode loads and valve sections. The use of a medium gain rather than a high gain valve here minimizes the effect of the following stage on the voltage amplifier.

Coupling between the voltage amplifier and the output stage is via high frequency boost networks designed to assist in the stabilizing of the amplifier. Because of the direct nature of this coupling, the grids of the output valves are held at about +10V to earth. The cathode resistor of the output stage is consequently larger than is usual, and this improves the performance of the cathode by-pass arrangements.

Careful construction was required in order to minimize the possibility of oscillation caused by pick-up of stray fields. The input stage in particular must be well screened and the output of the amplifier should be well removed from the input. Any controls should be so placed that they are easy to adjust. The layout should be logical, so that servicing and fault finding are simplified. For the same reason the construction and wiring should be as open as possible while minimizing pick-up.

In the layout adopted, a screen passing straight across the input valveholder and connected to pins 2 and 7 and to earth separates the grid circuit of the valve from the rest of the amplifier. The valve itself is contained in an earthed screening can. The under-chassis screen is fitted with a bottom lid electrically connected to earth so that the entire input circuit is contained in an earthed screen having as few gaps as possible. The input to the first stage is brought in via a coaxial cable and plug in order to complete the screening arrangements.

The rest of the amplifier is laid out in line, with the output transformer at the end of the chassis remote from the input circuits. The feedback resistors, which provide a voltage signal proportional to output current, are mounted at the output end of the chassis. The feedback lead is taken from there to the input compartment which contains the series feedback resistors used for gain adjustment. Heater leads are twisted pairs as usual, and multiple connexions to the chassis are avoided. The only layout problem which has been encountered was initial oscillation due to very bad siting of the wrong type of output stage grid stopper resistor.

Pre-Amplifiers

The pre-amplifier, like the power amplifier, must have a high loop gain. However, this is easier to achieve in the pre-amplifier because no output transformer or inductive load is used. The difficulties here were those of providing simple range switching facilities with remote control and of reducing the number of stages required in order to improve the reliability and reduce the size of the circuit.

Examination of the original specification showed that an accuracy of 0·1 per cent was not required on all ranges. For example, this accuracy corresponds to 100μV on the most sensitive range and the accuracy of the rest of the equipment in the computer does not warrant this. Consequently the equipment was designed for an accuracy of 10mV on all ranges, i.e. for 0·1 per cent of the basic range.

Various attempts were made to design a pre-amplifier which mixed the two input voltages, after phase shifting one of them by 90°, and which also had arrangements for range switching. While several possible solutions were found, none of them was reliable and easy to set up. It was therefore decided to divide the duties between two pre-amplifiers, one having mixing facilities and the other being an accurate variable gain amplifier.

The principle of the mixing amplifier is shown in Fig. 9. The output voltage of this arrangement is:—

$$V_o = \frac{1 + j\omega(C_1 + (A+1)C_2)R_1}{-(V_1 + V_2 j\omega C_1 R_1)A}$$

at a frequency where $\omega = 2\pi f$. Now if, at 75c/s one makes $\omega C_1 R_1 = 1$ and one also makes $C_1 = C_2$, then, at 75c/s:—

$$V_o = \frac{-(V_1 + V_2 j)A}{1 + j(A+2)}$$

If the gain of the amplifier at 75c/s is non-complex and very much greater than one, then:—

$$V_o \simeq jV_1 - V_2.$$

The arrangement thus phase shifts V_1 by 90° and adds it to V_2. For an error of 0·1 per cent and a phase shift of less than 4 minutes, the required value of A is about 66dB. There are only two setting up adjustments required for such an amplifier, one of which is the variation of R_1 until $\omega C_1 R_1 = 1$. Once this is done, the amplifier should never require further adjustment provided the values of R_1 and of the capacitors remain constant.

It was not only required that the voltages be mixed with the specified accuracy, but also that the interaction between the two inputs was minimized. The current flowing in resistor R_1 in Fig. 9 is:

$$I_1 = \frac{V_1 j\omega(C_1 + (A+1)C_2) - V_2 j\omega C_1}{1 + j\omega(C_1 + (A+1)C_2)R_1}$$

Under the conditions previously stated, this becomes approximately:

$$I_1 \sim (V_1/R_1) - \frac{V_2}{R_1(A+2)}$$

Thus V_2 has little effect on the current taken from the input V_1. With practical figures, the current flowing in R_1 due

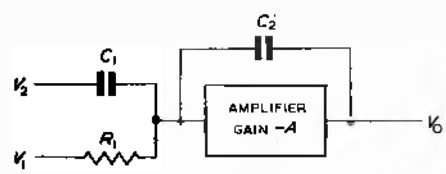


Fig. 9. Basic phase shifter and adder

to V_2 is less than 0·05 per cent of the current due to V_1, when both voltages are equal.

Similarly:

$$I_2 = \frac{V_2(1 + j\omega C_2 R_1(A+1)) - V_1 j\omega C_1}{1 + j\omega(C_1 + (A+1)C_2)R_1}$$

which gives approximately:

$$I_2 = V_2 j\omega C_1 - \frac{V_1 j\omega C_1}{A+2}$$

The arrangement shown is not the only method of using a high gain amplifier to give accurate mixing. An alternative would be to use a resistor to replace C_2. Examination of the frequency response characteristics of such an arrangement shows, however, that harmonics would be amplified more than the fundamental and this might cause a considerable error. The circuit given is to be preferred since it gives no preference to harmonics and in fact it has the effect of filtering out harmonics present in V_1.

The complete mixer circuit diagram is given in Fig. 10. The input stage is similar to that used on the power amplifier described previously, the requirements, being very similar. However, a capacitor has been added in the anode circuit, firstly to give a low frequency boost effect which helps to prevent oscillation, secondly to filter out high frequencies present across the 150B2 cold-cathode stabilizer and thirdly to help to reduce high frequency pick-up present at the input, which might otherwise saturate later stages of the amplifier. The 0·001μF capacitor in the coupling to the next stage, which is a cathode-follower, also assists in all three of these objects. The cathode-follower supplies a stabilizing circuit similar to those in the power amplifier, the values being fixed in this case since all component

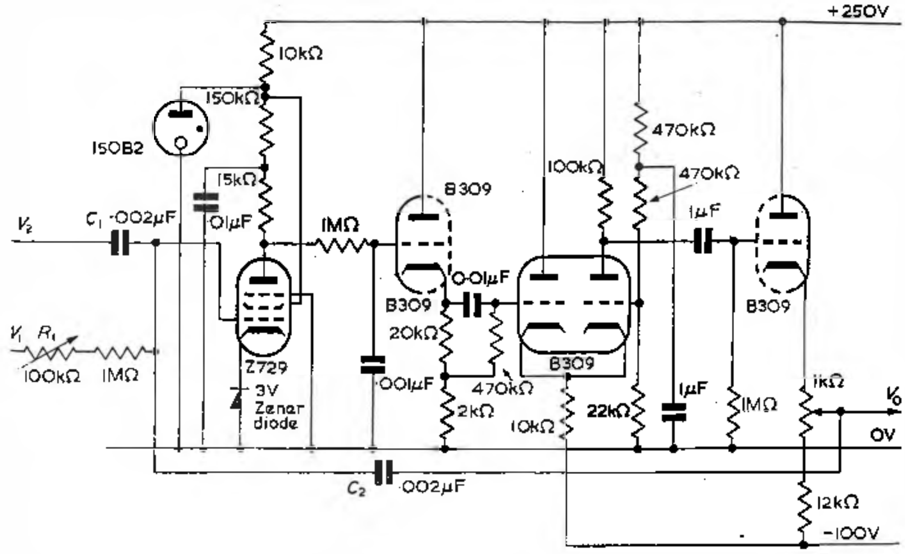


Fig. 10. Phase shifter and adder

values affecting stability are under reasonably close control.

The stage after the cathode-follower is a non-phase-inverting amplifier which is little affected by variations of supply voltage. This is a.c. coupled to the output cathode-follower. The output and feedback terminal is tapped down slightly on the cathode-follower load in order to ensure that the terminal is at earth potential. Although this increases the output impedance slightly, this is of little importance since the load has a high impedance and the overall negative feedback reduces the effect.

The performance is very dependent on the components R_1, C_1 and C_2 and these must have close tolerance and good stability.

The output from the mixing amplifier is the sum of the two inputs, one being phase shifted by 90° before addition. When both inputs have values of 7·07V, the mixer output is 10V. The input to the power amplifier must be 10V for full scale deflexion of the meters, so that the basic gain of the intermediate adjustable gain amplifier must be 1·0. The full list of gain ranges required is then as follows:—

RANGE (10V base)	GAIN
0·316 per unit	0·316
0·1 per unit	1·0
0·0316 per unit	3·16
3·16 per unit	10·0
1·0 per unit	31·6
0·01 per unit	100·0

The per unit terminology used above is common in power engineering, and it is used on the setting switches of the computer. As an example of its use, the range nominated as 0·316 per unit (or p.u.) must give full scale deflexion of the meters when the input is 0·316 × 10V, 10V being the base (or 1·0 p.u.).

Range Switching Amplifiers

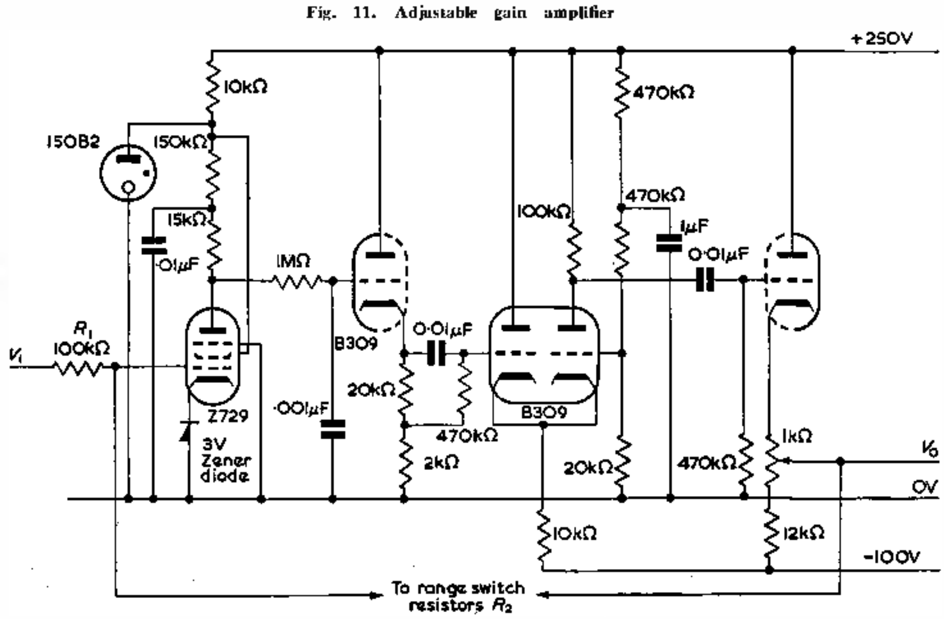
The circuit diagram of the range switching amplifier is shown in Fig. 11. It will be seen to be very similar to the mixer amplifier described above, except that some component values have been changed. Range switching is carried out by relays controlled by a remote switch. It is necessary to set up the ranges on test in conjunction with the mixer and preferably with the power amplifier as well. The performance is very dependent on the stability of the resistors R_1 and R_2, which must be high stability units.

The space available in the control desk for the mixing and range switching amplifiers was very limited, and this influenced the layout adopted. The resulting arrangement does not give easy access for servicing, but this is unavoidable, particularly in the case of the range switching amplifier where the relays take up a great deal of space.

As with the power amplifier, careful construction was necessary in order to minimize the possibility of oscillation caused by stray pick-up. The grid circuits of the input valves are completely screened from the rest of the equipment. On the range switching amplifier the relays were completely enclosed in a metal case since they are in the grid circuit. The grid circuit screens pass across the input valve holders and are connected to pins 2 and 7. The screens are fitted with earthed bottom lids so that the input circuit is completely enclosed, and the valve is fitted with an earthed screening can.

It had been intended originally to use six relays, controlled by the external range switches, to switch to the six gain ranges. However, there was little space available in the control desks for these amplifiers, and it was

Fig. 11. Adjustable gain amplifier



desirable to reduce the number of relays. This has the additional advantage of reducing the size of that part of the adjustable gain amplifier where pick up from stray fields is most likely to cause trouble.

Consideration of the binary nature of relay contacts shows that the number of different resistors which can be switched by n relays is 2^n. Therefore three relays are sufficient for the present purpose since they can switch eight resistors and only six resistors are required for this application. The usual method of connexion of relay contacts for this type of switching is in a 'tree' formation. This requires 2^n contacts on the last relay in the chain. Because only six resistors are required in this case, it has been possible to reduce the maximum number of contacts to four (or two changeover contacts) on any relay. The circuit arrangement is shown at the top of

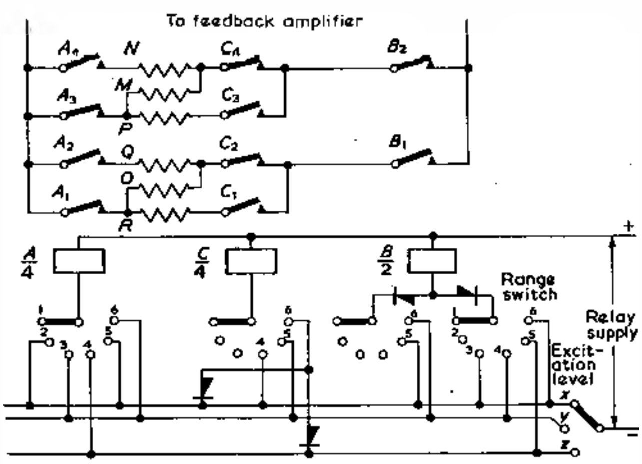


Fig. 12. Range switching circuits

Fig. 12, and the resistors in circuit as a result of the operation of any combination of relays are shown below:

RELAYS OPERATED	RESISTORS IN CIRCUIT
None	*M*
A	*N*
B	*O*
C	*P*
A and *B*	*Q*
B and *C*	*R*
A and *C*	*M*, *N* and *P*
A and *B* and *C*	*O*, *Q* and *R*

From this it can be seen that if relay *A* is never operated at the same time as relay *C*, then the resistors *M*, *N*, *O*, *P*, *Q*, and *R* can be switched into the circuit individually. If these resistors are the feedback resistors in the ajustable gain pre-amplifier, then six independent ranges can be obtained by actuating various combinations of relays, *A*, *B* and *C*.

The relay control circuit is the lower part of Fig. 12. The switch x, y, z, is the excitation level switch which determines the voltage to be regarded as base (1 p.u.) as discussed earlier. Position x corresponds to 1V, y to 3·16V and z to 10V base. The six position, four bank switch is the range switch, which determines the fraction of base voltage which must give full scale deflexion of the meters. The ranges are:

RANGE	1	2	3	4	5	6
P.U. for f.s.d.	31·6	10	3·16	1	0·316	0·1

Voltages corresponding to 31·6 p.u. on a 3·16V base (i.e. 100V) and to 10 p.u. and 31·6 p.u. on a 10V base (i.e. 100 and 316V) cannot occur on the analyser, and the metering system has not been designed to handle such voltages. Consequently it is required that ranges 1 and 2 should be the same when the excitation level switch is on position y (3·16V) and that ranges 1, 2 and 3 should be the same when the excitation level switch is on position z (10V).

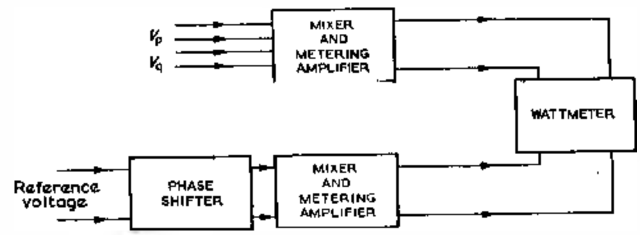


Fig. 13. Arrangement for phase measurement

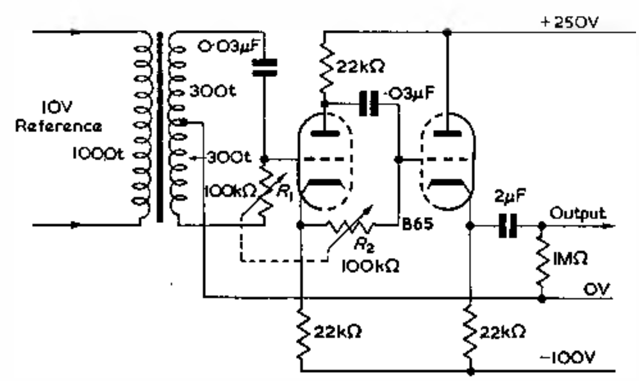


Fig. 14. Calibrated phase shifter

The ranges of all metering amplifiers must be changed when the excitation level is changed, so it has been arranged that the range switches are supplied from x, y, z busbars. An alternative arrangement, which is not described, is possible in which each amplifier requires three banks of the excitation level switch, but only a four bank range switch with no rectifiers.

The relays operated, and the corresponding resistors in circuit, for various switch positions have been tabulated in the preceding column. The resistors therefore correspond to the following amplifier ranges:—

RESISTOR	*M*	*N*	*O*	*P*	*Q*	*R*
F.S.D INPUT	31·6	10	3·16	1·0	0·316	0·1 volts
GAIN	0·316	1·0	3·16	10	31·6	100

The circuit ensures that if all relays fail, or if the relay supply fails, then the amplifier will be switched automatically to the lowest gain range. If any one or two relays fail, the higher gain ranges will not be obtained. Thus the meters are protected in the event of relay failure. This feature was not obtained with the original arrangement using six relays.

The four rectifiers in the circuit have been added to prevent interaction between the various relays on one unit and on other units connected to the x,y,z bars. As an example of the operation of the rectifiers, if the range and excitation level switches are on position 6 and x respectively, then the rectifiers in the C relay range switch circuit will prevent false operation of any relay connected to the z bus bar, while they will allow relay C to operate from the x busbar.

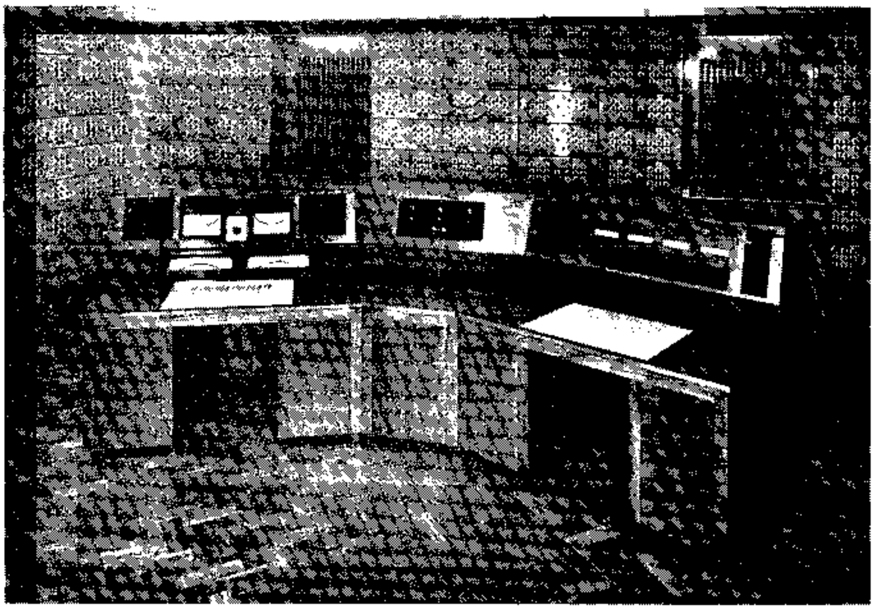


Fig. 15. General view of computer

Phase-Shifter

If two currents differing in phase by 90° are applied to the coils of a wattmeter, the wattmeter will indicate zero regardless of the amplitude of the two currents. This fact is used in the present equipment to determine the phase of any quantity on the computer with respect to a reference voltage. The system used is shown in block diagram form in Fig. 13. The phase of the voltage represented on the computer by V_p and V_q is to be determined with respect to the reference voltage on the left of Fig. 13. The reference voltage is passed through an adjustable phase shifter before being applied to the I_p input of the 'current' metering amplifier. The voltages V_p and V_q are applied to the 'voltage' metering amplifier in the usual way. The phase shifter is then adjusted to give a null reading on the wattmeter, when the required relative phase can be obtained from the calibration of the phase shifter dial.

The circuit diagram of the simple adjustable phase shifter is shown in Fig. 14. A centre-tapped input transformer is supplied from the 75c/s reference source. A CR phase shift circuit across the secondary of this transformer supplies about 3·3V to a phase-splitter V_1, the voltage being adjustable from in phase to about 100° lagging on the reference voltage.

The phase splitter provides about 3·1V at anode and cathode, and the second phase-shifter gives a further variation of phase from in phase to 100° lagging to the grid of the output cathode-follower. Simultaneous adjustment of the two resistors R_1 and R_2 on a common shaft gives a variation of phase of almost 200° at the output. The output voltage is constant at about 3V, so the 'current' amplifier is switched to the 3·16V f.s.d. range for this test.

A general view of the computer with the control desk in the foreground is shown in Fig. 15. The direct metering equipment is on the left of the control desk, the amplifiers being built into the cupboards under the desk. A closer view is shown on page 280. Here the ammeters can be seen vertically mounted on the sloping face of the desk. The wattmeters are horizontally mounted, and the scales can be seen on mirrors from the front of the desk. The push-buttons on the left are used to select the computer unit to be metered, and the switches on the right are used to select such things as meter ranges and polarities of input voltages. The phase-shifter calibrated dial can be seen between the two ammeters. Since the wattmeter and VAr-meter ranges depend on the settings of both the 'current" and the 'voltage' switches, the range in use is determined from both switches and displayed in the small square above the phase-shifter dial.

A short search of the literature to discover similar applications of electronic amplifiers has shown that an appreciable amount of work has been done in this field. Macfayden and Hill[8] developed a simple amplifier for dynamometer instruments, and this has been further developed by Arnold[9], who in addition has used amplifiers feeding both coils of a wattmeter[10], as has Boast[11]. Edmundson[12] has described another arrangement, together with various applications.

Acknowledgments

The writer wishes to thank the General Electric Co. Ltd for permission to publish this article.

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A CONSTANT-RESISTANCE NETWORK

Having the Properties of a Universal Shunt

By E. R. Wigan*, B.Sc., A.M.I.E.E.

By a simple modification, the bridged-T constant-resistance network can be transformed into a constant-resistance universal shunt, adjustable on a decimal basis; a three-decade 1kΩ shunt is described which has a maximum shunting ratio of 11·111/1·0 and is adjustable in steps of 0·001. Between these steps continuous interpolation can be made, the accuracy of which is discussed.

(Voir page 327 pour le résumé en français: Zusammenfassung in deutscher Sprache Seite 334)

THE title set out above is technically correct but is slightly misleading, for this article is not concerned with the conventional 'universal instrument shunt', but rather with networks which are of much higher resistance, and which are unlikely ever to be used in combination with 'measuring instruments' of the usual kind. However, the title has the merit of condensing into a few words the fuller description which might read as follows:—'A constant-resistance network the input/output transfer-constants of which are controlled by a series of decade-type dials which by the sum of their readings indicate the value of the constant directly'. The field of use of networks of this kind will lie in the domain of precision measurement, particularly the measurement of alternating voltage and current. The nature of the 'transfer-constants' mentioned in this extended title is left open, for the same network will serve to measure all of them; input/output voltage; input/output current; input-voltage/output-current, and input-current/output-voltage—in every case it is possible to arrange the circuit so that any one of these quantities will be indicated directly by the sum of the readings of the decade-dials.

It is important to notice that all these ratios are expressed in the form 'in/out' not 'out/in'; in this respect the networks have the properties of an instrument-shunt, the object of which is to provide an input/output current-transfer constant which is usually a round number and is always greater than unity. The networks to be dealt with serve nearly the same purpose, but their primary and novel feature is the continuous, or virtually continuous, adjustment of the 'shunting-ratio'. The importance of this and its application in precision measurements can best be explained in terms of the conventional multi-shunt measuring instrument.

Now the object of fitting a shunt to a measuring instrument is to ensure that, whether the input current be large or small, the meter deflexion shall be as near as possible to full-scale so as to obtain the highest possible reading-accuracy. When the choice of shunts is limited, as is necessarily the case in practice, the meter very rarely reads at full-scale; for if the set of shunts is in the ratio 2:1, the meter will often read near half-scale, and the reading-accuracy falls correspondingly. It would be necessary to make the shunting-ratio continuously variable if the maximum reading accuracy were to be available always.

In the shunts to be described this ideal is achieved; the 'meter' is in fact called upon merely to register at all times its maximum deflexion. It at once becomes obvious that the 'meter' is superfluous—any device will serve which will register the output current (or voltage) delivered by the shunt and will report when it departs from a predetermined or fiducial value. For this purpose some kind of 'null-indicator' can be incorporated. The shunt itself can then act as the 'measuring instrument', the shunting-ratio being adjusted until the null-indicator reads zero.

This reversal of roles has much more than trivial significance, for instance the accuracy of measurement no longer depends upon precise measurement of a pointer deflexion, but upon the restoration of a null-reading—an operation the precision of which depends upon (a) the sensitivity of the indicator—this can be made very high, and (b) the accuracy of the shunting network.

The various ways of arranging for the null-reading indicator will not be dealt with here, the discussion being mainly concerned with the second factor. It will be shown that three-figure accuracy is achievable without difficulty, and that four-figure interpolation can be made at the price of a minor circuit complication.

As the decade-shunt is not intended to be used in the conventional way with a relatively low-resistance instrument, the choice of resistance values is relatively free; it is found convenient to make the 'design-impedance' (R_o) about 1kΩ. The component resistors can then be measured and adjusted without any difficulty.

The Circuit Principle

Consider first the elementary network of Fig. 1 in which there are no arrangements for breaking down the network into decades, but, instead two resistances R_a and R_b are shown, which control the shunting-ratio. This is the well-known Bridged-T circuit, which has the property that, so long as the 'load' resistance—marked $R_{o(2)}$—has the proper value, the input resistance will also be equal to R_o when R_a and R_b are chosen so that:—

$$R_a \, . \, R_b = R_o^2 \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (1)$$

$$\text{moreover, if } R_a / R_b = N^2 \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad (2)$$

then the input/output voltage ratio e_1/e_2 is given by:

$$e_1/e_2 = (N + 1) \quad \ldots\ldots\ldots\ldots \quad (3)$$

* *The British Broadcasting Corporation*

There are three resistances, all of value R_o, in the diagram and they have therefore been distinguished with suffices. Now the circuit can be redrawn (Fig. 1(b)) so that two of them (1) and (3) together with elements R_a and R_b, form a ring, with the third element $R_{o(2)}$ as a diameter. On applying equation (1) to this diagram it can be seen that a voltage applied at e_1 cannot produce any current in $R_{o(2)}$—the 'bridge' is 'balanced'. It would seem therefore that element $R_{o(2)}$ is superfluous, and, so far as transmission of power from left to right in the diagram is concerned, this is indeed true; neither the input-resistance nor the value of e_1/e_2 is affected if element $R_{o(2)}$ is either eliminated or short-circuited*; however $R_{o(2)}$ critically affects the output- resistance of the network. It is to preserve this quantity that $R_{o(2)}$ is included; when it is present, any device, such as a

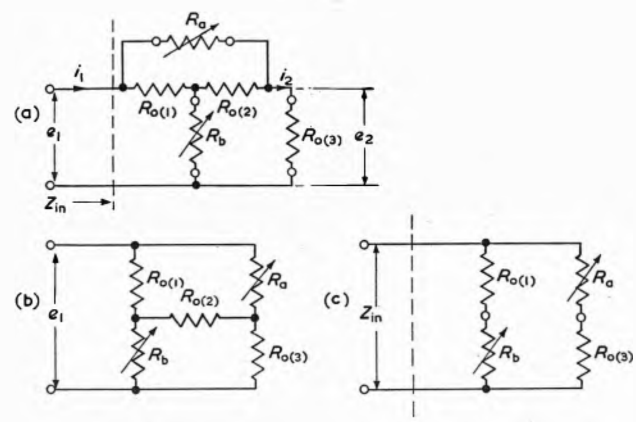

Fig. 1. Bridged-T network

meter, connected at $R_{o(3)}$, will operate as if supplied from a 'source' resistance the value of which can easily be calculated. In the special case that the source of power which produces e_1 has itself an internal resistance equal to R_o, the 'source' resistance feeding the meter is also equal to R_o, and this is constant for all values of N. These points are mentioned here in order to dispose of them and not because they have any bearing upon the operation or calibration of the decade-shunt; they become important only when attention has to be given to the behaviour of any instrument connected to replace $R_{o(3)}$—for instance the output-resistance of the network provides a damping resistance in shunt with the instrument. However such aspects of the networks will not be dealt with here.

Returning now to the case of transmission from left to right in the diagram; since $R_{o(2)}$ can be ignored, it can be taken as infinite—see Fig. 1(c)—and from equation (1) it can be calculated that the input resistance is R_o. Moreover, to arrive at the value of e_1/e_2, one can ignore the left-hand branch—which serves no other purpose than as an impedance-adjuster—and write

$$e_1/e_2 = 1 + R_a/R_o$$

Now from equations (1) and (2):— $\left.\begin{aligned} R_a &= N.R_o \\ R_b &= R_o/N \end{aligned}\right\}$ ..... (4)

$$\text{so } e_1/e_2 = (1 + N), \text{ as in equation (3)}$$

The clue to the method of building a decade-type shunt lies in equations (3) and (4), for it is not difficult to arrange a circuit in which a resistance (R_a) increases in the ratio (N), while the conductance of another element (R_b) increases in the same ratio. The final objective is to arrange the circuit components so that the shunting ratio—which is not equal to N but to $(N + 1)$—shall be indicated by the sum of the readings of the shunt's control-dials.

Before investigating how this can be done the following features of the shunt should be noted:—

(1) The 'load' R_o is a necessary part of the assembly; the equations given above apply only if this load is connected;

(2) Since the input resistance of the shunt is then R_o, and is constant for all values of N:

$$i_1/i_2 = e_1/e_2 = (N + 1)$$

and, moreover,

(3) For the same reason, $e_1/i_2 = (N + 1) \,.\, R_o$ and

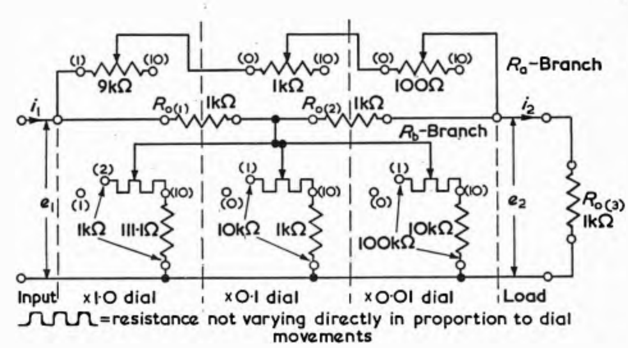

Fig. 2. Schematic diagram

$i_1/e_2 = (N + 1)/R_o$. Thus all the transfer constants are proportional to $(N + 1)$.

The Decade Principle

In Fig. 2 the general arrangement of a three-decade shunt is sketched, using for convenience a design-impedance of 1kΩ. The decades are marked off by the dotted lines and, reading from left to right, are '×1', '×0·1', '×0·01'. In order to simplify the diagram the resistors in both the R_a- and the R_b-branch have been shown as simple rheostats, the latter being suggested as of a log-type. It is to be understood that, in practice, these would be replaced by precision resistances selected by decade-dials; it is however permissible to use rheostats, as shown, for the ultimate decade of a shunt, thus making it possible to vary the shunting-ratio continuously. The bracketed numerals adjacent to the rheostat-symbols indicate the markings which would appear on the decade-dial scales. The 'wipers' of the associated switches would move from left to right in the diagram if the dials were turned away from their zero-marks.

To fulfil equation (4) the resistances in the R_a-element increase linearly with the value of N as each dial is turned to maximum; the resistance of the R_b-elements, on the other hand, decreases. Further, because $(R_a \,.\, R_b)$ = Constant, R_b must become infinite when R_a is zero; the R_b-branch is therefore broken under these conditions—as indicated by the isolated, insulated contacts marked with a bracketed-zero.

In the same way that the resistance R_a increases linearly as any one of the dials is turned, so also the total resistance of each rheostat in the R_a-branch decreases from

* *It may be noticed that these two conditions lead to the two well-known 'series' and 'parallel' constant-resistance networks.*

decade to decade being $10k\Omega$ in the first, $1k\Omega$ in the second and 100Ω in the last case. The inverse process goes on in the R_b-branch of the shunt.

This circuit arrangement, therefore, fulfils the requirements of equations (1) and (2), the resistance of the R_a-branch increasing directly in proportion to N—the number of units of R_o which have been put in circuit by moving the decade dials away from their rest positions—while the resistance of the R_b-branch falls in exactly the same proportion. The value of N which has to be inserted into equation (3) is therefore the number of 'steps' by which the dials have been moved off their rest positions. This being the case the shunting ratio e_1/e_2 will not be, as was intended, the sum of the dial indices, so long as the indices indicate the number of 'steps'; the solution to this difficulty is to mark the dial-scales so that they do not agree with the number of 'steps' but exceed them by unity. There is only one place where this can be done—the 'units'-dial has to be marked to start at '1' instead of starting at '0' as do the other dials; and this can be seen from Fig. 2.

In disposing of this problem, a minor difficulty is created: it does not arise in the case of a shunt of limited maximum ratio, such as that in Fig. 2 which has no dial indexed higher than '×1'; nor would it arise if this dial had 20, or even 100 contacts, for in all these cases the 'hidden' unity-correction is taken care of by the marking of the steps of the '×1' dial. If, however, a '×10' or '×100' dial were added, the reading of the shunt setting would be cumbersome, for a shunting-ratio of, say 100/1 would not be shown by setting the '×10' dial to '10', but by setting up '99' + '1' — for the '1', which is the minimum setting of the '×1' dial, cannot be disposed of.

The solution is to place in tandem with the decade-shunt one or more 'multiplying' attenuating sections, each increasing by 10-times the shunting-ratio shown by the decade-dials; the shunt can then be both adjusted and read without the possibility of overlooking the intrusive-1. The resistance values needed in these additional multipliers are given in Appendix (1).

A Three-Decade-Shunt with a Fourth-Figure Trimmer

For the reasons just stated a decade-shunt should include no dial indexed higher than '×1', multipliers being used for ratios higher than 10:1. The shunt specified in Table 1 can be set up in units of 1·0, 0·1 and 0·01 using the decade-dials; the 'trimmer unit' allows the ratio to be varied continuously over the range 0·01. The design of this unit is dealt with in detail in Appendix (2), and it is sufficient to mention here that it consists of two ganged rheostats as sketched in Fig. 2, one being linear and the other of a log-type. Luckily the main influence of the trimmer rheostats is confined to the linear element in the R_a-branch of the shunt—this is very easily calibrated *in situ*; the R_b-element has much less effect, and a crude approximation to the ideal inverse relationship to the R_a-element serves, even for use in a high-precision shunt.

Table 1 defines the number of ohms which must be connected in the R_a- and R_b-branches of the completed shunt when the various dials are operated. All the R_a-elements in the Table are connected in series (including the R_a-element of the trimmer) and all the R_b-elements are connected in parallel, following the sketch-diagram, Fig. 2. For convenience in building alternative forms of the shunt with a different resistance, the Table is based upon a design-resistance, R_o, of $1k\Omega$; if a value of R_o which is K-times larger is called for, every resistance in the Table has to be multiplied by K.

The 'trimmer-unit' which is to provide for continuous adjustment of the shunting ratio should have a linear 10Ω rheostat in the R_a-branch, and a log-law rheostat of about $1M\Omega$ in the R_b-branch. Alternatives are possible and are dealt with in Appendix (3).

Field of Application

The continuously variable shunt[1] has so far been applied only in alternating current and voltage measurements[2] in which a buffer amplifier has been used to isolate the network from the input signals; however it seems likely that it can be used with advantage wherever measurements, either of a.c. or d.c. have to be made with pointer-instruments. There may, for example, be a double advantage in using the instrument at full-scale deflexion: not only is there a large deflexion to observe, so that fractional changes are easily noticed, but it may happen also that the measurement system is such that, at full deflexion, the rate of change of deflexion with input signal is a maximum; this applies to instruments with 'square-law' scaling.

Again, in certain measurement techniques which may be purely electronic, there is an advantage in keeping relatively constant the signal level at the input of the measuring apparatus (e.g. to maintain a high signal-to-noise ratio); a decade-shunt may then be placed ahead of the apparatus, which is then allowed to operate under optimum conditions irrespective of the input signal-level.

No doubt the design-resistance of $1k\Omega$ used in the previous section will not be universally suitable, but there is no difficulty in changing it, nor would it seem difficult to build and operate shunts several orders of ten lower, and possibly an order of ten higher in resistance.

A constant resistance network of the type described in this article is a primary component of the Muirhead-Wigan Precision R.M.S. Decade Voltmeter type D930A. The design resistance is $1k\Omega$ and there are four decade

TABLE 1

DECADE RANGE	DIAL INDEX-MARK	SWITCH STEP-NUMBER	RESISTANCE SELECTED	
			R_a (Ω)	R_b ($k\Omega$)
'×1'	1	Rest-position	0·0	∞
	2	1	1000·0	1·0000
(Ten switch studs)	3	2	2000·0	0·5000
	4	3	3000·0	0·3333
	—	—	—	—
	9	8	8000·0	0·1250
	10	9	9000·0	0·1111
'×0·1'	0	Rest-position	0·0	∞
	1	1	100·0	10·00
(Eleven switch studs)	2	2	200·0	5·00
	—	—	—	—
	9	9	900·0	1·111
	10	10	1000·0	1·000
'×0·01'	0	Rest-position	0·0	∞
(Eleven switch studs)	1	1	10·0	100·0
	2	2	20·0	50·0
	—	—	—	—
	9	9	90·0	11·11
	10	10	100·0	10·00

N.B.—The tolerances permissible on these resistance values are discussed in Appendix (2).

dials reading from ' ×1 ' to ' ×0·001 '; the associated null indicator is calibrated to allow interpolation between the steps on the final dial.

Acknowledgments

Shunts of the decade type were designed in the course of an investigation at the BBC Engineering Research Laboratories, Kingswood Warren, Tadworth; acknowledgments are made to the Director of Engineering for permission to publish this material.

APPENDIX

(1) 'MULTIPLIER'-SECTIONS FOR A DECADE-SHUNT

The shunting-ratio, shown by the sum of the indices of the shunt, may be multiplied by any factor M by connecting, in tandem with the shunt, a set of fixed resistances as shown in Figs. 3 (a) and (b); it is immaterial whether they are connected ahead of or behind the shunt; but in the latter case the multiplier-section is inserted between the shunt and the load resistance, $R_{o(3)}$.

If it is desired to increase the maximum available shunting ratio from 11·11 (as shown in Fig. 2) to 111·1, one section only is needed, with $M=10$.

To increase the ratio to a maximum of 1111·0, a second similar section could be added, or alternatively a section with $M=100$ could be substituted. On the whole the circuit of Fig. 3 (a) would seem to be the easier to build as the shunt element is rather larger and therefore easier to adjust precisely. For the same reason two similar sections with $M=10$ are to be preferred to one with $M=100$.

(2) ERRORS PERMISSIBLE IN THE R_a- AND R_b-ELEMENTS

The precision with which the component resistors in a decade-shunt have to be adjusted depends upon their position in the circuit. In general the R_b-elements can have wider tolerances than the R_a-elements; this is because the R_b-element of the shunt is concerned, not with the transfer of voltage or current from one end of the shunt to the other, but with the maintenance of the input resistance.

The primary purpose of the shunt is to provide a known voltage- or current-transfer constant, and the input resistance is of less importance.

However, there is a second reason for a wider tolerance on the shunt element resistance; it arises from the presence of three fixed resistances, R_o, in the network; in the following discussion it is presumed that these will have been set to the correct value within very close tolerances; then, from Fig. 1 (b) it can be seen that writing Z_{in} as the input resistance:—

$$1/Z_{in}=1/(R_o+R_a)+1/(R_o+R_b) \qquad \ldots\ldots\ldots\ldots (6)$$

It will be assumed that the values of R_o and R_a are precise and that R_b is in error; being too high by the fractional ratio β, so that, instead of having the proper value called for by equation (4):—

$$R_b=(R_o/N)(1+\beta) \qquad \ldots\ldots\ldots\ldots\ldots\ldots (7)$$

whence, from the text:—

$R_o/Z_{in}=1/(1+N)+1/(1+(1+\beta)/N)$ which, for $\beta\ll 1$, reduces to:—

$$R_o/Z_{in}=1-\beta(N)/(1+N)^2 \qquad \ldots\ldots\ldots\ldots\ldots\ldots (8a)$$

For some purposes one can best use the number indicated by the dial indices, namely $(N+1)$ which will be referred to as D; then an alternative form of equation (8a) is:—

$$R_o/Z_{in}=1-(\beta/D)(1-1/D) \qquad \ldots\ldots\ldots\ldots\ldots\ldots (8b)$$

Now the correct value of Z_{in} is R_o, so the term in brackets represents the fractional error in Z_{in}; which is easily shown to be a maximum when $D=2·0$ and is then equal to $\beta/4$. For all larger or smaller values of D the error will be less than this, falling to zero when $D=1$, the minimum value available from the shunt.

Any error in the resistance in the R_b-branch of the shunt is therefore reduced in the ratio 1:4 at the outset; moreover R_b in the above equations implies, not the individual resistance in a single decade of the shunt, but the totality of shunt elements in all decades.

It can of course happen that R_b is represented by the shunt element of one decade alone, for instance if only the ' ×1 '-dial is operated. It follows that, if an accuracy of 0·1 per cent is called for in Z_{in}, the R_b-element controlled by this dial must be adjusted to at least 0·4 per cent; Z_{in} will then be within much less than 0·1 per cent of the correct value, R_o, for all values of D which are significantly greater or less than 2·0.

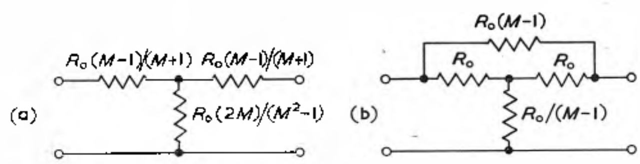

Fig. 3. Multiplying sections *(Multiplication factor is M)*

In the case when the setting D lies between 1 and 2, the only resistors included in the R_b-branch of the shunt lie in the second and lower decades and, again taking the most unfavourable case which assumes that the second but not the lower dials will show a reading, it is found that 0·4 per cent accuracy is more than sufficient for the associated resistors. Finally the third-decade R_b-resistances can be allowed a tolerance of 1 per cent. For in the most unfavourable case—i.e. when $D=2·09$, so that the third dial contributes a maximum of conductance to the R_b-branch of the shunt, while the first dial is set at 2·0, where Z_{in} is most sensitive to errors in the R_b-branch—it is found that the 1kΩ contributed by the first dial to the R_b-branch swamps the error in the third dial in the ratio of 10:1 and the value of β becomes 0·001, and the final error in Z_{in} is 0·025 per cent.

The resistance in the R_a-branch has to be adjusted to 0·1 per cent at all points on the first dial and at the higher end of the second dial, while 0·5 to 1·0 per cent will serve for the last dial. There is in this case no maximum at $D=2·0$.

It becomes very tedious to derive the errors in shunting-ratio and the Z_{in} if the elements $R_{o(1)-(2)-(3)}$ are incorrect, and it has to be assumed that their errors are an order smaller than any others dealt with above.

It is reckoned that if the elements of the shunt fulfil the error-specifications given above the random variation of the residual errors will ensure that the overall error due to all causes is extremely unlikely to exceed 0·1 per cent—for, because the specification calls for errors to be less than 0·1 per cent (but without consideration of sign), there will be some cancellation, due to positive and negative errors combining when more than one decade is in use.

(3) The Design of a 'Trimmer' Element for a Decade-Shunt

Permissible Approximations

The exact resistance values for a trimmer-element for the shunt described in Table 1 are given by equation (4), substituting for N values lying between zero and 0·01. The maximum value of the R_a-element—which has to be a 'linear' rheostat—will be 10Ω; but it is not possible to provide a resistance in the R_b-branch which will agree

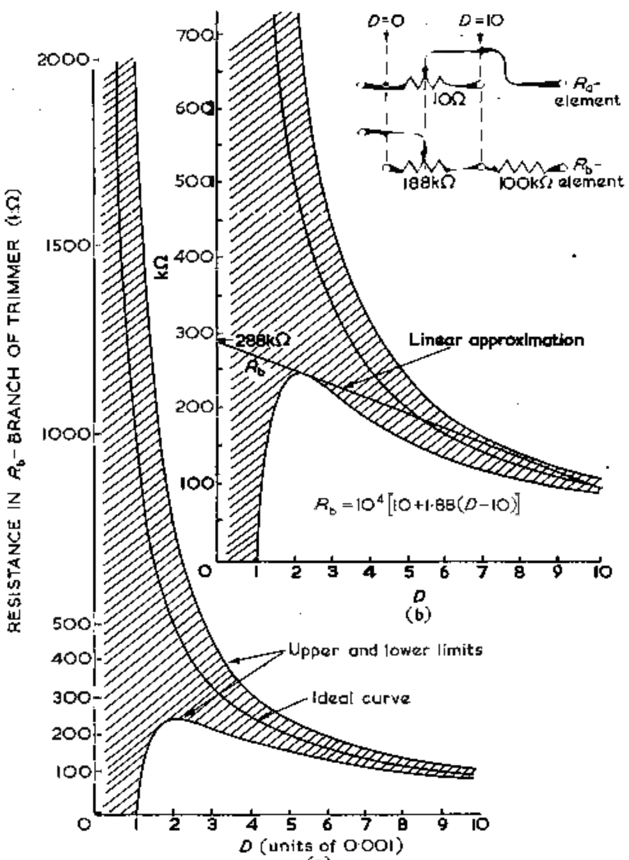

Fig. 4. Solution with a linear rheostat in the R_b-branch

with the equation, for although it is possible to make R_b infinite, when R_a reaches zero, it is not possible to make a smooth transition from infinity to some finite value. An approximation to the ideal value of R_b has obviously to be made in the region of $N=0{\cdot}001$.

From equation (8a) of the previous Appendix, one has, writing the error of Z_{in} as 1/1 000 to agree with the other decades:—

$$1/1\,000 = \beta(N)/(1+N)^2 \qquad (9)$$

Now, if the R_b-element is in error, it will have its maximum influence upon Z_{in} when no other R_a and R_b-resistances are in circuit—that is to say when none of the dials are off their rest-positions except the trimmer; N therefore will lie between 0 and 0·01. Let the decimal-ratio indicated by the trimmer-dial be t; then from equation (9) substituting t for N, and ignoring $(t)^2$

$$\beta = (1/1\,000)(2 + 1/t) \qquad (10)$$

The quantity t having a maximum value of 0·01, the term $1/t$ controls the right-hand side of this equation, and approximately:—

$$\beta = 1/(1\,000 \,.\, t) \qquad (11)$$

which gives the fractional error in the R_b-resistance which can be allowed before Z_{in} departs by more than 0·1 per cent from the ideal value $(R_o)=1\text{k}\Omega$. It is convenient to convert the fractional error into the error in ohms: using equation (7):—

$$R_b = (R_o/N)(1+\beta) \text{ in which } R_o \text{ is } 1\text{k}\Omega$$

one can abstract ΔR_b as the number of ohms error which can be allowed; from equation (1):—

$$\Delta R_b = \beta(R_o/N) = \beta(R_o/t) \text{ in this case.}$$

And from equation (11):—

$$\Delta R_b = (1/t)^2 \qquad (12)$$

A Solution with a Linear Rheostat in the Rb-Branch

In Fig. 4 three curves have been drawn corresponding to

(1) $R_b' = R_o/t$ the ideal value for R_b.

(2) $R_b' + (\text{tolerance}) = R_o(1/t + 1/t^2)$

(3) $R_b' - (\text{tolerance}) = R_o(1/t - 1/t^2)$

Any curve which lies within the shaded area will lead to values of resistance in the R_b-element which are close enough to the ideal value to ensure that Z_{in} will be within 0·1 per cent of the correct value, namely $R_o = 1\text{k}\Omega$.

It can be seen from the figure that it is possible to lay a straight line between the upper and lower tolerance-curves. A purely 'linear' rheostat can therefore be used in the R_b-branch of the trimmer-unit. It must be connected so that the resistance conforms to:—

$$R_{b(\text{linear})} = 10^4 \,.\, (10 + 1{\cdot}88(D - 10))\Omega \qquad (13)$$

the quantity D being the number of units of 0·001 shown by the control-dial pointer of the trimmer-unit.

The physical realization of this formula is a fixed resistance of 100kΩ in series with a rheostat of total resistance 188kΩ, the R_b-rheostat being coupled to the same control shaft as the R_a-rheostat and connected electrically so that R_b decreases linearly from 288kΩ down to 10kΩ, while R_a increases from zero to 10Ω, as the dial is turned from $D=0$ to $D = 10$.

Alternative Solutions when the Tolerance on Z_{in} is Reduced Below 0·1 per cent

First it must be emphasized that the keeping of Z_{in} constant to high precision has no direct bearing upon the precision of the shunt as a measuring-instrument.

Secondly it should be mentioned that the straight-line solution offered in Fig. 4 cannot be applied if the 0·1 per cent tolerance is reduced very much; for instance it seems likely that 0·08 per cent would be the limit; for any closer limit a non-linear rheostat would be called for, and the most likely to be effective would be the so-called 'log' type. With one of these tolerances much smaller than 0·1 per cent could be obtained. To justify this the tolerance on all the other (decade) dials would have to be closed to the same extent by more careful adjustment.

References

1. Wigan, E. R., Mayo, C. G. British Patent 786,189.
2. Wigan, E. R., Mayo, C. G. British Patent 783,967.

A Stroboflash Pulse Sequence Generator for the Study of Turbulent Flow in Fluids

By R. E. King*, M.Sc.

A cold-cathode selector tube is used to obtain a train of pulses of a given sequence over a wide range of pulse repetition frequencies. The output of the sequence generator triggers a stroboflash which provides a photographic indication of the directions and local velocities of turbulence in a fluid.

(Voir page 327 pour le résumé en français: Zusammenfassung in deutscher Sprache Seite 334)

In many problems in hydrodynamics it is desirable to measure simultaneously fluid velocities over an area through which the flow pattern is varying with time. If this flow pattern is repetitive, one way of doing this is to explore different points of the field with a probe at corresponding times. In most cases, however, flow is turbulent and even though the mean pattern may repeat at regular intervals the detailed flow pattern is very irregular. For such cases it is preferable to measure the local velocities simultaneously over the whole field. This is impossible if conventional flow-measuring devices are used, because they all involve introducing a large number of instruments into the fluid.

One method that has been used to explore this kind of problem is to introduce particles having the same density

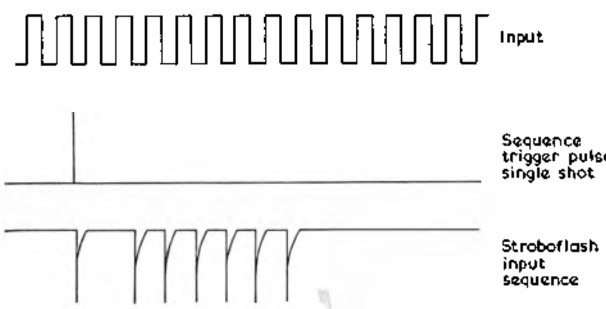

Fig. 1. Stroboflash trigger pulse sequence

as the fluid and to illuminate them by some kind of flashing light. By using shutters in front of a fixed lamp, a sequence of flashes of known duration can be obtained. An 'open shutter' photograph then produces a pattern of dashes which can be analysed to show the field of flow.

This article describes an apparatus to produce a similar effect using a high voltage discharge tube. There are two advantages to this: firstly, the high intensity of the flash makes photography easier, and secondly, the simple triggering required can be easily controlled by an oscillator over a wide range of frequencies.

The only difficulty in analysing a photograph taken by this means is in deciding the direction of flow, which may vary randomly across a given field. In order, therefore, to distinguish between the beginning and end of the flash sequence the first flash is followed by a delay of one trigger-pulse period after which a train of six flashes follow. The trigger sequence is shown in Fig. 1 and examples of the use of the stroboflash unit applied to the study of tidal wave motion are shown in Fig. 2.

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Principle of Operation of the Pulse Sequence Generator

One method of obtaining the desired pulse sequence is by using suitably gated flip-flops in cascade. The large number of such flip-flops required to produce the desired train of pulses is obviated by the use of a selector tube which is very convenient for pulse repetition frequencies up to 5kc/s.

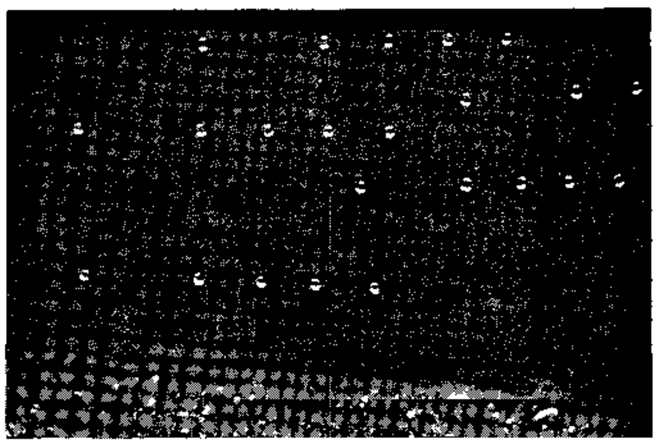


Fig. 2. The path of five globules of a mixture of xylene and carbon tetrachloride in a water stream over a sandy bottom

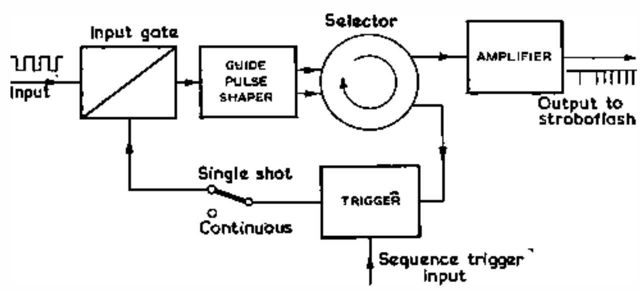


Fig. 3. Arrangement of pulse sequence generator

The pulse sequence generator employing a selector tube is shown in schematic form in Fig. 3.

The guide pulse shaping stage forms pulses suitable for transferring the glow in the selector tube in cyclic order at input signal frequency. By appropriate connexion of the selector tube cathodes the output pulse sequence controlling the stroboflash unit can be varied. Under continuous operation the input gate is at all times open and the output pulse sequence is repetitive. For single shot operation only ten input pulses are fed to the selector tube, allowing one revolution of glow transfer. On reaching the final cathode of the selector tube, the glow causes the input gate to be closed, thereby isolating the input signal from the guide pulse shaper. Initiation of the pulse sequence is brought about by re-opening the input gate.

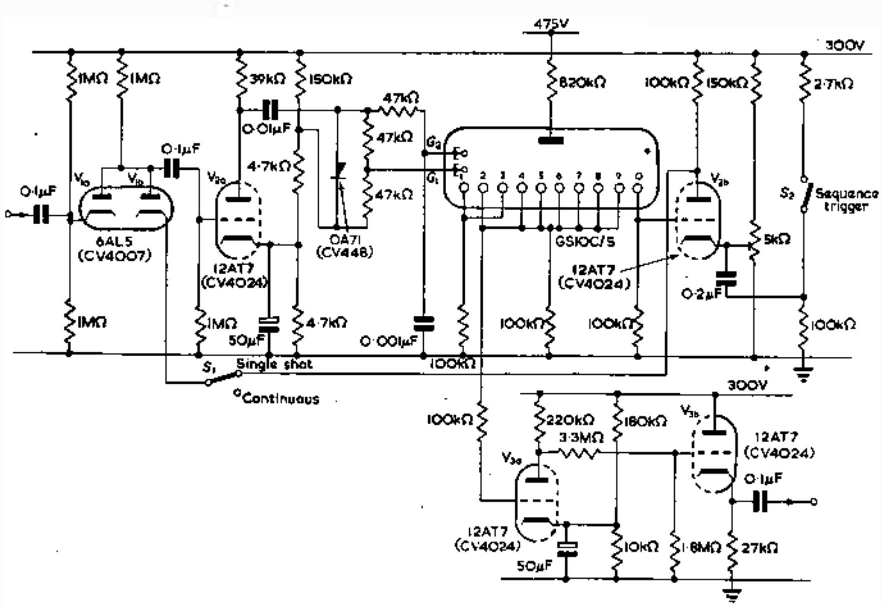
Description of Equipment

The complete circuit of the pulse sequence generator is shown in Fig. 4. Normally the gate control valve V_{2b} is cut off and therefore diode V_{1b} is also cut off, allowing the input signal to proceed to the guide pulse shaping valve V_{2a}. With the switch S_1 in the 'continuous' position, V_{1b} is inoperative at all times and the input signal causes the glow to transfer cyclically in the selector tube which is of the type GS10C/S which has all its cathodes brought out independently. When the glow is on cathodes 2, 4, 5, 6, 7, 8 or 9, positive-going pulses are obtained at the grid of V_{3a}, thereby turning it on and producing negative-going pulses at the output of V_{3b}. This negative-going pulse triggers the Neostron NSP2 which in turn, on discharging the 2μF capacitor, induces a very high impulse voltage in the secondary of the transformer. Due to leakage currents through the surface of the discharge tube, ionization of the gas occurs producing momentarily an intense glow discharge. The circuit of the stroboflash unit is shown in Fig. 5.

When the glow of the selector tube reaches cathode 0, the positive-going pulse at the grid of V_{2b} causes the valve to conduct and its anode drops to a low potential. When the glow is at cathodes other than 0, V_{2b} is maintained cut off by virtue of its large positive cathode potential. With switch S_1 in the 'continuous' position, this drop in anode potential of V_{2b} is of no consequence, and the glow is transferred to cathode 1 and the cyclic operation is maintained. The stroboflash thus operates continuously in the predetermined cyclic order.

When 'single shot' operation is required to obtain a single exposure, the selector tube glow will reach a steady state at cathode 0, for in this position V_{2b} is turned on, its anode is therefore at a low potential and V_{1b} is caused to conduct, thereby cutting V_{1a} off and isolating the input.

Fig. 4. Pulse sequence generator



When the 'sequence trigger' switch, S_2, is closed the cathode of V_{2b} is taken to a high potential, thereby cutting it off. The anode of V_{2b} thus returns to the h.t. potential and V_{1b} is cut off: V_{1a} conducts and the input signal is then passed to the guide pulse shaper V_{2a}. The glow is thus transferred cyclically to the other cathodes of the selector tube until cathode 0 is again arrived at, when V_{2b} is turned on and the input signal is isolated once again.

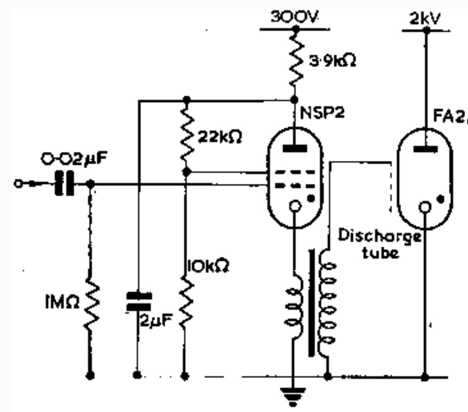


Fig. 5. The stroboflash unit

Only a single sequence of pulses is therefore fed to the stroboflash unit and the pulse repetition frequency is that of the input square wave.

Correct operation is obtained for input peak-to-peak amplitudes of from 20V to 100V. The upper frequency limit of the sequence generator is that of the selector tube which is approximately 5kc/s but for the complete stroboflash unit is dependent on the firing rate of the Neostron NSP2, which is 250 flashes per second.

When the sequence trigger switch is closed, thereby allowing the input to pass to the shaper V_{2a}, a slight delay might be encountered until the positive edge of the input turns V_{2a} on to form the guide pulses. As a result, the transfer of glow from cathode 0, the steady state initial position, to cathode 1 might be erratic, but once this glow has been initiated the pulse sequence follows correctly; for this reason it is preferable to commence the stroboflash pulse sequence with the transfer to cathode 2.

Conclusion

The pulse sequence generator is simple in construction, employing only three valves and a cold-cathode selector tube. Furthermore, it is extremely versatile in that any sequence of eight pulses can be readily obtained by suitable connexion of the selector tube cathodes.

Acknowledgment

The author wishes to thank Dr. J. L. Douce of the Electrical Engineering Department, and Dr. D. McDowell and Mr. R. McBride of the Hydraulics Laboratory of the Civil Engineering Department of Queen's University, Belfast, who have contributed to the design and application of the apparatus.

AN IMPROVED RECORDING OXIMETER

By A. W. Melville*, A.M.I.E.E., D. H. Smith*, M.Sc., and J. B. Cornwall*

An analysis of the theory of ear oximetry is given and an instrument is described which continuously solves the theoretical equation and provides a record of absolute arterial oxygen saturation. The methods previously used for deriving a signal proportional to the logarithm of the intensity of light incident on the earpiece photocells have been found unreliable and have been replaced. The new circuit introduces other advantages, resulting in a high degree of stability and repeatability. The instrument has been developed particularly for use by the anaesthetic staff during prolonged thoracic surgery but it may be employed wherever continuous recording of oxygen saturation is advantageous.

(Voir page 327 pour le résumé en français: Zusammenfassung in deutscher Sprache Seite 334)

THE earliest recorded work on the measurement of the oxygen saturation of circulating arterial blood by photo-electric methods is that of Matthes[1] and Kramer[2]. Many modifications and improvements have been reported since, among them being those of Millikan[3], Goldie[4], Wood[5] and Paul[6]. All recent work has been based on the unequal absorption by oxygenated and reduced haemoglobin of light of two different wavelengths. Light from a small filament bulb passes through the pinna of the ear to two photo-electric cells, fitted with red and infra-red filters respectively.

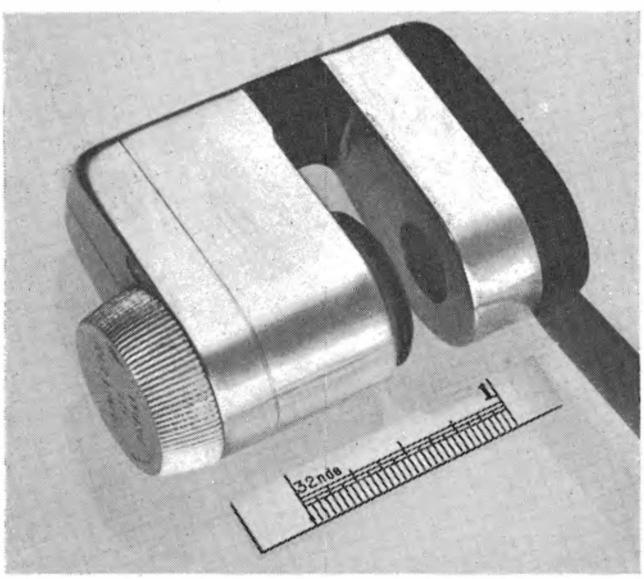


Fig. 1. Earpiece, showing clamping adjustment, O ring holding latex in position and photo-cell aperture

The filters are chosen so that the infra-red radiation (8 000Å) is absorbed approximately equally by oxygenated and reduced haemoglobin whereas the absorption of red light (6 400Å) is much greater for reduced than for oxygenated haemoglobin.

Normally the tissues of the ear contain a mixture of arterial, capillary, and venous blood, but by the gentle application of heat the blood flow can be increased many times so that very little combined oxygen is lost and the blood remains essentially arterial in composition.

The mathematical treatment given in the next section shows that if preliminary adjustments are made with the ear occluded, a continuous record of oxygen saturation can be made on the heat-flushed ear. Using this theoretical work, an oximeter free from errors caused by variations in thickness and pigmentation of tissue has been designed. It is independent of changes in the total amount of blood flowing, and of long-term variations in the illumination incident on the ear.

Instruments based on similar premises have been developed by Goldie[4], Wood[5] and Paul[6]. Goldie used electronic amplifiers, and incorporated a specially designed ratiometer to give a continuous indication of oxygen saturation. Wood used two galvanometers with logarithmic scales to indicate the outputs from the two photocells, and calculated the saturation from a nomograph. Paul used electronic circuits and presented the oxygen saturation directly on a recording milliammeter. In the oximeter to be described here, improved circuits have resulted in the reduction of instrumental errors, and the use of a servo-mechanism has provided a simple method for the continuous recording of the oxygen saturation. Calibration is decidedly simpler than in Paul's design.

Theory

This analysis is based on the assumptions that the blood in the vascular bed is essentially arterial blood and that the complex mass of capillaries can be treated as a uniform film of blood obeying Beer's Law, enclosed in a 'tissue' container. These assumptions have been justified by experimental results obtained over many years (Goldie[4], Wood[5]).

SYMBOLS USED

e_1	=	molecular extinction coefficient of oxy-haemoglobin at 6 400Å
e_2	=	molecular extinction coefficient of reduced haemoglobin at 6 400Å
e'	=	molecular extinction coefficient of haemoglobin at 8 000Å
c_1, c_2	=	concentrations of oxy and reduced haemoglobin respectively, in the blood
l	=	thickness of the equivalent uniform film of blood
d, d'	=	optical densities of blood film at 6 400Å and 8 000Å respectively, where $d = l(e_1c_1 + e_2c_2)$ and $d' = le'(c_1 + c_2)$
a, a'	=	optical densities of ear tissue at 6 400Å and 8 000Å
I, I'	=	illumination intensities incident on the ear, at 6 400Å and 8 000Å respectively
I_o, I_o'	=	intensities incident on red and infra-red photocells respectively, after transmission through an occluded ear
I_t, I_t'	=	intensities at red and infra-red after transmission through a flushed ear

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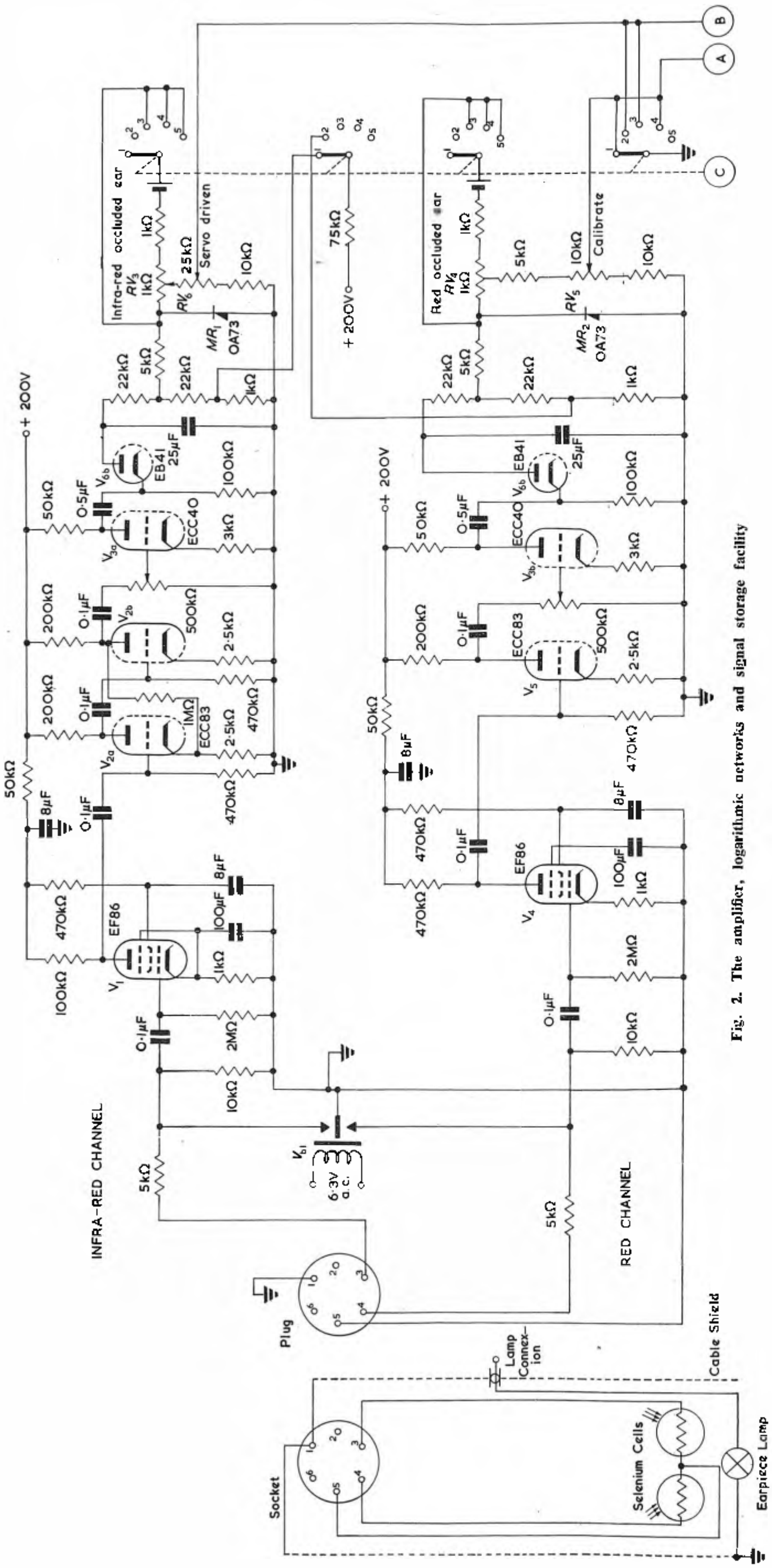

Fig. 2. The amplifier, logarithmic networks and signal storage facility

Applying Beer's Law, $I_0 = I \cdot 10^{-a}$

$$\log I_0 = \log I - a \quad \ldots\ldots\ldots\ldots (1)$$

Similarly

$$\log I_0' = \log I' - a' \quad \ldots\ldots\ldots\ldots (2)$$

$$\log I_t = \log I - (a + d) \quad \ldots\ldots\ldots\ldots (3)$$

$$\log I_t' = \log I' - (a' + d') \quad \ldots\ldots\ldots\ldots (4)$$

In the past photocell circuits which give voltage outputs related to light intensity by an equation of the form $E = k \log I$ have been used. Then equations (1) and (3) give:

$$E_0 - E_t = k(\log I_0 - \log I_t) = kd \quad \ldots\ldots\ldots\ldots (5)$$

and equations (2) and (4) give:

$$E_0' - E_t' = k'd' \quad \ldots\ldots\ldots\ldots (6)$$

$$\frac{E_0 - E_t}{E_0' - E_t'} = (kd/k'd') = (k/k') \cdot \frac{c_1 e_1 + c_2 e_2}{e'(c_1 + c_2)} \quad \ldots\ldots\ldots\ldots (7)$$

$$\frac{c_1}{c_1 + c_2} = (k'R/k') \cdot \frac{e'}{e_1 - e_2} - \frac{e_2}{e_1 - e_2} \quad \ldots\ldots\ldots\ldots (8)$$

where $R = \dfrac{E_0 - E_t}{E_0' - E_t'}$

The percentage oxygen saturation, $S, = 100c_1/(c_1+c_2)$ so equation (8) shows that S is linearly related to R. Thus if R is measured, S can be calculated.

It can be seen that R (and hence S) is independent of I, I', a, a' and l, i.e. R is not affected by changes in the incident light intensity, or by changes in the amount of tissue or blood in the light path. However, it must be remembered that equation (7) was obtained by treating equations (1) and (3), (2) and (4), as pairs of simultaneous equations. This assumes implicitly:

(a) That the incident light intensity remains constant for the time required to make measurements on the occluded and flushed ears.

(b) That flushing the ear does not alter the amount of tissue in the light path.

It is clear from equation (8) that if k, k', e_1, e_2 and e' are known, it is possible to obtain an absolute measurement of oxygen saturation. However it is found to be simpler to calibrate the instrument by comparing R with the value of oxygen saturation determined by laboratory methods, at a number of saturation levels.

Equation (8) must be modified since the photocell outputs have to be amplified electronically. If the amplifiers are linear and have gains of A, A' respectively, equation (8) becomes:

$$\frac{c_1}{c_1 + c_2} = (k'AR/k'A') \cdot \frac{e'}{e_1 - e_2} - \frac{e_2}{e_1 - e_2} \quad \ldots\ldots\ldots\ldots (8a)$$

This shows that for the relationship between oxygen saturation and R to remain constant, the ratio $k'A/k'A'$ must be fixed.

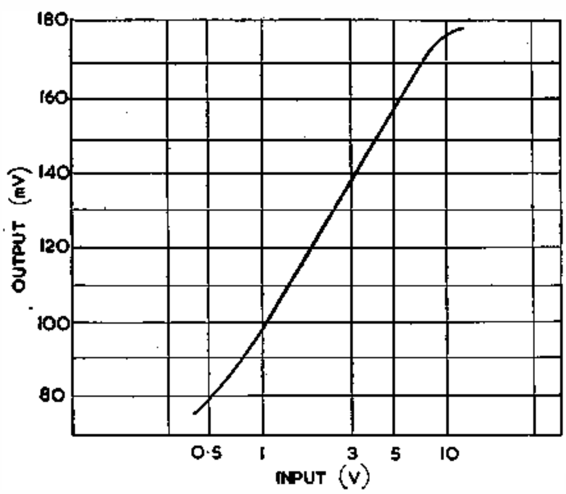

Fig. 3. Characteristic of logarithmic network

If instead of using a logarithmic circuit preceding the amplifier it is arranged that the input to the amplifier is related linearly to the intensity of light incident on the earpiece photocells, and the logarithmic circuit provided subsequent to amplification, then:

$$R = (K_2/K_3)\,\frac{\log AI_o - \log AI_t}{\log A'I_o' - \log A'I_t'}$$

$$R = (K_2/K_3)\,\frac{\log A + \log I_o - \log A - \log I_t}{\log A' + \log I_o' - \log A' - \log I_t'}$$

$$R = (K_2/K_3)\,\frac{\log I_o - \log I_t}{\log I_o' - \log I_t'} \quad \ldots\ldots\ldots\ldots (9)$$

where K_2 and K_3 are the constants of the new logarithmic circuits.

In this case the only requirements of the amplifiers are that they are linear and stable for the period of measurement and initially adjusted so that the output is within the range of the logging circuits.

The above analysis shows that a continuous record of oxygen saturation will be obtained if R can be presented in a suitable form. The value of saturation so obtained will be independent of variations in the amount and colouring of the ear tissue. It is independent of variations in the incident illumination, except such short term variations as occur during the course of an experiment. It is also independent of changes in the amount of blood in the vascular bed.

In the instrument to be described in the following sections the quantity R in equation (9) is continuously recorded.

Description of Apparatus

Earpiece

The earpiece, Fig. 1, is similar in principle and construction to others previously described. The selenium cells are in the form of four quadrants isolated by carefully cutting through the selenium layer on a 15mm diameter cell. Diametrically opposite quadrants are carefully covered with Wratten gelatin filter type 29 for the red band and type 88A for the infra-red band. Connexions are made by soldering with Wood's metal. The light source is an 'Atlas' miniature panel lamp 28V 0·04A chosen for its size and low current, enabling it to be operated from the controlled h.t. supply. Rotation of the knob provides adjustment of the grip on the pinna of the ear and a latex capsule over the light source may be pressurized through a fine polythene tube incorporated with the cable, so that the ear may be rendered bloodless during the preliminary setting up.

Amplifier Channels (Fig. 2)

A single 'Brown' chopper is used as a d.c. to a.c. convertor for the outputs from both pairs of selenium cells. The interrupted signal appears across a 10kΩ resistor and with this load the voltage is strictly proportional to the light intensity over the range used. The 5kΩ series resistor prevents short-circuiting of the cells when the chopper contact is made and eliminates any cell hysteresis which may result from this. The chopper system is isolated from the input grids to eliminate errors resulting from grid current chopping. To eliminate 50c/s pick-up it is important that a single earth connexion is used for the entire input circuit.

The input valves of both the infra-red and red amplifier channels have been chosen for their low noise and hum level. Degenerative triode stages are used to provide negative feedback throughout the remainder of the amplifiers. Since the output from the red filtered cell is approximately 10 times that of the infra-red cell, an extra triode stage is present in the infra-red channel, and some additional feedback is applied across the second and third triode stages. The final single triode stage, which is preceded by a gain control, supplies sufficient voltage to the diode rectifier to ensure strict linearity throughout the operation range.

Logarithmic Circuits (Fig. 2)

With certain germanium diodes it is found that the relationship between diode current and voltage is logarithmic for a limited range of low voltages in the forward direction. The smoothed d.c. voltages available from diodes V_{6a} and V_{6b} are applied to series circuits consisting of a dropping resistor and a germanium diode type OA73. With the constants given it will be seen from Fig. 3 that if the d.c. input is kept within the limits of 1 and 6V, the voltage drop across the diode bears a logarithmic relationship to the input voltage. The output voltage limits are 90 and 170mV. At these voltages the diode characteristics are particularly sensitive to temperature changes and it has been found necessary to provide temperature control. The diodes are located in close fitting holes drilled in a brass block which also houses a bimetal thermostat and a small heating element. The temperature of the block is maintained at 96°F ± 0·2F which provides a voltage stability of ±1 per cent.

Signal Storage (Fig. 2)

Immediately following the logging diodes are the storage circuits in which a voltage equivalent to the occluded ear

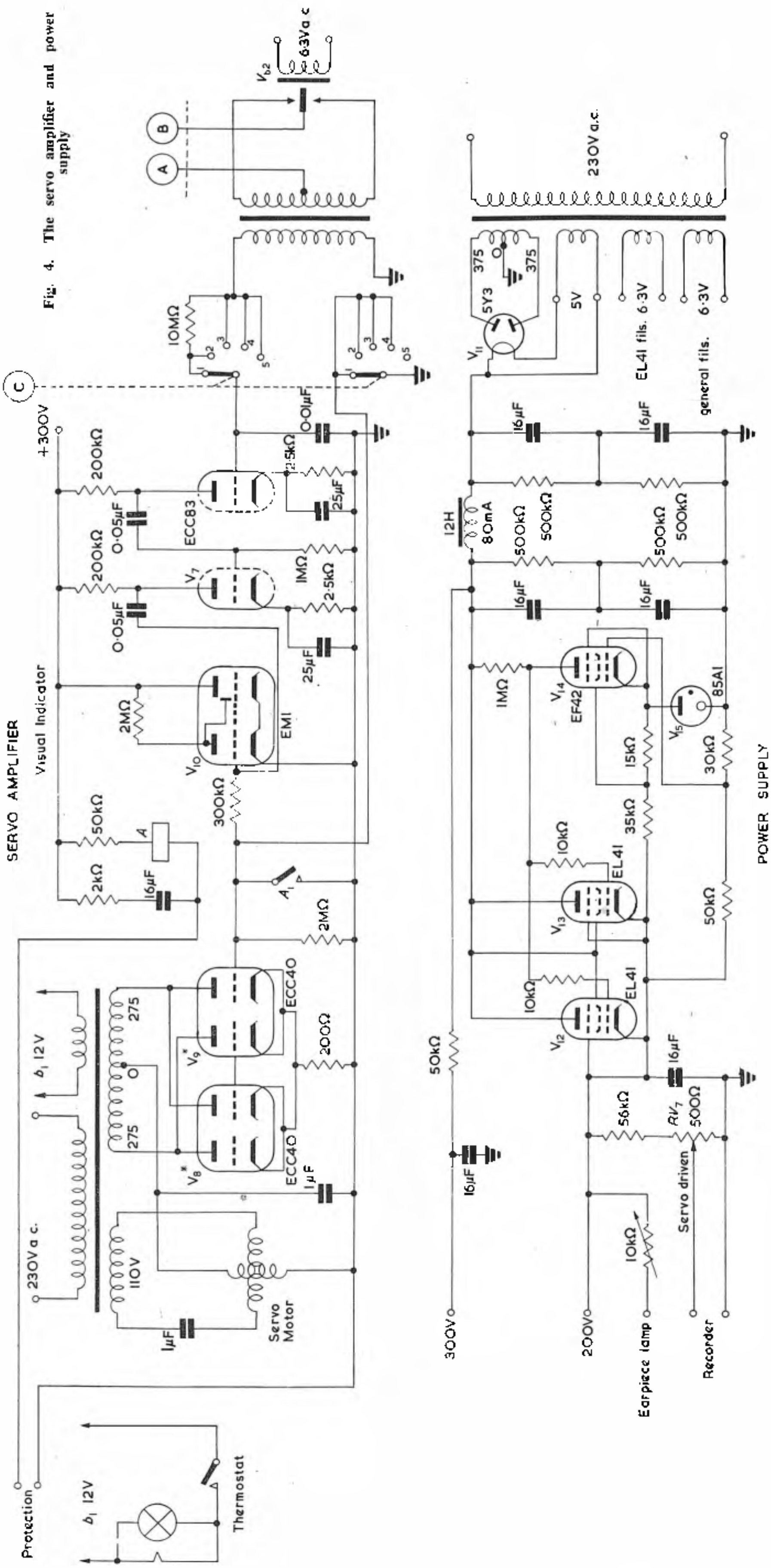

Fig. 4. The servo amplifier and power supply

signals, that is $K_2 \log I_0$ and $K_3 \log I_0'$, may be set up on the potentiometers RV_3 and RV_4. The source of bucking voltage in each case is a 1·5V dry cell. When the ear is allowed to flush, the voltage from each logging circuit decreases and the outputs from the two channels represent $K_2 (\log I_0 - \log I_t)$ and $K_3 (\log I_0' - \log I_t')$ respectively. It remains to record the ratio of these two signals.

The Ratiometer (Fig. 4)

The difference of these voltages is chopped into the input transformer of a servo amplifier patterned on the Brown Honeywell system. The output of this amplifier energizes a servo motor which drives the potentiometer RV_6, Fig. 2, to restore equality in the outputs. The position of the servo motor is related to the ratio of the two signals and is indicated by a 0 to 1mA Esterline Angus recorder energized from RV_7 which is driven by the same shaft as RV_6. If r.f. diathermy interference is encountered in the theatre, the diathermy unit may be equipped with a relay which short-circuits the terminals marked 'Protection' (Fig. 4). This operates an internal relay which earths the grid of the final stage of the servo amplifier. The de-energizing of the internal relay is delayed so that the amplifier channels have fully recovered from any blocking signals before the servo drive is reconnected.

Power Supply (Fig. 4)

A conventional series regulated power unit supplies h.t. requirements of all amplifier stages and the ear-piece lamp.

Standardizing

In order to compensate for long term changes in the constants of the circuit an artificial ear has been constructed so that the overall performance can be checked at any time. In fact, the standardizing procedure is so simple that it may be carried out in a few minutes prior to each recording.

The assembly consists of a small carrier, Fig. 5, one side of which is clear glass and the other side opal ground on one side. The carrier is clamped in the ear piece and its transmission simulates that of the bloodless ear. The sliding filter consists of one thickness of Wratten gelatin filter type 29 held between two surface ground microscope slides. The transmission of the carrier and filter together simulates that of the flushed ear at a particular oxygen saturation, in this case 90 per cent.

RECORDING PROCEDURE

(a) The equipment is warmed up for 20min before adjustments are made and the earpiece should be on the ear for 10min to ensure com arterialization. These periods may be simultaneous.

(b) The gain of each channel is set so that the operating range is within that of the logarithmic circuits.

(c) The ear is occluded by pressurizing the latex capsule

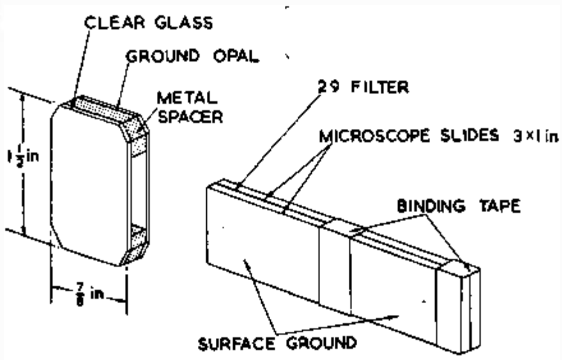

Fig. 5. Details of standardizing filters

above systolic pressure and the amplifier outputs are reduced to zero by potentiometers RV_3 and RV_4.

(d) When the ear is allowed to flush, the oxygen saturation percentage is continuously indicated on the recorder.

When standardizing, this cycle is carried out with the artificial ear in position. Occlusion is simulated by removing the sliding filter and flushing by replacing it. With the filter previously described the standard scale setting is 90 per cent. Operations (b) and (c) are facilitated by a built in 'magic eye' indicator and switching positions which provide for:

Position 1 Setting of infra-red channel gain so that output is within the required limits by comparison with a standard voltage developed across the 1kΩ resistor included in the diode loads of V_6.

Position 2 Ditto for red channel.

Position 3 Storing of occluded ear signal in infra-red channel by adjusting RV_3 for zero output.

Position 4 Ditto for red channel.

Position 5 Recording.

CALIBRATION

The instrument is being calibrated against blood samples withdrawn and examined spectographically. Calibration points obtained to date indicate a scatter of less than ± 3 per cent from a mean curve. The instrument is in routine use on a variety of thoracic surgery cases and has shown excellent long term stability and reliability.

The setting up time is mainly due to the necessity for adequate arterialization of the ear. Attempts have been made to reduce this time by supplying heat at a higher rate from a small infra-red heater but there is apparently a slow process of migration of cellular fluids as the ear expands, which changes the occluded ear setting and it has not been found possible to accelerate this process.

Acknowledgment

The authors gratefully acknowledge the help and encouragement of the Medical Staff at the Greenlane Hospital, Auckland and of the Greenlane laboratory staff for their assistance in the calibration and evaluation of the instrument. They are also indebted to Mr. J. Boyes, technician at the laboratory for his ingenuity in devising and assembling the earpiece.

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A High Speed Pantograph Test Car

Tests have recently been carried out by the Department of the London Midland Region Chief Mechanical and Electrical Engineer, in order to examine in more detail the problem of current collection at high speeds on the 25/6·25kV overhead system proposed for the electrification of British Railways.

A high speed test unit comprising a pantograph test car and a generator van, has been designed and constructed to be hauled either by an electric locomotive or a main line diesel locomotive at speeds up to 100mile/h or by a multiple train unit.

Equipment is installed within the test car to record the following values which can be recorded continuously when the overhead equipment which is being examined is either earthed or alive at 25/6·25kV:

(1) *Pantograph Height:* The static contact wire height of the overhead equipment is measured and recorded at the time of erection. The height recording taken by the test car, however, is the height under dynamic conditions and the uplift of the overhead equipment, which is variable with speed, can be examined at various speeds.

(2) *Pantograph Pressure:* The static pressure of the pantograph on the overhead equipment is 20lb and it is necessary to examine the increase in pressure in relation to train speed produced by the vertical component of the aerodynamic forces and also the forces experienced when running into a graded equipment.

(3) *Acceleration:* The vertical acceleration of the pantograph head is recorded, giving an indication of the contact restoring properties of the pantograph used on the particular trial.

(4) *Loss of Contact:* This recording gives the places where the pantograph leaves the overhead equipment, the length of loss being useful in the study of bouncing.

(5) *Train Speed:* This value is taken from an axle mounted alternator and this source is used to drive the recorder charts via a velodyne set. A wheel diameter compensating device is included in the circuit.

(6) *Track Marks:* It is necessary to give indications of various locations on a test run and this is done by the use of two lamp/photo-electric cell units slung under the vehicle on the centre lines of the two pantographs and aluminium reflectors which are located on the sleepers between rails.

(7) *Time Marks:* Facilities are provided for pre-selected period marks to be given on the charts.

(8) *Event Marks:* Push-buttons are provided within the dome of the test car to give an indication of any passing object such as stations, mile posts, etc.

An instrument console is supported on resilient mountings between the floor and the roof of the test car and contains the necessary recorders, amplifiers, time and event markers and control equipment.

A Physiological Stimulator Using Junction Transistors

By W. T. Catton* and L. Molyneux*

After two years experience in physiology classes with an earlier stimulator using a single point-contact transistor it was decided, in view of the obsolescence of this device, to design a new circuit using junction transistors. The new circuit described involves no alterations in the front-panel control arrangements, provides a better pulse shape and a lower output impedance, with the added advantage of a lower supply battery voltage. The facilities provided are as for the earlier model; output voltage variable from 0 to 25V; frequency variable from 2 to 100/sec; constant pulse-width of about 1msec

(Voir page 327 pour le résumé en français: Zusammenfassung in deutscher Sprache Seite 334)

A TRANSISTOR stimulator previously described (Catton, Molyneux and Schofield[1]) employed a point-contact transistor, a type now obsolescent. A new circuit has been devised using standard junction transistors which provides the same facilities and makes use of the same control panel and switching as the earlier model. In this way the new model may be put into service alongside the old, without the need to alter the operating instructions or procedure in any way. In order to achieve the same facilities as the original model it was found necessary to use three transistors, leading to an increased cost of construction. To set against this are a substantial reduction in the size of battery required (45V for the old model, 27V for the new) and a better performance in that the output impedance is reduced to less than 1kΩ and the output pulse shape is improved. As a result of these improvements the new model has proved more effective, when tested on standard frog preparations, in that only about half the nominal voltage is needed as compared with the old model.

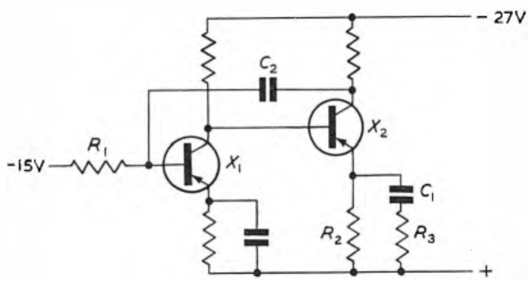
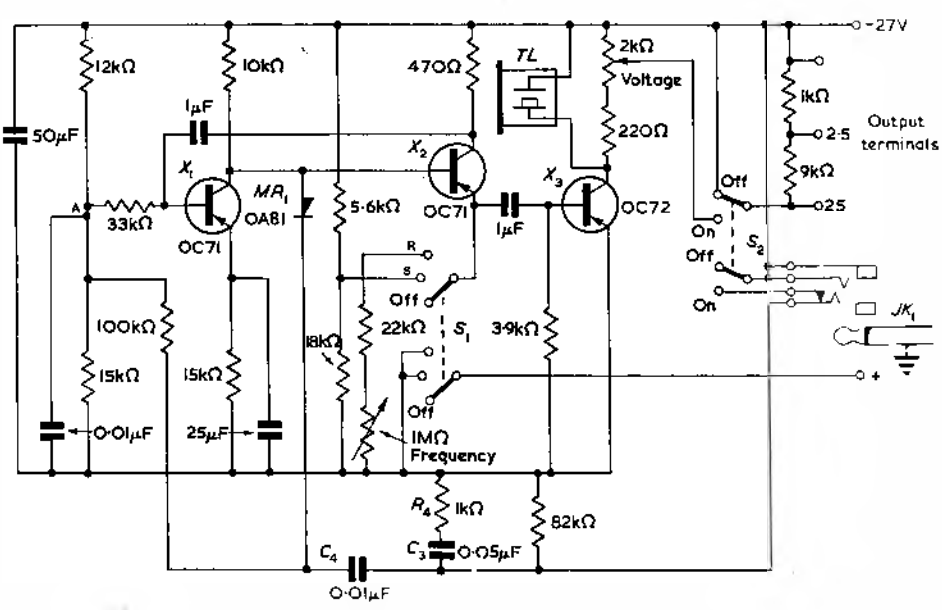


Fig. 1. Diagram of basic circuit for repetitive pulses

The operation of the circuit is as follows. Fig. 1 shows the basic circuit for generation of repetitive pulses and Fig. 2 is the complete circuit. All potentials will be described with reference to the battery positive line. If the potential drop due to the flow of base current in R_1 is neglected, then the potential of the base of X_1 will be set by the potential divider (12kΩ and 15kΩ) shown in the complete circuit, and is about 15V. Due to 'emitter-follower' action the potential of the emitter is also about 15V. The current in the emitter lead is therefore defined by the size of the emitter resistor (15kΩ) and is thus 1mA. Neglecting the base current the collector current is the same as the emitter current, and the collector potential is therefore determined by the supply voltage and the potential drop across the collector load resistor. With the circuit values shown the potential of the collector of X_1 (and base of X_2) is about 17V. Suppose now that C_1 is charged to more than 17V, say 20V. At this potential the base of X_2 is positive with respect to its emitter and the transistor draws negligible current. If the charge is now allowed to leak away through R_2 the potential on C_1 will decrease and there will come a time when the base of X_2 is negative with respect to its emitter. When this happens X_2 will draw collector current and its collector potential will move in the positive direction. This positive-going change is conveyed to the base of X_1 by C_2. As a result X_1 will draw less current, causing its collector potential, along with that of the base of X_2, to move in the negative direction; thus X_2 is driven into heavy conduction and C_1 is recharged through R_3. In the final circuit (Fig. 2) R_3 provides the input resistance of the base of the output transistor X_3, and the charging current of C_1 is used to 'bottom' X_3. The resulting output pulse is flat-topped with fast rise- and fall-times (less than 5μsec). In the final cir-

Fig. 2. Circuit diagram of complete stimulator

S_1 = switch selecting operations; R, repetitive pulses; s, single pulses; off
S_2 = stimulus on/off switch
JK_1 = Jack plug with connexions to external switch (kymograph contacts)
TL = crystal earpiece



* *King's College, University of Durham.*

cuit also a 1MΩ variable resistor forms the major part of R_2, and this together with a value of 1μF for C_1 provides a range of frequency control from about 2/sec to 100/sec. The pulse width is about 1msec with the circuit values shown, and is dependent on the value of C_1 and of the components associated with X_2, particularly its collector load. Thus an alteration of the value of C_1 affects both frequency and pulse width.

For single stimuli, manually controlled, C_1 is disconnected from R_2 and connected instead to a potential divider (5·6kΩ and 18kΩ) which sets it (and the emitter of X_2) to about 19V. A single output pulse is now initiated by momentarily driving the base of X_2 more negative than this. This is achieved by depressing the stimulus key (S_2), when C_3 is charged through R_4 from the negative line and a negative pulse is transferred via C_4 and diode MR_1 to the base of X_2. The facility of external triggering from the kymograph drum contacts by jack plug JK_1, and other details of the control switching are as described for the earlier model.

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A PULSE WIDTH MODULATOR

By D. C. Brown*, Ph.D., and J. E. Baughen†, D.C.Ae.

A pulse width modulator has been produced utilizing the hole storage effect in a junction transistor. Using a signal sampled at 10kc/s, a sensitivity of 20µsec/V was realized.

(Voir page 327 pour le résumé en français: Zusammenfassung in deutscher Sprache Seite 334)

It has been suggested by Price[1] and Deuitch[2] that the hole storage effect encountered in junction transistors might be used to produce a type of pulse width modulation. Fig. 1 shows a typical cathode-ray oscillogram of pulse widening produced by hole storage in the circuit shown in Fig. 2. Deuitch has analysed the form of the output current pulse for the grounded emitter configuration in terms of T_v the hole storage time and T_f the interval between the trailing edge of the input pulse and the time when the trailing edge of the output pulse has fallen to 90 per cent of the final current amplitude.

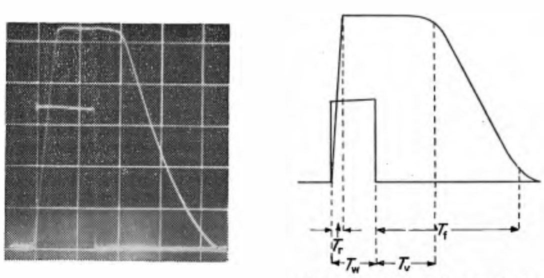

Fig. 1. The input and output pulses showing pulse widening

T_r is the rise-time of the output pulse and T_w the width of the input pulse (7·5µsec in this case). The pulses are displayed in this manner for convenience, in fact they are of opposite polarity

$$T_v = \beta_o T_{co} \ln [\beta_o I_b / I_{cs}] \quad \ldots\ldots\ldots\ldots\ldots\ldots (1)$$

$$T_f = \beta_o \{T_{co} + T_o\} \ln [10\beta_o I_b / I_{cs}] \quad \ldots\ldots (2)$$

In order to produce a good output pulse shape, the rise- and fall-times should be as short as possible. The rise-time of the pulse depends on T_o and T_{ce} and can be reduced by increasing f_o and making the value of the collector resistor (R_1 in Fig. 2) as small as possible. The fall-time is made small by making $T_f - T_v$ as small as possible. However it is also desirable to make T_f large so that the pulse widening effect is great. From equations (1) and (2) it will be seen that to make $T_f - T_v$ small and yet to have T_f large imposes contrary conditions, the former demanding that β_o and T_{ce} be small, the latter that β_o and T_{ce} be large. A compromise solution is to select a transistor with a low value of β_o and as small a value of T_{ce} as possible, such that a satisfactory rise-time is still obtained from the output pulse. A number of transistors were examined and an S.T.C. transistor type TS1 was chosen which gave satisfactory rise- and fall-times when the value of the collector resistor was 1kΩ. (β_o being 15 and T_{co} 3×10^{-7}sec for a typical TS1 transistor).

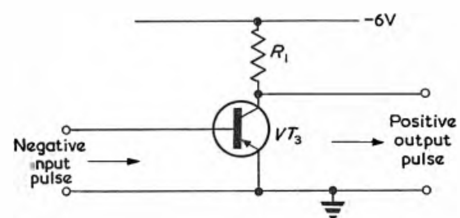

Fig. 2. The hole storage stage

The Modulator Circuit

From equation (2) it will be seen that, as the base current is temperature dependent, then T_f is also temperature dependent, increasing as the temperature increases. If constant output pulse width is to be preserved for a given

Symbols Used		
β_o	=	The static current gain in the grounded emitter configuration
I_b	=	The base current
I_{cs}	=	The collector current in the saturation state
f_o	=	The α cut-off frequency
C_c	=	The collector junction transition capacitance
R_c	=	The collector resistor (in this case R_1)
T_{co}	=	The diffusion time-constant = $1/2\pi f_o$
T_o	=	$R_c C_c$

* *Now with Associated Engineering Ltd.*

† *Now with G.E.C., Ltd.*

steady signal, then some temperature compensation should be introduced. The pulse shape appearing at the collector of VT_3 is not suitable for use in a pulse width modulator system without shaping and it would be convenient if temperature compensation could be produced by the shaping circuit.

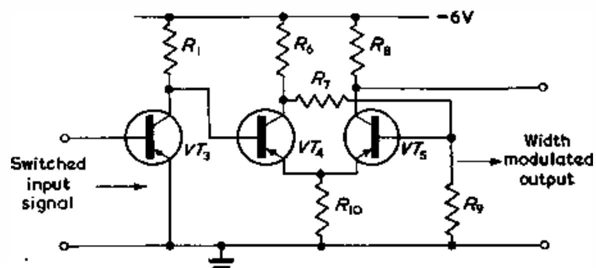

Fig. 3. The hole storage and pulse shaping circuit

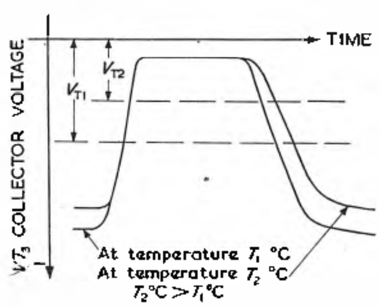

Fig. 4. The effect of temperature on the pulse width from and the triggering level of the shaping circuit

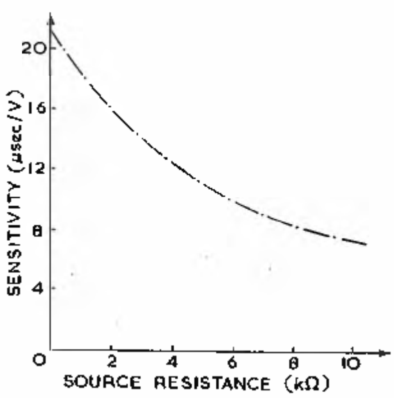

Fig. 5. Variation of the sensitivity of the circuit shown in Fig. 3 with signal source resistance

This is achieved by shaping the pulse with an emitter coupled bistable circuit (Fig. 3). The necessary trigger level of the pulse shaping circuit changes with temperature from V_{T1} to V_{T2} in a manner tending to compensate for the effect of temperature on the pulse width as indicated in Fig. 4. In this case the temperature coefficient of the pulse width was better than ·003/°C from 20°C to 50°C, and as it was slightly over compensated the pulse width decreased as the temperature increased. The rise- and fall-times of the output pulse from the circuit shown in Fig. 3 were better than 0·5μsec.

In the complete modulator system shown in Fig. 6, a 5μsec duration pulse at a repetition rate of 10kc/s was used to sample the signal input by means of a transistor switch. The efficiency of the system (i.e. change in output pulse width produced by change in input signal voltage amplitude) decreases as the signal source impedance increases as shown in the graph in Fig. 5. This is due to the behaviour of the transistor switch[3]. By interposing a grounded collector stage VT_1 between the signal and the switch transistor not only is the effective source resistance reduced, but by d.c. coupling VT_1 to VT_2 a standing negative d.c. level is obtained which means that the system can

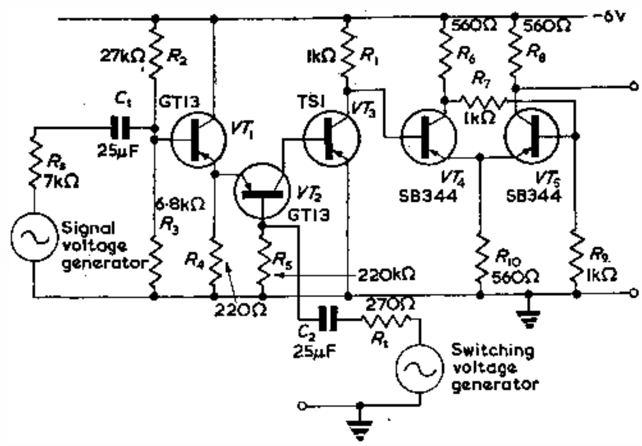

Fig. 6. The complete modulator

R_s being the internal resistance of the signal voltage generator and R_t the internal resistance of the switching voltage generator

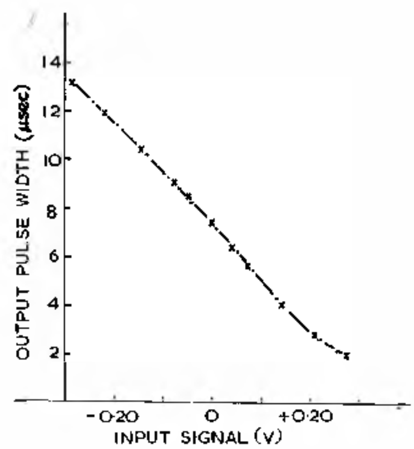

Fig. 7. Output pulse width from VT_5 as the input signal amplitude to VT_1 is varied

The experimental points are shown. In this case the input signal source resistance was 5kΩ

now handle positive as well as negative signals. The relationship between input signal voltage and output pulse width from the collector of VT_5 is shown in Fig. 7. This graph was obtained by applying a 5kc/s square wave of varying amplitude to the system in order to have a truly dynamic, rather than a static, calibration. The accuracy of measurement of the pulse width was only about $\pm$0·1μsec, but it can be seen that a reasonable linearity was obtained over the central operating region.

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A Voltage-Modulated Variable Pulse-Rate Generator

By E. J. C. Fowell*, B.Sc., A.M.I.E.E., and A. Cowley*

This article describes a pulse generator circuit, the output frequency of which is a linear function of the input voltage.

The circuit employs a screen coupled double Phantastron pulse circulating system with suitable modifications to improve the accuracy of transfer at low output frequencies.

Four output frequency ranges are provided between 5 and 2 500p/s.

A complete circuit diagram together with performance results are included.

(Voir page 327 pour le résumé en français: Zusammenfassung in deutscher Sprache Seite 334)

A PULSE generator circuit is described which is suitable for the generation of variable pulse rate signals according to the fluctuations of an input voltage which itself is an analogue of the desired pulse rate at any given time.

Such a device is very helpful in carrying through tests on complete digital servo systems of the incremental pulse counting type or alternatively on the separate components of such a system; for example, the harmonic or step response of a pulse rate meter may be deduced by supplying a sine or step wave input to the pulse generator which then supplies the appropriate input signal to the pulse rate meter under test.

This generator has been designed with this type of analysis in mind and is arranged to work in either of two modes:—

(a) For real input simulation the variation from zero to full pulse rate is obtained by supplying a voltage varying between the limits of −75 and +75V (to match the Servomex LF51 function generator output).

(b) For harmonic analysis, the connexion of the input terminal to earth potential provides a steady pulse output at half the maximum pulse rate. If a sine wave signal is now inserted at the input instead, with its mean level at earth potential, then the mean pulse output at half maximum pulse rate is modulated at the input frequency, the depth of modulation depending on the input signal height, full modulation being obtained with a 75V peak signal.

The circuit itself consists basically of two identical Phantastron units, connected to form a pulse circulating loop (Fig. 1). The pulse frequency is determined directly by the grid return voltages of the triangle forming valves, these voltages being obtained from the input terminal. The pulse frequency is then linearly proportional to the input voltage, the accuracy of transfer declining rapidly at near zero value of output frequency.

A certain amount of circuit sophistication has been introduced to maintain adequate performance of the basic Phantastron at low grid return voltages, i.e. at the low frequency end of the operating scale.

Further circuit complication arises from the need to make the pulse circulation self-starting under all conditions of use, i.e. on switching-on the circuit or after a below-cut-off excursion of the input voltage.

Circuit Description

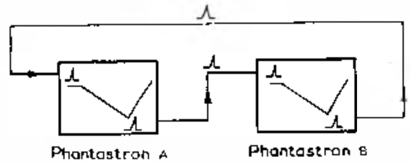
Timing Circuit Description

As stated above, the waveform generation is carried out in turn by two Phantastron generators[1,2,3] each of which produces a linear descending ramp of voltage, whose slope is determined by the instantaneous input signal acting as the control grid return voltages, for the Phantastron pair.

Each ramp wave is started and terminated at specified potential levels, so that the time for the production of the voltage ramp is inversely proportional to the input voltage.

At the termination of the ramp a trigger-voltage then operates the second Phantastron circuit, while the first circuit resets itself slowly to its original state. The second Phantastron then re-triggers the first Phantastron at the end of its ramp waveform and the process once started maintains itself unless one or both of the ramp forming circuits is disabled by some cause.

Fig. 1. Basic connexion of two Phantastrons



Referring to Fig. 7 it will be seen that the Phantastron valves (V_1, V_9) have cathode-follower connexions between the anode and the run-down capacitors (C_a, C_b, respectively). This well-known modification[4] allows the capacitor in question to recharge itself from the cathode of the cathode-follower rather than through the high anode load resistance of the Phantastron valve, thus providing the necessary quick return time required for properly defined circuit action.

The input and output circuits of the Phantastrons are directly coupled in order to make the generator self-starting after any enforced quiescent period. (This matter is described in more detail later in this article.)

The input positive step trigger signal to a Phantastron stage is applied via a diode gate to the suppressor grid of the ramp forming valve (Fig. 2). The diode gate serves to reduce the loading of the suppressor grid circuit on the trigger circuit and vice versa, and also prevents unwanted one sided oscillations in the circuit due to subsidiary closed loops.

The run down at the top end of the capacitor C (Fig. 2) is arrested at some value above that corresponding to the 'bottomed' condition of the ramp forming valve.

This is done by means of a triode voltage comparator which is cathode coupled to the cathode-follower in such a way that the space current flow changes from the cathode-follower triode to the comparator triode at some voltage determined by the potential at the comparator triode grid.

An enlarged portion of the run down near the point of arrest appears at the anode of the comparator triode and

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provides a trapezoidal step output signal from the Phantastron irrespectively of ramp slope.

Current injection via the 'zero set' resistor to the junction of the resistor R and capacitor C (Fig. 2) enables the input signal voltage to be arranged so that an input varying from -75 to $+75$V gives a complete range of pulse outputs from zero to maximum.

Method for the Maintenance of Active Coupling Between the Two Phantastrons

Chance and Williams[3] give two double Phantastron circuits based on the generic form of Fig. 1. One is a double version of the screen coupled circuit[5], the other a double version of the cathode coupled circuit[6] of the Phantastron.

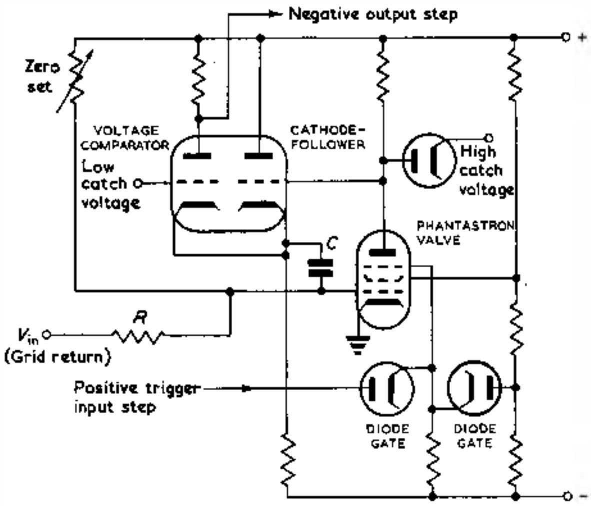


Fig. 2. Single Phantastron stage

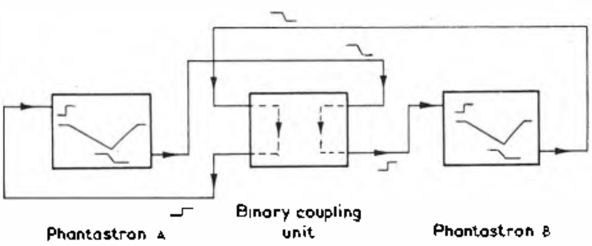


Fig. 3. The method of coupling

It is perhaps worth commenting on some of the properties of these two circuits:—

The screen coupled version gives a better run-down linearity as the grid-to-anode circuit gain is not reduced by cathode degeneration. This circuit would also appear to lend itself to modulation of the grid return voltage as the input grid swing is smaller and has better defined end stops than the cathode coupled version.

It suffers from one major defect; it is unable to recover from the quiescent state should it once attain it owing to the a.c. couplings between the two circuits.

The cathode coupled circuit on the other hand has an extremely elegant d.c. coupling system which will not allow the quiescent state to be maintained. However, the circuit suffers in the comparison as to waveform linearity and the large step voltages which occur at the control grid terminal.

Taking these considerations into account it was decided to go ahead with the clumsier screen-coupled version and complicate the inter-coupling arrangements still further to prevent the maintenance of the quiescent condition.

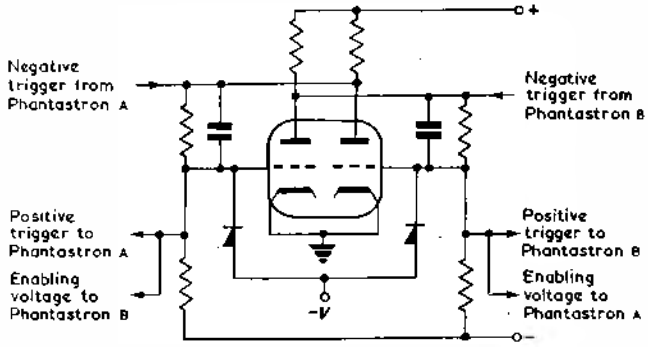


Fig. 4. The binary circuit

The method adopted involves the use of a d.c. coupled binary circuit which ensures that either one or other of the Phantastrons has its suppressor grid lifted to earth potential. Under these conditions the circuit will only become quiescent if the input voltage is held low enough to cut off one or both of the Phantastron valves. The method of coupling is indicated in Fig. 3 which shows the signal flow path.

The binary circuit is entirely conventional but is shown in Fig. 4 for the sake of completeness.

Modifications Incorporated to Sustain the Operation of the Phantastron Circuits at Low Grid Return Voltages

In the operating condition, the Phantastron valve is only passing a small space current and the control grid voltage V_g will be somewhere near the cut-off value.

At low repetition rates (see Fig. 5) the charging current i into capacitor C will be small and the grid return voltage, normally held at some large positive voltage, will have to tend towards the value of V_g which may be typically 5 to 10V negative with respect to earth potential.

Under these conditions, variations in V_g over the waveform time produce significant variations in the value of $(V_{grid\ return} - V_g)$ and therefore in the current i so that the generated waveform becomes sensibly non-linear.

If continuity of output at the low frequency end of the scale is more important than linearity of output, then these departures may not be troublesome.

Of more moment is the fact that, having taken the grid return voltage to a small negative value, there is now no lifting mechanism to swing the control grid into

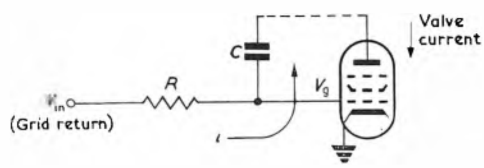
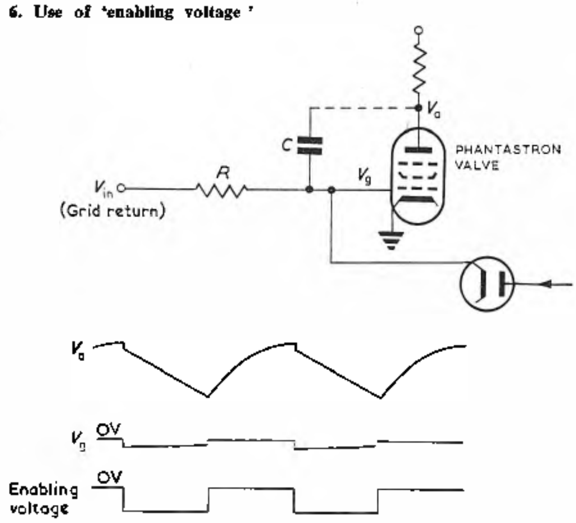


Fig. 5. Phantastron at low repetition

grid current at the end of the ramp waveform so that a large enough space current may not be produced in the valve to give screen-suppressor re-triggering. Also the capacitor C may not now be re-charged via the grid-cathode diode. Altogether, the effect is to cause a failure of the waveform generating action.

To overcome this collapse of action, an 'enabling voltage' waveform is connected to the control grid of the Phantastron via a disconnecting diode. This waveform is negative-going during the sawtooth generation period, but during the reset time the waveform rises to approxi-

Fig. 6. Use of 'enabling voltage'



mately earth potential and lifts the control grid of the Phantastron valve up with it (if necessary) through the diode (Fig. 6). The two waveforms required are obtained from the binary circuit as shown in Fig. 4.

By this method, the minimum stable idling frequency was reduced from 1/10th to 1/25th of the maximum operating frequency in any one range.

Performance of the Complete Circuit

The complete pulse generator is shown in Fig. 7.

The pulse frequency output is provided in ranges to give outputs of 5 to 100p/s, 20 to 500p/s, 50 to 1 250p/s, and 100 to 2 500p/s by switching various values of charging capacitors into the two Phantastron circuits.

Fig. 8 shows a typical calibration curve for the lower pulse rate range (i.e. 5 to 100p/s). The maximum departure from linearity on any one range does not exceed 5 per cent of the actual pulse rate.

Setting Up Procedure

With the input voltage set to -75V the idling control potentiometer R_1 of Fig. 7 should be adjusted until the output pulse rate corresponding to the lower end of the

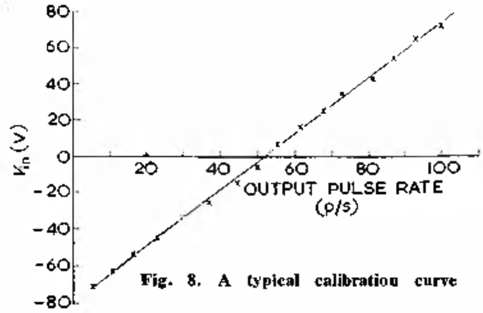


Fig. 8. A typical calibration curve

particular range is obtained. The l.f. balance potentiometer R_3 should now be adjusted for a 1:1 mark-to-space ratio of the output waveform. The maximum frequency control R_2 may now be set for the desired maximum pulse rate with an input of $+75$V.

This procedure should be repeated until the conditions stated are satisfied.

Acknowledgment

The work described has been carried out on a Machine Tool Development Project under the sponsorship of the National Research Development Corporation at the Manchester College of Science and Technology. The work is continuing under a grant of the Department of Scientific and Industrial Research.

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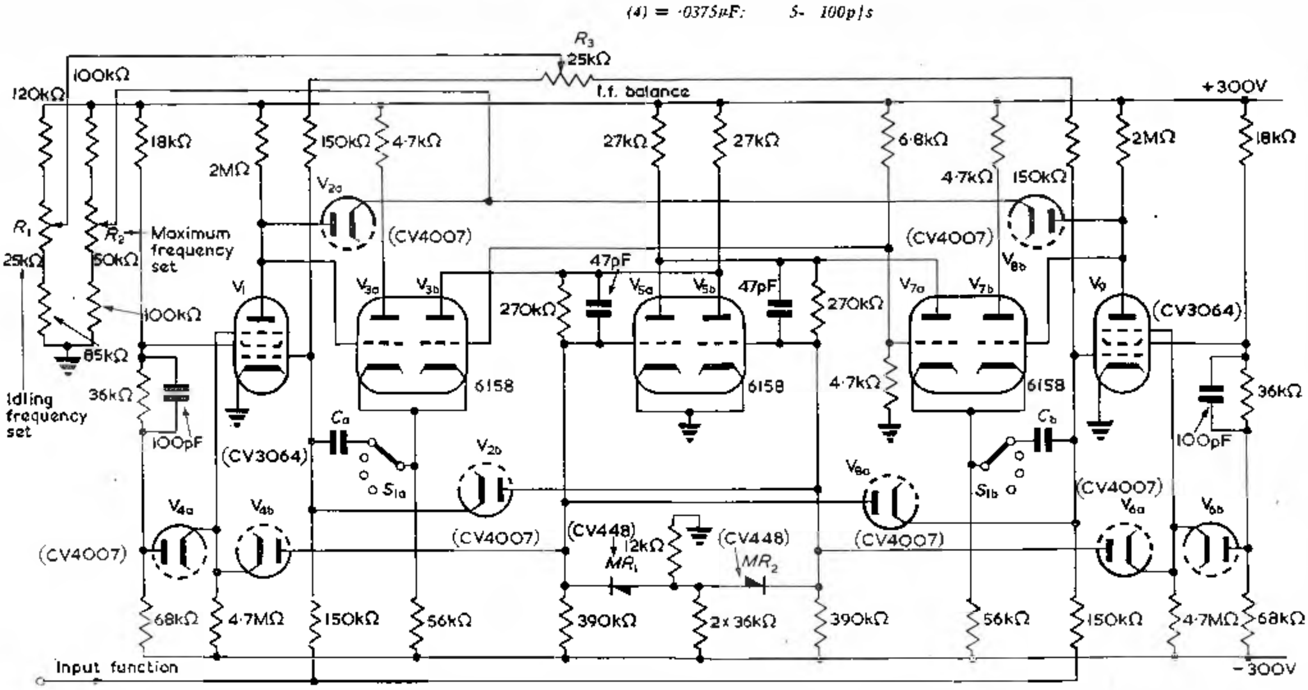
5. Chance, B., Hughes, V. W. Ibid p. 199.
6. Chance, B., Hughes, V. W. Ibid p. 204.

Fig. 7. The complete pulse generator

C_a and C_b :—Position (1) = 1 500pF: 100-2 500p/s
(2) = 3 000pF: 50-1 250p/s
(3) = 7 500pF: 20- 500p/s
(4) = ·0375µF: 5- 100p/s



Short News Items

The Royal Society announces that the following are among the recently elected Fellows of the Royal Society:

Professor R. H. Brown. Professor of Radio Astronomy in the University of Manchester, Jodrell Bank Experimental Station for his many contributions to radio astronomy on galactic and extra galactic radio emissions.

Dr. Louis Essen. Deputy Chief Scientific Officer at the National Physical Laboratory, Teddington, Middlesex for his work on the precise measurement of frequency and of the velocity of light.

Dr. D. K. MacDonald. Principal Research Officer, Division of Pure Physics, National Research Council of Canada, Ottawa for his investigations on the thermal and electrical properties of metals with particular reference to the study of electron interactions.

The Radiotelevisione Italiana has placed an order with E.M.I. Electronics Ltd for four television camera channels Type 203 for use in covering events at this year's Olympic Games taking place in Rome in August. Among other recent Italian orders for electronic equipment placed with E.M.I. is one for an analogue computer variable time delay unit which is being used by Centro di Studi Nucleari di Ispra.

The number of combined television and sound licences during February, throughout Great Britain and Northern Ireland, increased by 148 456, bringing the total to 10 368 323. Sound only licenses total 4 597 891, including 422 780 for sets fitted in cars.

Electronic computer users and potential users of data transmission will shortly receive a questionnaire from the British Post Office asking what data transmission facilities may be needed in the future. The survey will be made of about 350 organizations including banks, insurance companies, finance houses, and industrial concerns and Government Departments. As more and more computer and punched card equipment is brought into use for commercial, industrial and scientific purposes the necessity for new and sometimes more speedy transmission of data arises. This usually means sending information from branches or departments of an organization to a central point for processing, by computer or punched card equipment, and sometimes the return of information to the originating point after processing. The Post Office has a varying range of apparatus and lines which can be used for this purpose and is anxious to keep abreast of events and to develop new facilities.

The first British closed-circuit television system to be installed in Wall Street is now enabling executives of a large New York stockbrokers' office to study the latest ticker-tape stock market movements without leaving their offices. The television system relays a continuous picture of moving ticker-tapes giving price changes on the New York and American Stock Exchanges, to seven large-screen monitors in offices throughout the building. The order for this equipment, which was designed and manufactured by E.M.I. Electronics Ltd, follows the recently announced trading agreement between the British company and the Fairbanks Whitney Corporation of New York. The equipment was delivered and working within three weeks of the agreement's announcement. Without closed-circuit television, fourteen ticker-tape machines and seven optical projectors would have had to be installed in order to provide the same on-the-spot service. The television system is much less expensive and eliminates the need for noisy tape machines throughout the building.

The Indian Government and Marconi's Wireless Telegraph Co. Ltd, have signed an agreement in New Delhi for co-operation in the local manufacture under licence of equipment of Marconi design. The agreement, which provides also for technical assistance and the supply of materials and components, will form the basis for Indian manufacture of equipment in the aeronautical radio, sound and television broadcasting, communications and radar fields. The range of aeronautical radio equipment has particular application to the AVRO 748 transport aircraft which is being produced at Kanpur.

British Electronic Engineers on Thursday, 24 March, showed technical delegates from the six Common Market countries equipment and methods developed here to control rapidly-increasing upper airspace traffic in Western Europe. Forty delegates of the Eurocontrol Technical Working Group were invited by the Ministry of Aviation to discuss a series of technical papers, presented by members of the Electronic Engineering Association, before visiting London Airport to be shown British electronic equipment already available. An experimental electronic air traffic control system, which is being developed for use at the Oceanic Air Traffic Control Centre at Prestwick for the Ministry of Aviation, was described. Captain V. A. M. Hunt, Director, Control and Navigation, Development and Planning, at the Ministry of Aviation, presided at the presentation of the papers.

A Comet 4 Simulator, designed and manufactured by General Precision Systems Limited, Aylesbury for Aerolineas Argentinas, has now been completed for shipment to South America. Coupled to the simulator is an AT.500 Mark 2 Automatic Radio Aids Unit. This unit has a chart recorder in which a map of the route or airport area is placed and on which the track flown by the aircraft during the exercise is traced automatically. Controls on the unit enable the instructor to place up to eight separate ground radio stations anywhere in the map area to correspond with those on the actual route or airport area. Each station is given its appropriate identification call sign and frequency. During the exercise, the crew use the radio and navigational equipment in exactly the same way as they would in the air and the resultant signals are automatically fed to the instruments, indicator lights and to the earphones.

An international colloquium will be held at Grenoble, France from 27 September to 1 October, 1960 on "The Physics of Electrostatic Forces and their Applications." Full particulars can be obtained on application to Professor N. J. Felici, Institut Fourier, Place Doyen-Gosse, Grenoble, Isère.

The Pye Instrument Landing System has been selected by the East African Civil Aviation Authority for installation at the newly developed Nairobi airfield. The use of this airfield, with its 13 500ft runway for the new jet main route through Africa, makes precision radio guidance essential.

An experimental installation of closed circuit television has been completed by Marconi's Wireless Telegraph Co. Ltd at London Airport to enable the authorities to determine whether the control of aircraft parking can be improved. To this end, two BD871 type television cameras have been installed at the top of lighting stands, one at either end of a main parking area—an area chosen because it is out of the Marshalling Supervisor's direct line of vision. The signals from the cameras are fed, via control units installed at the bases of the lighting stands, to television monitor screens at the Marshalling Supervisor's position, to give him a view of the parking area from either end. At present a Marshalling Supervisor, situated in the main passenger arrival and departure building, relies on radio-telephony for reports of aircraft parking movements because his view of the parking area is restricted to that portion which lies in front of the building.

The Audio Manufacturers' Group of the British Radio Equipment Manufacturers' Association has now published its specification for the methods of measuring and expressing the performance of audio frequency amplifiers. Performance standards are also given in the specification, one of the aims of which is to provide parameters for high fidelity equipment. The specification, in its provisional form, is being made available to all those who are thought to be in a position to make comments on the suitability of the methods and requirements; these comments, which should be submitted not later than 31 July, 1960, will be taken into account in the preparation of the final specification, which will be issued later this year. Requests for copies should be addressed to The Secretary, Audio Manufacturers' Group, 49 Russell Square, London W.C.1.

Dr. H. G. Taylor, Director of the Electrical Research Association, and Mr. L. Gosland, Research Manager, have left for a six weeks' visit to the United States and Canada. Among the places to be visited in the United States will be the Massachusetts Institute of Technology, The Batelle Research Institute, The Armour Research Foundation and the Underwriters' Laboratory at Chicago. In Canada they will visit members of the Electrical Research Association which include the National Research Council and the Defence Research Board. Dr. Taylor will be returning early in May, while Mr. Gosland will be going to the West Coast to visit the British Columbia University and Research Institute.

Mr. L. T. Hinton (Standard Telephones and Cables Ltd) and Mr. R. R. C. Rankin (Mullard Ltd) have been re-elected Chairman and Vice-Chairman respectively of the Electronic Engineering Association following the annual general meeting and the election of its new Council on 17 March.

A Symposium on the reliability of Service equipment was held last month at the Headquarters of the Federation of British Industries. The 300 present included senior representatives of the Admiralty, War Office, Air Ministry, Ministry of Aviation and industries contributing to the supply of military equipment. Papers were presented outlining the various aspects of the problems relating to the high reliability required for the modern complex weapons of war. The symposium arose out of the full recognition that these problems require the closest co-operation among Government Departments, research establishments, Service organizations and many sections of industry. A number of the papers will be circulated among professional societies, technical journals, etc., for the purpose of spreading knowledge to commercial interests.

The Central National d'Etudes des Telecommunications (C.N.E.T.) at Issy-les-Moulineaux, Paris, is the central telecommunications research laboratory of the French P.T.T. With a total staff of some 2 000 it is, it is claimed, the largest research centre in Europe engaged wholly in electronics. Among its recent achievements has been the development of a unipolar field-effect germanium transistor known as the Tecnetron and capable of operation up to 1 000Mc/s. C.N.E.T. has also developed an electronic letter sorting machine and a prototype electronic telephone exchange.

The English Electric Valve Company Limited, England, and Eitel McCullough Incorporated, U.S.A. have concluded an agreement for the exchange of technical and manufacturing information over a large range of klystrons, travelling wave tubes and power tubes. This agreement will enlarge the range of tubes in their respective catalogues and enable both Companies to get into production quickly on a composite range of each others products.

The British Council are sponsoring Sir Willis Jackson, President of the Institution of Electrical Engineers, and Mr. W. K. Brasher, Secretary of the Institution, to hold discussions in Poland on the training of electrical engineers and they have now left for Warsaw. They have been invited by the Polish Minister of Higher Education, Mr. Henryk Golanski, a recent British Council visitor to this country. Mr. Golanski studied in the Electricity Department of Warsaw Polytechnic and is a former Vice-Minister of Industry and Trade, and Vice-Minister of Light Industry.

V.H.F. radio communication equipment valued at £180 000 has been ordered from Plessey International Limited by the South African High Commissioner in London. The equipment which will be used by the armoured fighting forces of the Union was developed by Plessey in close co-operation with the S.R.D.E. (Signals Research and Development Establishment).

An extensive system for relaying national broadcast programmes is being installed in Korea as part of a Government scheme to promote widespread distribution of news and general information. Radio amplifier systems, supplied by Philips, Holland, are being installed at 403 points throughout the country and each of these will feed 100 loudspeakers. In this manner, the Korean Government is hoping to cover many of the towns and larger villages in South Korea where privately owned radios are still rare. With each of the 403 radio amplifier units, Philips are supplying microphones to enable local 'broadcasting' through the loudspeaker systems of items of purely local interest.

MEETINGS THIS MONTH

THE BRITISH COMPUTER SOCIETY LIMITED

Birmingham Branch

Date: 11 May. Time: 6.30 p.m.
Held at: Large Theatre, Electrical Engineering Department, The University, Edgbaston
Annual General Meetings of the Branch and of the Midland Region

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, unless otherwise stated

Computer Group

Date: 4 May. Time: 6.30 p.m.
Lecture: Computer Controlled Television Displays for Flight Simulators.
By: J. N. Naish

Radar Group

Date: 11 May. Time: 6.30 p.m.
Lecture: Radio Guidance in the Automatic Landing of Aircraft
By: J. S. Shayler.

Students' Meeting

Date: 20 May. Time: 7 p.m.
Lecture: Servomechanisms and Electronics Engineers—the Present and the Future.
By: M. James
(Note: Attendance at this meeting is not limited to Students.)
Date: 25 May Time: 3 p.m.
Held at: The Gustav Tuck Theatre, University College, London.
Symposium on: Stable Frequency Generation

Liverpool Section

Date: 4 May. Time: 7 p.m.
Held at: University Club, 2 Mount Pleasant, Liverpool
Annual General Meeting.

THE INSTITUTION OF ELECTRICAL ENGINEERS

Electronics and Communications Section

Date: 2 May. Time: 5.30 p.m.
Held at: Savoy Place, London W.C.2.
Lecture: Planning and Installation of the Sound Broadcasting Headquarters for the B.B.C.'s Overseas and European Services.
By: F. Axon and O. H. Barron.

East Anglian Sub-Centre

Date: 2 May. Time: 6.30 p.m.
Held at: Cavendish Laboratory, Cambridge.
Annual General Meeting.
Lecture: Fundamental Particles.
By: Professor O. R. Frisch.

Northern Ireland Centre

Date: 3 May Time: 6.30 p.m.
Held at: David Keir Building, Queen's University, Stranmillis Road, Belfast.
Annual General Meeting.

North Lancashire Sub-Centre

Date: 4 May. Time: 7.15 p.m.
Held at: N.W.E.B. Demonstration Theatre, Friargate, Preston.
Annual General Meeting.
Lecture: Storage and Manipulation of Information in the Brain
By: Dr. R. L. Beurle.

North Staffordshire Sub-Centre

Date: 16 May. Time: 7 p.m.
Held at: Technical College, Stafford.
Annual General Meeting.
Lecture: The Characteristics and Protection of Semiconductor Rectifiers
By: D. B. Corbyn and N. L. Potter.

South Midland Centre

Date: 2 May. Time: 6.30 p.m.
Held at: James Watt Memorial Institute, Birmingham.
Lecture: Radio Aspects of the International Geophysical Year.
By: R. L. Smith-Rose.

South-Western Sub-Centre

Date: 5 May. Time: 3 p.m.
Held at: Electric Hall, Torquay.
Lecture: The Application of Transistors to Line Communication Equipment.
By: H. T. Prior, D. J. R. Chapman and A. A. M. Whitehead.

WOMEN'S ENGINEERING SOCIETY

Date: 18 May. Time: 7 p.m.
Held at: 'Hope House', 45 Great Peter Street, Westminster, London S.W.1.
Lecture: Electronic Weighing.
By: Miss R. Winslade

LETTERS TO THE EDITOR

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

Rectangular Hysteresis-Loop Magnetic Cores as Switching Elements

DEAR SIR,—With reference to Mr. Kaposi's article in the May 1959 issue, I should like to make the following remarks:

(1) The expression obtained by the author and determining the switching time of a core, when a constant voltage is applied is:

$$\tau = \frac{(I_B - I_A)/V_0}{(N^2A/l)\mu_2} \text{ .... (8)*}$$

where N = number of turns on the coil

l = mean magnetic path length

A = cross-sectional area of the core

V_0 = constant voltage applied.

For the significance of other symbols see Fig. 1.

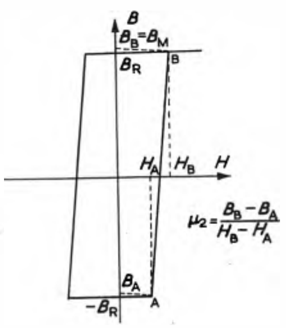


Fig. 1. An idealized dynamic magnetization curve for the rectangular hysteresis-loop magnetic material.

The differential permeability μ_2 is assumed being constant

While equation (8)* is not incorrect, it is well known that the currents I_A, I_B are the dynamic constants and the differential permeability μ_2 is generally a function of magnetic field applied to an arbitrary core. Therefore the calculation of the switching time by the use of equation (8)* is an extremely difficult procedure. Fortunately we can obtain a simple, practically useful equation without the complicated derivation, based on the integral form (the principle of volt-seconds absorbed by a core) of the well known Faraday's law. Then neglecting the series resistance in circuit as well as leakage inductance we have simply:

$$\tau = \frac{NA(B_R + B_M) . 10^{-8}}{V_0} \text{ sec ...... (1)}$$

Where the significance of symbols used is the same as in equation (8)* and is evident from Fig. 1. B_R, B_M is in Gauss, V_0 in volts, A in square centimetres. For magnetic cores of high rectangularity $B_M \simeq B_R \simeq B_s$ where B_s = saturation flux density, and:

$$\tau = \frac{2NAB_s . 10^{-8}}{V_0} \text{ sec ...... (1a)}$$

Equation (1) and especially (1a) shows that the switching time τ does not depend upon the differential permeability of the core material. This explains the good temperature stability of constant voltage switching, because the saturation flux density depends only slightly upon the temperature in the range well below the Curie point (especially for metallic tape cores). On the other hand the switching coefficient S_w as well as threshold field H_0 (the field beyond which the irreversible changes of magnetization take place) decreases strongly with increasing temperature; thus the switching time τ under constant current conditions depends strongly upon the ambient temperature:

$$\tau = \frac{S_w}{H_M - H_0} \text{ .... (2)}$$

Where H_M = applied magnetizing field. The simple expression (1) for τ under constant voltage conditions can be obtained from Mr. Kaposi's equation (8)* replacing μ_2 by $\frac{B_A + B_B}{N/l(T_B - T_A)} = \Delta B/\Delta M$ and assuming $B_A \simeq B_R$ and $B_B \simeq B_M$ (see Fig. 1). Nevertheless equation (8)* in its original form is not useful.

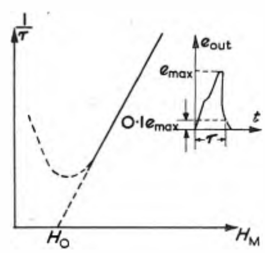


Fig. 2. The graphical representation of $1/\tau$ vs H_M under constant current magnetizing conditions. obtained experimentally.

The intersection of the extrapolated line with t axis gives the H_0 value. The switching time τ is proved immediately by oscilloscopic observation of the output voltage pulses, induced during magnetization

(2) I do not agree with the statement (see the article under discussion, p. 278) pertinent to the equation marked by us as (2): as H_0 cannot be calculated very easily, the utility of the equation is limited. When H_M is very large, H_0 can be neglected and the equation reduces to $\tau H_M = S_w$. In this form it can be used to deduce the switching time from the magnitude of the switching field . . . It is in my opinion not fully true, because: (a) The value of H_0 can be easily found by the extrapolation of the experimentally obtained line relating $1/\tau$ with magnetizing field H_M as it is shown in Fig. 2.

(b) H_0 is an important physical parameter of the core material and cannot be neglected under constant voltage switching conditions, when H_0 and H_M are very often of the same order of magnitude. The average magnetizing current I_M (related simply to H_M) under constant voltage magnetizing conditions can be found, if we take into account the physical realm of the flux reversal process, caused by action of a resulting field being the difference between H_M and H_0. Then combining equation (1) and (2):

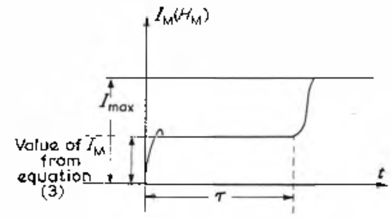


Fig. 3. The characteristic I_M/H_M vs time under constant voltage magnetization conditions.

During the magnetization the magnetizing current value is limited by the latitude of the dynamic magnetization loop. After the magnetization process has been accomplished, the I_{max} value is only limited by the sum of ohmic resistances in the driver source circuit

$$H_M = \left[\frac{S_w . V_0 . 10^8}{2\,N\,A\,B_s} + H_0\right] \text{ Oe;}$$

$$\text{and } I_M = 2{\cdot}5/\pi \; \frac{H_M . l}{N} \text{ (A) .... (3)}$$

B_s in gauss, V_0 in volts, A in square centrimetres, S_w in Oe sec^{-1}. The significance of symbols being as previously explained. The experimental evaluation performed under constant voltage switching conditions shows (Fig. 3) that the magnetizing current exhibits a sharply defined peak. This can be explained by the presence of a rotational magnetization process as well as by air transformer coupling effect, which occurs at the beginning of the flux reversal. Thus the magnetizing current value I_M derived from H_M (see equation (3)) is an average giving a good approximation.

(3) It can be easily proved that, in equation (9)* of Mr. Kaposi's article the switching time τ is left out. Further equation (10)* is not incorrect, but it is a definition rather than an equation useful for calculation. The original relationship presented by Mrs. Sands[1] determining the

* *The equations marked by asterisk are quoted from Mr. Kaposi's article.*

equivalent energy resistance of a core magnetized is the following:

$$R_{ec} = \frac{0{\cdot}4\pi\,(B_R + B_M)\,N^2A}{H_M\,.\,l\,.\,\tau}\,.\,\omega^{-s}\,[\Omega] \ldots (4)$$

This equation is pertinent to the constant current magnetization conditions; nevertheless by substituting into equation (4) the value of H_M given by equation (3), we can find the relationship pertinent to the constant voltage conditions. This is the following:

$$R_{ev} = \frac{0{\cdot}4\pi\,N^2}{(S_w\,.\,\omega^s/2\,A\,B_s + H_o/V_o')l}\quad[\Omega] \quad\ldots\ldots\ldots\ldots\ldots (5)$$

Where V_o' is the voltage per turn applied, other symbols being unchanged. High rectangularity of the core material ($B_R \simeq B_M \simeq B_S$) is assumed. The great influence of H_o at moderate values of V_o' is evident.

To sum up: the understanding of dynamic core behaviour is as yet not satisfactory. The only paper dealing with the problem generally and regarding to terminal properties of the core is the excellent work of Papoulis and Chen[2]. For many practical purposes however a knowledge of simplified average parameters is useful and sufficient. It is of great importance in both theoretical and practical considerations, to evaluate carefully the proper starting point for every case to be solved. A more exhaustive discussion on the same properties of bistable magnetic elements is beyond the scope of this letter, and is the subject of the authors book[3].

Yours faithfully,

A. GORAL,

Zegrze, Poland.

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1. SANDS, E. A. The Behaviour of Reactangular Hysteresis Loop Magnetic Materials Under Current Pulse Conditions. *Proc. Inst. Radio Engrs.* *40*, 1246 (1952).
2. PAPOULIS, A., CHEN, T. C. Domain Theory in Core Switching. Proc. Symp. on the Role of Solid State Phenomena in Electric Circuits. N.Y. (1957).
3. GORAL, A. Magnetic Contactless Relays. To be published by Ministry of National Defence of Poland.

The Author Replies:

DEAR SIR,—I was very interested to read Mr. Goral's comments and was glad to learn that in his yet unpublished book he will give us an exhaustive discussion on the subject of bistable magnetic elements: I can only hope that his book will be available for us here in England as well.

By and large, I agree with Mr. Goral's remarks and we seem to come to the same general conclusions. However, I would like to make the following comments.

(1) Equation (8)* in the form of equation (1a) is indeed more practical. Nevertheless I would not say that equation (8)* is not useful.

(2) The finding of H_o by extrapolating is of little interest. Having obtained the curve by measurement, equation (2) is solved and there is no need to know the value of H_o. Further, my remarks quoted by Mr. Goral referred only to constant current switching.

(3) The switching time τ was left out of equation (9)* by mistake, for which I apologize.

Yours faithfully,

J. KAPOSI,

Electronic Switching Division,
Ericsson Telephones Limited,
Nottingham.

Local Feedback in Transistor Amplifiers

DEAR SIR,—I was surprised at the large errors Mr. Pfyffer obtained when using the expression I quoted in my previous letter (April issue). (Expression C in Mr. Pfyffer's letter.) I seem to have obtained much better agreement between theory and practice. I am enclosing an example of my measured results (see Fig. A) and I should be very grateful if you could include these in the published correspondence for comparison with Mr. Pfyffer's measurements.

With reference to Fig. B in Mr. Pfyffer's reply (April), the equivalent circuit I have used is essentially that of Fig. 12 in the article "Equivalent Circuit for Junction Transistors", Mullard Technical Communications No. 25, p. 151, with the assumption that the feedback term u may be neglected. Hence the capacitor in Mr. Pfyffer's Fig. B should be C_c rather than $C_c\beta$.

Yours faithfully,

D. G. W. INGRAM,
Harrow, Middx.

The Author Further Replies:

DEAR SIR,—The parameters of the transistor used by Mr. Ingram which are listed beside his measured figures, indicate that the equivalent circuits used by the author in the article and by Mr. Ingram in his letters are not the same. The fact that Mr. Ingram used capital letters in his formulae, suggested that he was referring to the 'Double Base Resistance' equivalent circuit (see Fig. 16 in Mullard Tech. Com. No. 25) while he was actually using the circuit of Fig. 12. The table comparing the results in the reply to Mr. Ingram's first letter is therefore somewhat misleading.

Further measurements have been carried out by the author, using a similar

Fig. A. Frequency response of common emitter stages

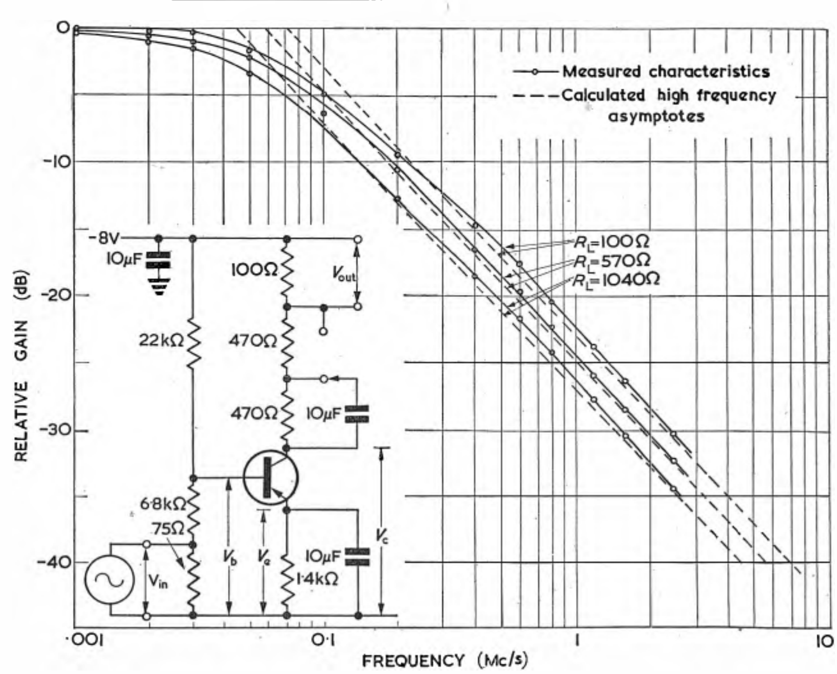
D.C. conditions $V_b = -1{\cdot}8V$, $V_e = -1{\cdot}65V$, $V_c = -6{\cdot}6V$

Transistor OC44 for which $R_b = 100\Omega$, $C_c = 11pF$, $f_\alpha = 10Mc/s$, $\beta_o = 195$

For the circuit as above $R_s = 5\,200\Omega$, $R_e = 22\Omega$

The calculated 'cut-off frequencies are then

R_L	100Ω	570Ω	$1\,040\Omega$
f_c	45kc/s	58·5kc/s	73kc/s



circuit as shown in Fig. 12 in the article (Mr. Ingram's circuit is also of the same type). Using formula (40) of the article under the assumption that $R_L \ll R_d$:

$$f_a \simeq f_B \frac{R_s + R_b + R + R_e}{R_s + R_b + (1 + \beta)(R + R_e)}$$

And referring to the 'Double Base Resistance' equivalent circuit, errors of up to 50 per cent are obtained, the results being on the low side compared with the measurements. This deviation is caused by the large value of R_b. If the circuit of Fig. 12 in Mullard Tech. Com. No. 25 is used but neglecting the capacitances, the same formula can again be used for f_γ, substituting R_b by $r_{bb'}$ and R_e by r_e $(=25/I_e)$ since the formula (40) was derived without considering the capacitances, and the equivalent circuit is of the same structure. In this case the calculated cut-off frequencies are 10 to 30 per cent on the high side.

With reference to Fig. B in the reply to Mr. Ingram's first letter, the capacitor across R_d would be C_e only if the current generator was defined in terms of i_e instead of i_b; also R_d would have to be replaced by $R_c = R_d/1 - \alpha$; hence Fig. B is correct.

In conclusion the author appreciates Mr. Ingram's contribution and agrees that his formula is more accurate and is useful if all the parameters of the transistor are known.

Yours faithfully,

H. Pfyffer,

General Electric Company Ltd,
Coventry.

New Technique for Multi-Track Recordings

Dear Sir,—The advent of general purpose magnetic tape decks with several parallel information tracks has led to many new applications for the storage, replay and analysis of data in a manner which calls for the use of at least two tracks, and often more, simultaneously. For instance, the required path of a machine tool cutter may be recorded as the phase modulation of a convenient carrier relative to an unmodulated carrier (or one modulated in reverse phase) on another track. In another example readings from several instruments and from a clock may be recorded on parallel tracks for later study of their time relationships.

In each case the information may to some extent be distorted if either on record or on replay, the line of heads is not at the standard angle of 90° to the travel of the tape. This error can be overcome if one of the recorded signals, and preferably one which is a simple sine wave, is recorded on duplicate tracks near the outer edges of the tape.

In the simple case where there are only two signals, one will be recorded at both edges and one at the centre of the tape. If the head is tilted, one of the outer channels will be advanced, and the other retarded, relative to the timing of the centre track, but if the two outer head signals are added together they will provide a reference signal in which the errors are very nearly compensated. The limitation is now eased to the extent that the edge tracks can be mistimed to, say, plus or minus forty electrical degrees of the clock signal before the reference becomes inadequate.

If there are several live information channels, as well as the duplicated reference channels at the edges of the tape, the signals from the two reference heads can be mixed in different proportions to generate the reference for the various data tracks. For instance a track 30 per cent across the width of the tape should be demodulated or otherwise interpreted with regard to a reference generated by adding 70 per cent of the first head signal to 30 per cent of the last.

For any but a central data track the compensation is not quite accurate, but as the errors depend on the third or higher powers of the electrical error angle, they are quite small for angles appreciably less than one radian.

Yours faithfully,

R. E. Spencer,

E.M.I. Electronics Ltd.,
Hayes, Middlesex.

A Drip-Rate Recorder for Intraveneous Solutions

Dear Sir,—I have read with great interest the article by Melville and Cornwall in the October 1958 issue and I thought you might possibly be interested in an automatic drip rate controller for intraveneous solutions which was developed at the Department of Telecommunications of the Swiss Federal Institute of Technology (Prof. Hch. Weber), following ideas of Dr. G. Hossli, Head of the Department of Anaesthesia at the Chirurgische Universitäts-Klinik (Prof. Dr. Brunner), Kantonsspital Zürich. The development was supported by a grant from the Fritz Hoffmann-La Roche-Stiftung zur Förderung wissenschaftlicher Arbeitsgemeinschaften in der Schweiz.

The instrument, which was built 1953/54, uses the same basic switching circuit as Messrs. Melville and Cornwall, relying on the voltage/time-relationship in an RC-combination. The voltage across C_2 (see Melville and Cornwall, Fig. 2) is sampled immediately before S_3 discharges C_1 and C_2 (in our set-up, S_2 remains closed during this phase). The sampled voltage is applied through a cathode-follower to one coil of a sensitive polarized relay. The other coil is fed from a reference source, for reasons of drift correction also through an identical cathode-follower. The relay armature assumes a neutral position (not making either contact) when the two voltages are equal; if the interval between two subsequent drops is too long, the voltage across C_2 rises above the reference voltage and the relay armature closes a contact on one side. Power is applied to the relay armature only during the sampling interval; a voltage output is thereby obtained on one set of contacts for drip rates exceeding a predetermined value, and on the other set of contacts for a drip rate slower than that. These voltages are used to actuate a small servomotor, which in turn operates a lead screw squeezing the rubber tubing of the infusion system.

A photoelectric arrangement was used to register the drops. A number of fail safe devices are incorporated into the equipment to make prolonged unattended operation possible. The drip rate can be selected between 6 and 180 drops/minute. The drip rate is kept at the selected value against changes in viscosity, venous pressure or position of the bottle within ±5 per cent.

A number of warning devices are included which should alert the nurse or an attendant in case of any malfunctioning, such as:

(1) When the infusion stops.
(2) When the liquid level in the drip vessel varies considerably (indicating a leak in the system and danger of air embolism).
(3) When the lead screw has travelled full way to either limit position.
(4) When a failure in the electrical system occurs.

The instrument was used on a number of patients, during surgery and post-operationally. In one case it was used for one week on one patient without interruption. It was particularly valuable for administering low drip rates (such as 6 per minute), which is very difficult to do with manual control.

An undesirable feature of the instrument is the noise associated with the operation of relays which are actuated at every drop. It would certainly be possible today to completely transistorize the equipment, make it noiseless in operation and considerably reduce bulk and weight.

Yours faithfully,

M. Muller,

Switzerland.

The Authors reply:

Dear Sir,—Our drip rate recorder was designed to assist in a study of the relationship between the drip rate of intravenous pitocen and the rate of contraction and the tone of the uterus. The question of drip rate controlling was considered separately after the above relationship had been established and we considered that control by feedback from a rate sensing circuit was unnecessarily complicated. In its place we used a variable speed, positive displacement pump sold under the trade name of 'Sigmamotor', with some modifications to the drive and equipped with safety devices. We appreciate Dr. Muller's interest in our equipment and it is clear that our rate sensing circuit is similar to his. Drip rate control by the positive displacement pump has been entirely satisfactory, and we think it is somewhat simpler than the feedback control system.

Yours faithfully,

A. W. Melville and J. B. Cornwall,

Industrial Development Laboratories,
Auckland, New Zealand.

BOOK REVIEWS

Linear Network Analysis

By S. Seshu and N. Balabanian. 571 pp. 371 Figs. Medium 8vo. John Wiley and Sons Inc., New York. Chapman and Hall Ltd, London. 1959. Price 94s.

HERE, at last, is a textbook which presents the modern approach to linear electric network analysis at a tutorial level. The need for such a book has been felt for some time. The authors are professors of electrical engineering in Syracuse University and the book is based on the lecture notes of the graduate course on network analysis given at that university for the past three years.

The book begins with an introduction to the electric network concept. The two postulates (Kirchoff's 'laws') upon which network theory is built are discussed in detail and the elements of an electric network are defined. In Chapter 2, the loop and node methods of analysis are described. A knowledge of the Laplace transformation is assumed, but the book contains an appendix devoted to theory of functions of a complex variable and the Laplace transformation.

Matrix algebra and the topology of linear graphs are the subjects of the third chapter. It is a combination of these two branches of mathematics which provides the basis to the modern approach to network analysis. The incidence matrix representation of the network connexions is introduced and, in the fourth chapter, it is shown how the incidence matrices combine with matrices representing the electrical properties of the network to yield a complete analysis. Several pages are devoted to the important question of determining initial conditions as 'right-hand' limits. The use of impulse functions in this connexion is also considered.

Chapter 5 deals with network theorems and the steady-state response of networks to both sinusoidal and periodic excitation. The convolution integral and DuHamel's integral are introduced in the following chapter as means of determining the response of a network to an arbitrary excitation function from a knowledge of its impulse response or indicial response. These results are then brought together as a set of uniqueness theorems and the chapter concludes with a discussion of the relationships between frequency-response and time-response.

The representation of network behaviour on the 'complex frequency' plane is the subject of Chapter 7. The topics considered are: Bode diagrams; minimum phase and non-minimum phase transfer functions; the calculation of network functions from given magnitude, angle, or real part; integral relationships between the real and imaginary parts of a network function; and the potential analogue. Chapter 8 is concerned with two-port network analysis. The two-port parameters, including the scattering parameters, are introduced and the matrix theory of interconnected two-ports is explained. In Chapter 9 the analytical properties of passive network functions are considered at length. Positive real functions are defined and studied and Foster's reactance theorem is presented and extended to the general single-energy case. The last section of the chapter contains an interesting treatment of the topological significance of network determinants. Chapter 10 deals with feedback structures, signal-flow graphs, stability criteria and root locus techniques, and the final chapter is concerned with the image-parameter theory of filter networks. Associated with each chapter is an extensive list of problems.

There is ample evidence that network analysis is still regarded in technical teaching circles as the study of filters in the sinusoidal steady-state. As the authors put it, network theory has developed from a collection of approximations which were useful in a large number of practical situations to the status of an exact science. The modern concept of electric network is a dynamical system of passive and active elements and modern network analysis is the study of the reaction of such systems to applied electric forces. It will be interesting to see if this book (and those which will surely follow it) has any effect on the teaching of network theory in this country.

S. R. DEARDS

Mathematics in Physics and Engineering

By J. Irving and N. Mullineux. 873 pp. 104 figs. Medium 8vo. Academic Press Inc., New York. 1959. Price $11.50.

THE reviewer remembers that many years ago Dr. Irving talked of the preparation of this book and since then he has several times wondered when it would appear. Now that it is in print it is clear that the authors have made very good use of the time by using their greater experience in its preparation and incorporating many recent applications of the methods they discuss. The book shows all the signs of very careful preparation and the authors and publishers are to be congratulated on the presentation, and print.

This book is suitable for engineering and physics students and graduates at a fairly advanced level, somewhere between that of a first-year university course mathematics and that required by the comprehensive books on mathematical methods by Jeffreys and Jeffreys, or by Morse and Feshbach. It fits admirably into this gap and the text includes examples from engineering and physics to illustrate the application of each method to a problem where it is needed. At the end of each chapter there is a selection of more problems, some elementary and others more advanced.

The book starts with the solution of the more important partial differential equations, such as the wave equation, and this leads naturally to the solution of some ordinary differential equations and a discussion of the properties of various types of Bessel and Legendre function though, oddly enough, there is no mention here or later of the Bessel functions with imaginary argument, which occur frequently in electrical engineering. Matrices and determinants are well described, together with practical methods of solving simultaneous equations illustrated in detail, and then there is a chapter on classical and wave mechanics. (This chapter does not fit in very well with the main theme but furnishes good examples of the use of various techniques.) There are two chapters on transforms, Laplace and others, with excellent illustrations of their use, two on complex variables, and one on numerical methods.

Special mention ought to be made of the chapters on the calculus of variations and on integral equations which are important topics often inadequately discussed. Here there is a successful attempt to explain these topics to the initiated as far as is possible in the space of a hundred pages, without sacrificing clarity.

To sum up, this book is a really good example of how to describe quite advanced techniques in analysis and illustrate them by well chosen applications. It is suitable as a text-book and should be bought by engineers and physicists who are curious about these techniques. It may appeal more to the engineer.

G. K. KYNCH

Transistor Circuits

By K. W. Cattermole. 442 pp. 306 figs. Demy 8vo. Heywood & Co. Ltd. 1959. Price 70s.

THIS book (as its title implies) is devoted almost entirely to transistor circuits, the chapters on semiconductors and the electrical properties of transistors being very short. It is intended for engineers who already have a knowledge of thermionic valves, and electrical circuits. And for these readers it should prove quite useful.

An almost inevitable feature of books on transistor circuits is that statements about the limitations of available transistors are out of date by the time they are published. This is true of the book under review, thus on page 2 it is stated that "In frequency, the range of commercially available transistors is from d.c. to about 20Mc/s". When making a statement such as this it is desirable that the time at which the statement was valid should also be given.

The subject of amplifiers is covered in five chapters—single stage, multi-stage, power, high frequency and bias stabilization—the subject is well covered, but only one example of a practical circuit—an audio amplifier—is given. The avoidance of circuits with given component values, is a deliberate policy of the author, who considers that this information is best obtained from technical journals.

The subject of pulse circuits is introduced by a chapter on negative resistance. A rather surprising feature of this and following chapters, is the amount of space devoted to an obsolete device—the point contact transistor.

The chapter on binary circuits deals with the transistor as a switch and the large signal behaviour of junction transistors as well as bistable circuits. Sine wave oscillators are included in the chapter on waveform generation, together with monostable and astable circuits. The use of pulse circuits as counting and timing devices is considered in the following chapter.

The last three chapters in the book are on—"Modulation detection and frequency changing" "Measurement of transistor parameters" and "Fields of application".

A. F. NEWELL

Progress in Nuclear Energy—Physics and Mathematics

Edited by D. J. Hughes, J. E. Sanders & J. Horowitz. 403 pp. 176 figs. Medium 8vo. Pergamon Press Ltd. 1959. Price 105s.

PERGAMON Press have for some time now been issuing books covering all the main aspects of nuclear technology. There are 12 series in all, each covering a different aspect and contributed to by the most prominent workers in the appropriate field. The above book is the third volume to be issued in the series devoted to the physics and mathematics of nuclear energy. This particular volume itself does not contain previously unpublished matter, as it consists of a selection of the more important papers delivered at the second Geneva Conference held in September 1958, relating to reactor physics. The selection of papers while representing only a small fraction of the total number delivered at the Conference, has nevertheless been so skilfully made that the reader is given a well balanced cross-section of the material covered at the conference.

A notable feature of the volume and which fairly reflects the trend of current thinking in reactor physics, is the relatively large proportion of space devoted to the process of thermalization and the associated problem of spectrum determination. Of a total of 21 papers in the volume, five are devoted to this topic and of these four are of a theoretical nature, while the fifth, deals with the measurement of neutron spectra in uranium water lattices.

As one expects the theoretical treatment of neutron thermalization is one attended by great complexity of both mathematics and physics. The physical complexity of the problem arises from the fact that unlike the slowing down process a detailed knowledge of the thermal motion of the moderator nuclei is required. The problem is further complicated by the fact that this thermal motion both in liquids and solids depends strongly on the interactions binding the moderator nuclei. A particularly detailed model for taking account of binding effects in crystalline graphite is contained in a contribution from the U.K.

An interesting phenomenological approach to thermalization in water is contained in a Russian contribution. They suggest that binding effects can be taken into account by using the monatomic gas model kernel, with a fictitious mass adjusted to depend on energy in such a way that agreement with experimental scattering cross-sections are obtained. They claim that this approach gives the correct transport mean free path.

Among other papers of theoretical interest is a Swedish contribution that discusses the calculation of temperature coefficients of resonance integrals. This topic, in common with thermalization is one that will gradually assume more and more importance as specific ratings of reactors are increased.

Doppler and self shielding effects are treated rigorously in an American contribution, using transport theory. There is a further excellent American paper which discusses the application of the Schwinger variational method to the inhomogenous Boltzmann equation. Thermalization and space energy problems are among those discussed. The remainder of the book is mainly devoted to review articles dealing with cross-sections and their determination and the fisson process.

While most of the items in the book will previously have been available to workers in the field, it is nevertheless convenient to have them together in handy book form. The value of the book is somewhat reduced by the fact that the complete official proceedings of the Geneva Conference containing the papers reviewed above will be readily available soon. The book would furthermore be more attractive if its price were substantially lower.

A. Z. KELLER

Advances in Electronics and Electron Physics Volume XI

Edited by L. Marton. 499 pp. 165 Figs. Medium 8vo. Academic Press Inc., New York, London. 1959. Price $15.00.

REVIEW volumes of this nature are an essential part of modern scientific literature. The various sections written by leading experts contain the distilled essence of hundreds of articles and highlight as well as summarize the important advances in a subject. A very welcome feature of this volume is that the articles are by Russian, German and Australian scientists as well as by leading authorities in the United States. It is to be hoped that this international representation will become a characteristic feature of the science of the future.

The subjects dealt with are of great practical as well as theoretical importance. Görlich reviews recent advances in photoemission from metals and semiconductors throughout the region from the short wavelength vacuum ultra-violet to the near infra-red. The behaviour of compound photo-cathodes is discussed in terms of their band structure and it appears that the application of the modern concepts of solid-state physics is giving a new insight into the problems of this subject. Somewhat related is the view by Hachenberg and Brauer on secondary electron emission from solids and a similar effort to relate the empirical facts to modern theories has been made in this review.

In the article by Clark Jones on the quantum efficiency of detectors chief emphasis is given to photo-emissive detectors, photo-conductive detectors, television camera tubes, photographic negatives and the human eye. He introduces the concept of detective quantum efficiency in order to define the detecting performances of a wide variety of detectors.

Reviews on automatic data processing by Barlow, Ovenstone and Thoneman, on radio telemetering by Riblet and on operational amplifiers by Konigsberg could hardly be more topical in this age of rockets and satellites. Of a more specialized and fundamental nature is the article on parity nonconservation in weak interactions by Sternheimer.

Structural studies by electron diffraction are reviewed by Pinsky and many beautiful illustrations show the advanced techniques of the Russian school. While little reference is made to the excellent work of other schools this can be excused as the boot has so often been on the other foot and the advantage of having this account in English has to be set against this small deficiency.

W. C. Price

Information Transmission, Modulation and Noise

By M. Schwartz. 454 pp. 309 Figs. Medium 8vo. McGraw-Hill Publishing Co. Ltd. New York, London. 1959. Price 85s. 6d.

MUCH change has taken place in methods of communication during the last two or three decades, and the exploitation of frequency modulation and the various types of pulse modulation has made it necessary for the present day student to learn much more of the scope and limitation of these systems as compared with amplitude modulation. Most books in this field are written for post-graduate study. The present volume is aimed at an earlier level, and the treatment is, in consequence, more detailed.

Following a brief, but necessary, preamble about the information content of messages, chapter two deals with the properties of Fourier series and integrals, and their use in determining the transmission of signals of various kinds through networks, and the limiting effect of bandwidth, i.e. non-zero response time, on the information that may be transmitted. The next two chapters deal with the theory and practice of amplitude, frequency, and pulse modulation and demodulation.

Chapter five introduces the other system limitation, viz. noise arising from spontaneous fluctuations of voltage and current in the circuit components. A very good discussion is given of shot noise, thermal noise, signal-to-noise ratio, and noise figure at low frequencies, including the effects arising from the use of transistors.

With this general basis, the book then turns to the effects of the various modulation systems on noise, and signal-to-noise ratio, and the improvements that can be obtained by trading bandwidth for noise, as in wideband frequency modulation systems, and by pulse code modulation. The discussion of the statistical methods of analysis necessary for a deeper understanding of the effects of noise in practical systems is undertaken in the final chapter.

The author has well borne in mind the educational purpose of the book, and many problems for solution by the student are included at the end of each chapter. The continual use of bracketed phrases, which appear on most pages, is unusual, and would have been better avoided, as they could be interpreted as afterthoughts. The brief mention of Laplace transforms in the discussion in chapter two does not seem necessary, and would appear to assume a higher standard of knowledge than does the rest of the book. These are, however, minor criticisms of a book which admirably serves its purpose in stressing the basic unity in transmission systems.

W. H. Aldous

Masers

By J. R. Singer. 141 pp. 42 Figs. Medium 8vo. John Wiley & Sons Inc., New York, Chapman & Hall Ltd, London, 1959. Price 52s.

DR. SINGER'S book is the first to be published on this new subject of masers.

The field of solid state electronics of which this is part is one which requires contributions from the theoretical physicist, the experimental physicist and the microwave engineer. It is somewhat disappointing to find that the book, admittedly described as an introductory treatment, is directed at none of these groups but appears to be directed at the post-graduate student of physics.

An historical introduction is followed by a theoretical discussion of induced emission and absorption. This is the least satisfactory chapter in the book; the treatment being confusing to one unacquainted with the phenomena and unhelpful to one who, knowing the fundamentals, wishes to apply them to masers. The other theoretical chapter—on the treatment of electron para-magnetic resonance in solids—also suffers from shortcomings of the same kind, although not to the same extent.

Gas maser devices, in particular the ammonia maser, are treated in detail and there are adequate descriptive treatments of both two-level and three-level solid state cavity masers.

Travelling wave masers are dealt with in a short final chapter which provides only a superficial introduction—it is a pity that this part of the subject, which is of very great importance for the future, is treated so briefly.

It is somewhat unfortunate that the physical basis of maser noise is neglected at the expense of the equivalent circuit treatment of noise from the device as a whole. Here, however, Dr. Singer omits to mention a most important contribution to the system noise—the effect of loss in the input signal line.

The reviewer would venture to disagree with the discussion on experimental equipment for cavity masers (Section 6.8). Two things should, he feels, be made quite clear. As has already been proved in practice, packaged solid-state masers can be made extremely compact and trouble-free—so that they can be left unattended at the focus of a radio-telescope, the provision of the refrigerant for the required low temperature being no more than a minor inconvenience. On the other hand, laboratory versions of masers with ancillary electronic equipment for experimental purposes are bulky and heavy—but here the requirements are, as in every experimental apparatus, for flexibility and ease of adjustment to the changing experimental conditions.

On particular points in the text—the statement on page 16 that for electron spins aligned in a magnetic field "the parallel orientation is a higher energy state than the antiparallel orientation" is surely not correct. It would seem more valuable to have combined Appendices A and B and derived expressions for simple three-level maser operation, and the various double pumping schemes possible all on the basis of a system containing four magnetic energy levels—because the chromium ion which has been used extensively possesses four levels.

The presentation of the text is extremely readable and is amply illustrated with well drawn diagrams, many taken from original papers. It is largely free from errors, although the units of magnetic field are somewhat capricious, in one instance (pp. 80-81) both gauss and oersteds are used on adjoining pages. The bibliography is extensive and there is an adequate index.

The book is well worth reading by all interested in masers, but it does not provide an authoritative handbook on the subject for those working in the field.

S. A. Ahern

Laboratory Instruments

By A. Elliott and J. Home Dickson. 490 pp. 227 figs. Demy 8vo. 2nd Edition. Chapman & Hall Ltd. 1959. Price 55s.

THIS new edition has been completely revised and much material added, but, as in the previous edition, the authors have confined themselves almost wholly to mechanical and optical matters.

The introduction of the subject of corrosion-resisting metals is among the additions. The appendix on optical crystals has been enlarged into a chapter with the latest possible information on the refractive index of materials useful in infra-red spectrometry. A complete new section on radiation has necessitated another new chapter which includes the section on photometry. New matter has also been added on colour vision, to the diagrams of photographic lenses, and in the chapter on photography.

The Instruments, Electronics and Automation Exhibition

A description, compiled from information supplied by the manufacturers, of some of the exhibits at the Instruments, Electronics and Automation Exhibition to be held at Olympia, London from 23 to 28 May.

(Voir page 321 pour la traduction en français: Deutsche Übersetzung Seite 328)

A.P.T. ELECTRONIC INDUSTRIES LTD

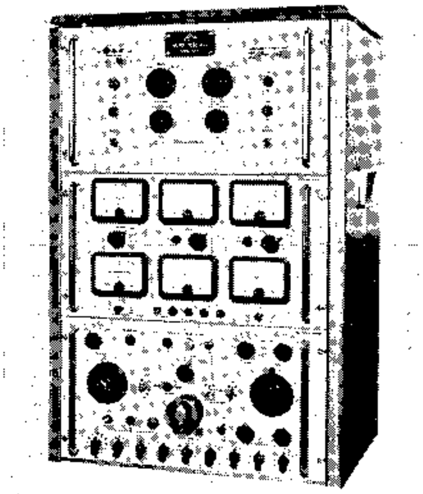
50-52 King Street, Southall, Middlesex

VALVE TESTER

(Illustrated below)

This test set incorporates stabilized power supplies, multi-range metering for the examination of all valve parameters, and an a.c. bridge for g_m measurement. These three sections are built into 19in rack units, each with a 10½in panel, and supplied in an instrument case.

Valves can be tested to 'CV' or other specifications up to the limits of the power supplies, and mutual conductance to 100mA/V. The performance of small



signal diodes (single and double), power rectifiers (half and full wave), h.t. metal rectifiers and gas-filled stabilizers can also be measured.

The metered power supplies are available for separate use when the test set is not being operated, and jack points allow external meters to be plugged in for calibration purposes requiring extreme precision.

The power supplies provided are as follows:

Anode Supply

0-300V (in 3 ranges) stabilized to 0·1 per cent

0-250mA (metered 1/10/30/100/300mA f.s.d)

Screen Supply

0-300V (in 3 ranges) stabilized to 0·1 per cent

0-50mA (metered 1/10/30/100mA f.s.d)

Grid Supply

0-100V neg. and 0-25V pos. Stabilized to 0·1 per cent (metered 100-0-100μA f.s.d)

Heater Supply

0-55V a.c. (in 4 ranges, each 25W max) (metered 10/3/0100V f.s.d and 0·5A f.s.d)

EE 21 751 *for further details*

ADVANCE COMPONENTS LTD

Roebuck Road, Hainault, Ilford, Essex

TRANSISTORIZED POWER UNITS

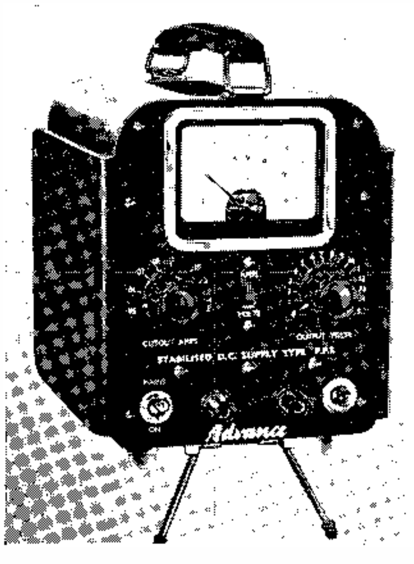
(Illustrated below)

The power pack type PP. 5 is a small compact transistorized instrument providing a stabilized d.c. voltage supply at a maximum current of 0·5A; the output voltage being continuously variable in one directly calibrated range from 0 to 15V.

Protection against current overloads including short-circuits is performed electronically and is conveniently reset from the front panel. In addition to this, the operating point of the overload protection circuit is adjustable to provide protection for the load circuit in use (that is, protection against thermal runaway in amplifier, etc.). Protection is also effected against excessive ambient conditions at maximum load.

Metering is performed by a switched dual range meter measuring both output current and voltage.

These units will operate from a mains



ELECTRONIC ENGINEERING will occupy stand number L.502 at the I.E.A. Exhibition and visitors will be welcome

input of 200 to 250V a.c. The output impedance is less than 0·01Ω at d.c., and less than 0·2Ω up to 100kc/s.

EE 21 752 *for further details*

AMPHENOL (GREAT BRITAIN) LTD

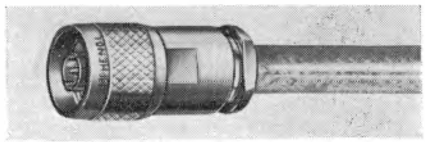
Victoria Road, Burgess Hill, Sussex

R.F. CONNECTORS

(Illustrated below)

The American 'C' and 'N' type r.f. connectors are now being produced in British versions with an improved cable clamping mechanism to British Military requirements.

These are available with alternative bushes to take Uni-Radio 102 and 73, 67 or 81 cables and all have 'captivated' contacts. Both types are being made as straight and right-angled plugs, free jacks



and panel jacks and except for the clamping mechanism they are equivalent to 'C' types Nos. UG-573/U, UG-710A/U, UG-570/U, UG-572/U and UG-571/U and 'N' types Nos. UG-21B/U, UG-22B/U and UG-23B/U. All have a nominal impedance of 50Ω. Joint Service Pattern Numbers are being issued for these components.

A special spanner is needed to assemble these connectors.

EE 21 753 *for further details*

ASSOCIATED ELECTRICAL INDUSTRIES LTD

Rugby, Warwickshire

ANALOGUE COMPUTER

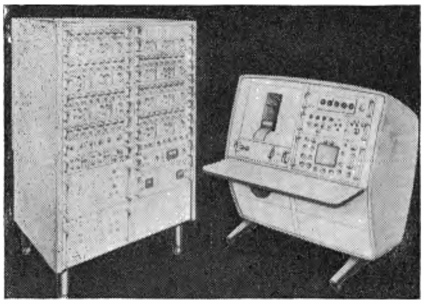
(Illustrated next column)

The AEI 955 analogue computer will solve linear and non-linear differential equations and can be used for the study of the dynamic performance of physical systems.

The computer offers facilities for setting-up linear functions (adding and integrating etc) and non-linear functions (dead-zone/limit, arbitrary functions and multiplying, etc) by use of standard sized computing units. Thus the size and layout of the computer may be arranged to suit any particular problem, setting up being achieved by 'patch-cords' across the front of the computer.

The computing impedances may be of 1·0 per cent or 0·1 per cent accuracy according to choice. The computing amplifier, which is continuously drift-corrected, is characterized by high gain, wide bandwidth, low drift, low grid current, and low noise.

The mode of operation of the computer, which may be either continuous



or repetitive, is selected by controls which are integral with a high accuracy (±0·5 per cent) dual trace oscilloscope. An auto-hold unit allows the computer to be put into the 'hold' condition after any pre-specified time or at any predetermined problem voltage. A permanent record of results may be obtained from the two-channel pen-recorder which is of the rectilinear write-out type and uses heat-sensitive paper as the recording medium.

EE 21 754 *for further details*

B & R RELAYS LTD

Temple Fields, Harlow, Essex

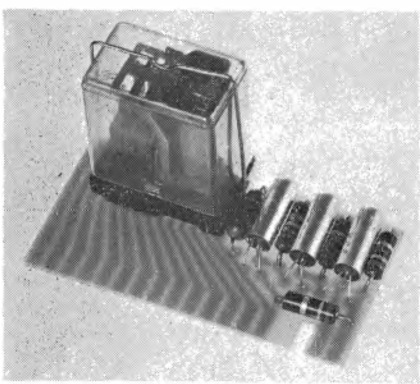
TRANSISTORIZED POWER RELAY

(Illustrated below)

This transistorized relay is designed for switching power where only a small operating signal is available. It is a three stage silicon transistor device and requires a power supply of 30mA at 12V d.c.; it operates with a current of 15μA applied to the control circuit and will withstand an overload of 100 times. The maximum ambient operating temperature is 60°C.

The overall dimensions of this printed circuit transistorized relay are 2in by 1½in by 1¼in and the relay itself plugs into the amplifier. The contacts and connectors are gold plated throughout. The printed circuit fits into a standard polarized 18 way socket.

EE 21 755 *for further details*



BELLING & LEE LTD

Great Cambridge Road, Enfield, Middlesex

FIXING AND MOUNTING TECHNIQUE

The 'Prestincert' process is a completely new method of mounting components of all kinds on panels and chassis of any material, on die castings etc, or of riveting laminæ together with no preparatory punching or drilling operations whatsoever.

This process produces joints which are mechanically and, where this applies, electrically of very high efficiency, and can therefore be used with equal success in the electrical, electronics and all engineering industries wherever rivets are normally applicable. In the electronics industry, for example, this method lends itself to the mounting and wiring of components on printed circuit boards in one simple operation, rendering soldering optional; it can be carried out either as a series of single point operations, or on a simultaneous multi-stage basis, according to the dictates of the job.

For large scale production the "Prestincerts" and components may be hopper fed to the fixing sites and the mounting and wiring thus fully automated. For small runs, hand fed presses are used and for single units portable hand tools may be employed.

EE 21 756 *for further details*

BLACKBURN ELECTRONICS LTD

Brough, Yorkshire

HALL EFFECT MULTIPLIERS

(Illustrated above)

These multipliers (types BIE 293 and BIE 294) have been designed to operate in conjunction with analogue computers and instrumentation systems.

The units satisfy the equation:—

$\dfrac{XY}{10^3} = Z$ (for BIE 293) $XY.10^3 = Z$ (for BIE 294)

They use the Hall Effect in semiconductors and are transistorized.

The multipliers have an accuracy of ±0·5 per cent of full scale and a frequency response from d.c. to 100c/s.

EE 21 757 *for further details*

TRANSISTORIZED BINARY CLOCK

This unit generates 1sec timing pulses with an accuracy of 0·05 per cent. Each 1sec pulse is uniquely identified by an associated 16 bit binary code group for any second in a 9 999sec period. The pulses are suitable for recording on magnetic tape or other systems.

When the recorded time signals are played back on the decoder (BIE 417 not on display) it is possible to identify any 1sec interval irrespective of the speed or direction in which the tape is run.

EE 21 758 *for further details*

"DIAMOND H" SWITCHES LTD

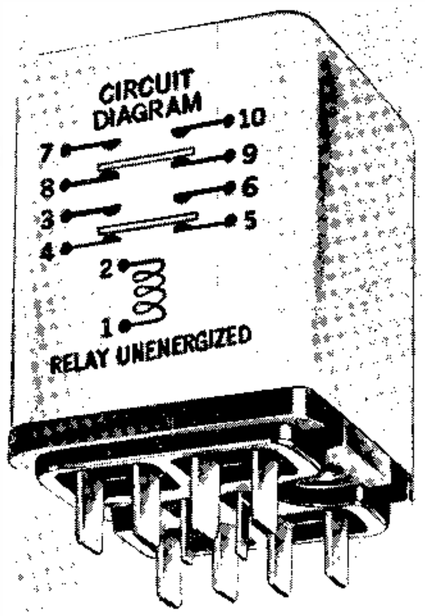
Gunnersbury Avenue, London W.4

RELAYS

(Illustrated below)

The series R relays are miniature, hermetically sealed units for aircraft use and have an operating temperature range of −65°C to 200°C. Weighing 4oz or less, they offer vibration resistance up to 2kc/s at 20g and operating shock resistance to over 50g.

The BW series relays (illustrated) are general purpose, low cost units measuring 1½in by 1¼in by 1⅞in. They are rated at 25A resistive at 250V a.c., and 1 h.p. at 115V a.c. or 2 h.p. at 250V a.c. They are



available for side, panel or plug-in mounting with various a.c. and d.c. coils.

EE 21 759 *for further details*

EKCO ELECTRONICS LTD

Southend-on-Sea, Essex

MOISTURE GAUGE

A transistorized instrument, the new Ekco moisture gauge is primarily designed for determining the moisture content of paper. It normally operates at the dry-end of a paper-making machine, utilizing a self-balancing capacitance bridge to measure the effective capacitance of the paper plus moisture. No contact is made with the paper and the instrument is self-compensating for variations in spacing of the measuring components. Measurements in the range 2 to 15 per cent moisture content are obtainable and the repeatability is ±¼ per cent. An output signal of 0 to 15mA is available for control purposes and a direct reading of 'percentage moisture content' is provided. When coupled to a standard Ekco beta gauge, the moisture gauge will provide continuous compensation to ensure a true 'dry weight' reading on the beta gauge.

EE 21 760 *for further details*

ENGLISH ELECTRIC VALVE CO. LTD

Chelmsford, Essex

TRANSMITTING VALVES

Three new triodes, types BR1131, BR1138 and BW194, have been added to the range of E.E.V. Co. high power transmitting valves.

BR1131 is a forced-air cooled triode with characteristics similar to those of the BR152B (CV28) but employs a thoriated tungsten filament, rated at 8·5V, 22A, and has a maximum anode dissipation of 3·5kW.

BR1138 is a forced-air cooled triode identical with BR152B (CV28) apart from the filament which is of thoriated tungsten and rated at 8·5V, 22A. The BR1138 may be used as a direct replacement for the CV28 in existing equipments, the difference in filament voltage being readily accommodated by connecting the filaments of pairs of BR1138 in series.

BW194 is a water cooled version of the well established BR194, with an anode dissipation of 50kW.

The main details of these valves are tabulated below:

TYPE	V_f (V)	I_f (A)	I_k (pk) (A)	μ	g_m (mA/V)	V_a (kV)	P_a (kW)	f (Mc/s)
BR1131	8·5	22	4	40	3·1	10	3·5	15/80
BR1138	8·5	22	2	40	3·1	10	1·0* 1·25†	15/80
BW194	13·0	240	100	34	38·0	15	50·00	5/30

* *Natural Cooling* † *Forced-Air Cooling*

EE 21 761 *for further details*

EVERSHED AND VIGNOLES LTD

Acton Lane, London W.4

RECORDER-CONTROLLER

The new Evershed three term recorder-controller is contained in two small units and requires only a standard panel cut-out of 144mm by 144mm. Despite these small dimensions a 4in chart display has been achieved through suitable curving of the front of the panel mounted part of the instrument. A 'desired value' indicator, and adjustment knobs for various control functions, are also carried by this section. Transistorized circuits are used and all power requirements (including a transistorized stabilized power source) are contained within the two units.

EE 21 762 *for further details*

GOODMANS INDUSTRIES LTD

Axiom Works, Wembley, Middlesex

VIBRATION GENERATOR

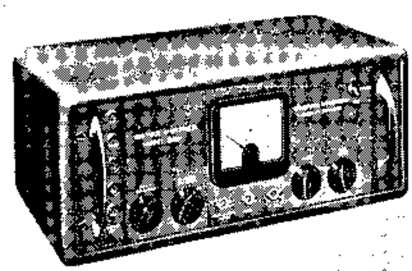
The new VG.109 Mk II vibration generator is capable of thrusts up to 18 000lb and employs oil cooling instead of the air cooling of the earlier machines. The table diameter of this generator is 23in and the mass of the moving coil and table assembly is approximately 100lb. The maximum stroke is ±1in and the static force factor approximately 28lb/A. The thrust available for 10kW dissipation is 7 500 to 10 000lb and that for 30kW 13 000 to 18 500lb. The annular table top is sealed for climatic chamber operations.

EE 21 763 *for further details*

CONTROL AMPLIFIER

(Illustrated below)

Designed to complete the Goodmans vibration system, the control amplifier model E.501 is capable of controlling vibration amplitude, acceleration, or velocity to a level, which can be pre-set, or programmed from an external source.



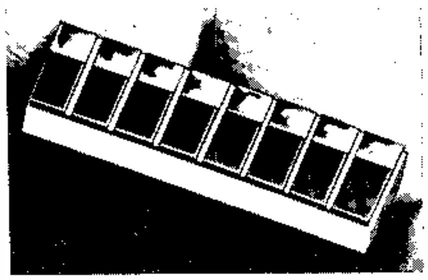
The unit consists of a special input stage, a head amplifier, a metering circuit, an error detector stage, a compressing amplifier and a cathode-follower. An error of 25dB can be compressed to about 4dB under suitable circumstances. The amplified and rectified output from a suitable vibration pick-up is compared with a predetermined direct voltage, to derive a signal indicative of the error of the vibration level from the desired value. The error signal controls the compressing amplifier, to which an external sine wave oscillator supplies alternating input thus giving a sine wave signal of frequency determined by the oscillator setting, but of such amplitude that it will produce the correct level of vibration (after passing through a power amplifier if necessary). The working frequency range is from 20c/s to 10kc/s, but response can be obtained beyond these limits.

Input sockets are provided for five signal sources, which can be switch selected at the input of the special cathode-follower; for crystal accelerometers of sensitivity from 10 to 20mV/g the meter sensitivity adjustment can be set for full scale values of 10g and 50g respectively on the two ranges provided.

The special design of the input cathode-follower enables crystal accelerometers to be used with cable lengths up to 100ft without disadvantageous results. A co-axial socket is fitted which provides an output from the cathode-follower stage and is suitable for connecting directly to any conventional cathode-ray oscilloscope without additional impedance conversion.

An oscillator input signal of 150mV r.m.s. is suitable and a maximum output of 4V r.m.s. can be provided by the amplifier. At an output level of 1V the total harmonic distortion is less than 2 per cent.

EE 21 764 *for further details*



INTERNATIONAL RECTIFIER CO. (Great Britain) LTD

Hurst Green, Oxted, Surrey

SILICON READ-OUT PHOTOCELLS

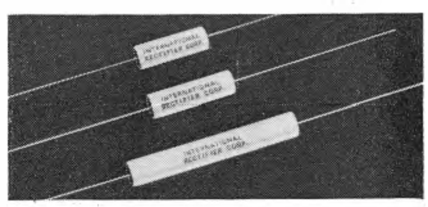
(Illustrated above)

These silicon photo-voltaic read-out matrices have a response time of the order of 10μsec and are specifically designed for computer and data processing equipment where rapid detection of light passing through punched cards or tape is required.

Each matrix is made up of a multiple array of individually segmented silicon cells. Light energy striking a particular segment will cause power to flow from that segment only. Typical current generated is 300μA for 0·01m² of active cell area at 1 000ft-candles illumination.

Flexibility of design enables these units to be supplied in single cell or multiple-cell assemblies, as required. They are characterized by negligible temperature dependence, long operating life, uniform response from cell to cell in one matrix, and rugged construction.

EE 21 765 *for further details*



HIGH VOLTAGE RECTIFIERS

(Illustrated above)

This series of silicon cartridge rectifiers covers the range from 600V to 10kV p.i.v. The rectifiers are ceramic encased and are suitable for use at altitudes up to 90 000ft. These units are available for d.c. outputs between 75 to 250mA (at 25°C) and have an operating temperature range of −55°C to +150°C. The maximum reverse current at rated p.i.v. and 25°C is 10μA.

EE 21 766 *for further details*

KELVIN & HUGHES (INDUSTRIAL) LTD

Kelvin House, Wembley Park Drive, Wembley, Middlesex

DATA HANDLING SYSTEM

On view for the first time will be a data handling system designed to give the greatest flexibility of application and capable of scanning and logging an unlimited number of process variables to an accuracy of 0·1 per cent of reading and at higher speeds than previously available with such an equipment.

The standard input is a direct current or voltage, so that the application potential of the equipment is only limited by the availability of suitable transducers. Such transducers are readily available for applications involving pressure, flow, strain, pulsed and a.c. outputs, while thermocouples serve temperature applications. Cold junction compensating units are provided for each type of thermocouple input.

Representation of information is in digital form throughout. High and low limit checking with alarm indicators, range variations, zero suppression and inbuilt linearization, as required for thermocouples, flow, etc., are available for integration into the scanning sequence. Logging is by electric typewriter on blank or pre-printed log sheets, off-normal points being printed in red.

A digital voltmeter is used to measure the process variable and present the result in digital form. It contains a sequencing unit, relays, two-stage amplifier, and power supplies. The latter includes a supply stabilized within 0·05 per cent for measurement reference purposes. Voltages are sequentially switched and compared with the input voltage until a final out-of-balance value within 0·1 per cent is achieved.

The system, including transducers and power supplies is housed in a cubicle type panel, the units being rack-mounted for easy removal and servicing.

EE 21 767 *for further details*

GEORGE KENT LTD

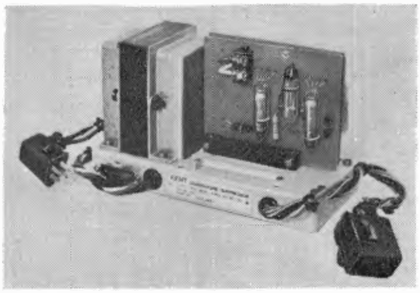
Luton, Bedfordshire

QUADRATURE SUPPRESSOR

(Illustrated above right)

The type 1 quadrature suppressor is designed for use with the George Kent range of electronic instruments when applied to the measurement of a.c. signals.

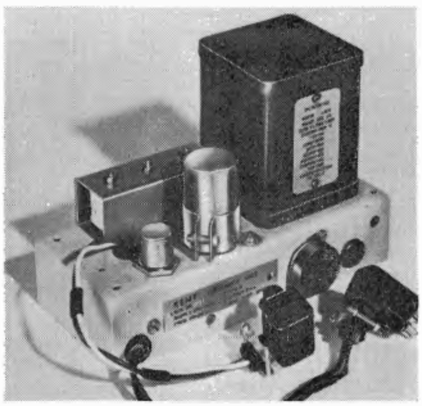
The suppressor is a small plug-in electronic assembly, which eliminates the possibility of amplifier saturation and errors in measurement arising as a result of the presence of a quadrature component in the input signal. A high degree of reliability and robustness is achieved by using only transistors, thermistors and other solid-state components. The principal part of the suppressor is the KENT-patented thermistor potentiometer, which comprises a plug-in printed-circuit carrying the main components. These are collectively comparable to the amplifier, balancing motor and slidewire of a conventional servo system.



The new unit effectively combats the quadrature effects that frequently occur in a.c. servo systems such as those measuring the electrolytic conductivity or the rate of flow of liquids. These spurious signals can have very detrimental effects, for any quadrature components present in the input of an a.c. servo system cannot be reduced to zero by the main a.c. feedback, and give rise to an irreducible error signal. This in turn may cause an error in the output of the system or may cause partial or even complete saturation of the amplifier and resultant loss of sensitivity.

The quadrature suppressor provides a solution to the problem by applying an automatic cancelling action to any incoming quadrature components, whether stable in magnitude or not, while in no way interfering with the main input signal. The residual quadrature error signal, derived from the servo amplifier, is amplified and demodulated by a circuit containing four low-power germanium transistors and used to control differentially the power supplied to the heaters of two indirectly-heated thermistors. The two thermistor beads form a voltage divider of variable gain controlling a signal of precise quadrature phase (derived by a 90° shift from the a.c. reference) which is fed back to the input of the servo amplifier and suppresses the residual signal. The circuit is direct-coupled from end to end and a third thermistor provides over-all negative d.c. feedback from the output to the input. This stabilizes the operating points of all four transistors without the use of any capacitors, and maintains the output characteristics of the thermistor potentiometer substantially constant in ambient temperature between 0° and 60°C.

EE 21 768 *for further details*



VOLTAGE REFERENCE UNIT

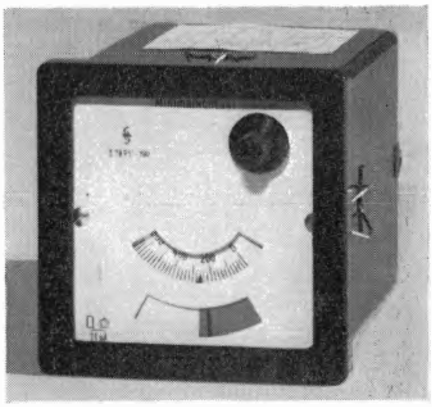
(Illustrated below left)

The type 3 reference unit is a solid-state d.c. reference source designed and produced by George Kent Ltd for use with the Company's range of electronic self-balancing recorders, indicators, and controllers. A compact plug-in assembly operating directly from the normal a.c. mains supply, it provides an extremely stable, ripple-free output of 5·0mA at 5V for the slidewire of the instrument measuring circuit.

Two Zener-diode stages in cascade reduce the effects of normal supply variations by a factor of at least 200, while individual temperature compensation of each of these stages ensures that the output is substantially independent of changes in ambient temperature between 10° and 70°C. An inherent advantage is that the unit continues to operate with only slightly reduced accuracy if the supply voltage is drastically reduced, even to as little as 30 per cent of its nominal value.

The d.c. output is isolated from earth, and a specially screened and balanced mains transformer effectively isolates it from the mains supply. This feature is of great value in certain measuring circuits having a finite impedance to earth.

EE 21 769 *for further details*



KYNMORE ENGINEERING CO. LTD

19 Buckingham Street, London W.C.2

HIGH SENSITIVITY RELAY

(Illustrated above)

These relays, manufactured by Siemens and Halske, may be used where signals in the picowatt region have to be monitored or where they have to be capable of initiating or stopping a controlled process and where reliably repeatable positive contact operation within closely preset limits is required.

The operating principle is as follows. When the preset value is reached the contact is positively operated at the expense of only very minute power dissipation. This results from combining the operation of the core-magnet strip-suspended moving-coil armature with a magnetic contact-pressure amplifier. Even when extremely slow gradient signals form the input, the contact will close with snap-action. The pointer of the strip-suspended armature contains a short piece of iron wire. In the preset operating bandwidth, a small permanent magnet forms a stop at the adjustable limit of deflexion. When the input signal increases, the pointer moves towards the magnetic contact-pressure amplifier until the strip-suspension torque is balanced by the core-magnet flux. At this stage, the smallest signal increase causes the pointer to snap over to the magnetic stop where a contact pressure of about 2g will be maintained. As the contact surfaces of pointer and magnetic stop are gold-plated, safe contact will be re-established even after prolonged storage or inoperation. The pointer-contact energizes a built-in slave relay whose coil is supplied from a 24V d.c. source. The contacts of the slave are rated to work into industrial contactors or alarms. In this way various separate circuits may be controlled. The slave relay can be supplied with a number of different contact combinations. One of its contacts is used as a hold-on contact in parallel with the relay-pointer contact. In this way contact is maintained until a manual push-button switch is operated to re-set the relay. The manual re-set button performs a dual function in that it first energizes a drop-off coil whose armature pulls away the pointer contact which is 'dead' since it is in parallel to the hold-on contact of the slave relay. Only afterwards the coil circuit of the slave is broken. Thus no arcing can take place at the pointer contacts. If on operating the 're-set' the original signal still persists, the relay will immediately re-energize the slave relay.

These relays have an external adjustment control for setting the operating point and a linearly divided scale. They are available in a range of current and voltage ratings, the maximum sensitivity available being 10^{-10}W.

EE 21 770 *for further details*

J. LANGHAM THOMPSON LTD

176 High Road, Bushey Heath, Hertfordshire

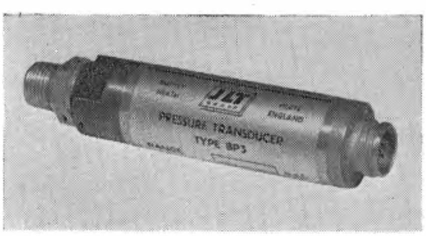
PRESSURE TRANSDUCER

(Illustrated below)

Designed for the measurement of both steady and rapidly fluctuating pressures, the frequency response of this transducer makes it specially suitable for the investigation of transients in hydraulic and pneumatic systems.

Models are available covering the ranges 0 to 2 000lb/in² to 0 to 20 000lb/in² gauge pressure with high overload rupturing pressures.

The pressure sensing element consists of a cylinder of beryllium copper partially bored-out to form a tube with a solid end, having two active resistive strain gauge windings over the bored-out section, and two temperature compensating windings over the solid section. These are connected as a four-arm balanced bridge network.



Pressure applied within the cylinder causes dilation of the bored-out section with resultant resistive changes in the active windings, thus with an a.c. or d.c. voltage applied to the bridge, a signal proportional to pressure is derived. The full range output is 1·5mV/V applied, the maximum operating voltage being 30V. The overall accuracy is better than 0·5 per cent including linearity and hysteresis.

EE 21 771 *for further details*

LAURENCE, SCOTT & ELECTROMOTORS LTD

Mansfield House, 376 Strand, London W.C.2

CONSTANT SPEED UNIT

This unit is capable of maintaining a shaft rotation at a speed which is constant to an accuracy better than one part in three thousand over the full rated range of the driving motor. Two types are available, one for use with a small induction motor as a prime mover and the other with a d.c. permanent magnet motor. The unit provides a speed reference signal which is obtained from a pick-up coil, the poles of which are mounted near the periphery of a toothed-wheel coupled directly to the driving motor shaft. At the designed constant speed the pick-off signal is 50kc/s. This signal is amplified and fed to a crystal filter circuit resonant at just over 50kc/s. The output of this filter is rectified and ultimately controls a cathode-follower output stage. In the case of the d.c. drive, the motor is connected directly in the cathode circuit, but with the a.c. version the cathode-follower load is the control winding of a magnetic amplifier which in turn controls the motor speed.

EE 21 772 *for further details*

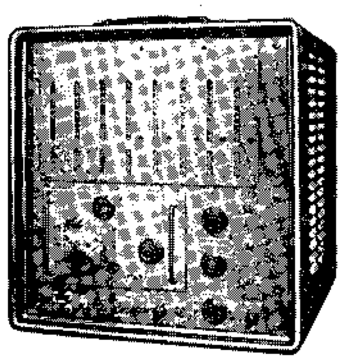
MARCONI INSTRUMENTS LTD

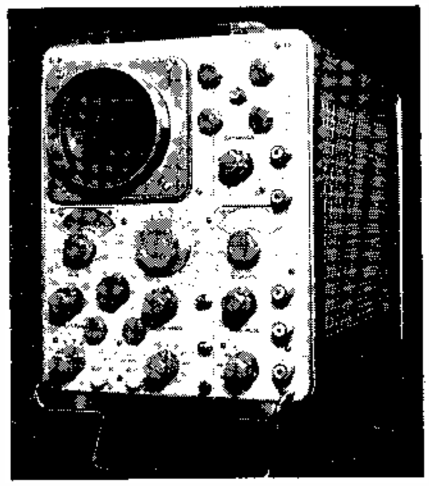
St. Albans, Hertfordshire

10Mc/s COUNTER

(Illustrated below)

The 10Mc/s electronic counter type





TF1345 is a high-speed counter-timer with built-in precision frequency standard. Read-out is by neon indicators on an 8-decade digital display. The instrument counts up to 10 millions per second; measures frequency from 10c/s to 10Mc/s, period of waveforms up to 100kc/s. A selection of plug-in accessories extends the frequency range to 220Mc/s, allows time measurement down to 1μsec and increases sensitivity to 10mV. Display time control is either manual, or continuously variable from 0·1 to 10sec with automatic and repetitive resetting. The equipment is available for bench or rack mounting.

EE 21 773 *for further details*

DUAL-TRACE OSCILLOSCOPE

(Illustrated above)

The TF 1331 is a d.c. coupled dual-trace oscilloscope with direct-reading time and voltage calibration. Electronic beam-switching enables two independent signals to be displayed either on alternate sweeps or at a switching rate of 100kc/s. Either input can also be displayed separately without switching. The frequency response extends from d.c. to 15Mc/s, and sensitivity is variable in seven steps from 50mV/cm to 50V/cm. The time-base sweep velocity is variable from 1sec/cm to 0·1μsec/cm in fifteen ranges, and can be increased up to 0·02μsec/cm using X expansion. Triggering can be applied internally or externally, or from an internal supply-frequency source. A 10MΩ input probe is available as an optional accessory. A single beam version, type TF 1330, is also available.

EE 21 774 *for further details*

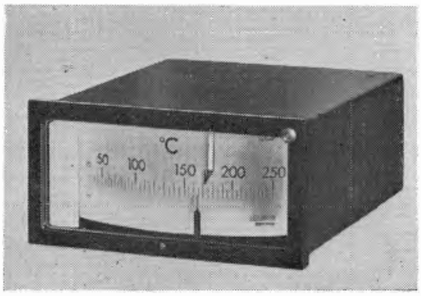
METRIX INSTRUMENTS LTD

54 Victoria Road, Surbiton, Surrey

PROCESS CONTROLLER

(Illustrated next column)

The Metrawatt transistorized controller offers an inexpensive yet accurate method of control of temperature, pH, conductivity, power and other variables whose function can be translated in terms of an electrical voltage.



The control system employs one or more high frequency oscillator circuits whose coupling coils are mounted on adjustable slides which can be set to any desired control values on the instrument scale. A small vane on the moving-coil meter arm enters the air-gap of the oscillator coils when the pointer reaches the values of the pre-set control positions. The oscillator circuits in turn operate mercury or high power contactors as required to effect the ultimate control of the system. Since there are no mechanical contacts on the meter, the movement can therefore be extremely sensitive giving a high accuracy of control (0·2 per cent) with a fast response.

A simple thermal feedback system is available if required, which converts the instrument into a proportional controller; the effect of thermal inertia in the controller system is then almost completely eliminated.

The instrument, which is fully transistorized, embodies a protection circuit which ensures that the control supply is switched off in the event of a fault either in the thermocouple or tranducer or in the instrument itself.

The effect of mains voltage fluctuations is only 0·1 per cent for a mains change of 15 per cent.

EE 21 775 *for further details*

MUIRHEAD & CO. LTD

Beckenham, Kent

Precision R.M.S. Decade Voltmeter

(Illustrated below)

The Muirhead-Wigan D930A is a very accurate voltmeter having a wide voltage

range (1mV to 300V) and a wide frequency range (5c/s to 100kc/s). Being a true r.m.s. voltmeter, it is capable of measuring voltages having sinusoidal and non-sinusoidal waveforms.

Voltage readings are taken directly from four decade controls marked 'units' (1-10) 'tenths' (0-10), 'hundredths' (0-10) and 'thousandths' (0-10) and a six-position range switch covering the multiplying ratios from ×0·001 to ×100.

Dial reading accuracy is thus always possible to 0·1 per cent and readings can be interpolated to 0·025 per cent.

The constant input resistance of 100kΩ makes low voltage measurement possible with low power consumption. Even at 300V the input power dissipation is less than 1W.

The tentative specification of this instrument is as follows:

Accuracy of voltage measurement

30c/s-5kc/s	60mV-10V: ±0·05 per cent (except at the supply frequency) 1mV-300V: ±0·1 per cent or 30μV (whichever is greater)
20c/s-20kc/s	1mV-300V: ±0·2 per cent or 30μV (whichever is greater)
5c/s-100kc/s	10mV-300V: ±0·25 per cent up to 50kc/s decreasing to ±1 per cent at 100kc/s (with the aid of a stable external oscillator such as the D-890-A/2 for standardizing at the test frequency)

Accuracy of voltage ratio measurement on any one range

Ranges: 100V-300V 10V-100V, etc., down to 10mV-100mV	±0·03 per cent to 0·15 per cent (according to ratio and voltage) up to 50kc/s (not worse than ±0·2 per cent at supply frequency)
Range 1mV-10mV	±0·05 per cent to 0·15 per cent (according to ratio) up to 10kc/s (not worse than ±1 per cent at supply frequency)

Input impedance 100kΩ shunted by 200pF (max)

Maximum peak/r.m.s. ratio of input waveform 3:1

EE 21 776 *for further details*

SHORT BROTHERS & HARLAND LTD

Seaplane Works, Queens Island, Belfast, Ireland

Analogue Computer

This new general purpose computer, known as the 'Simlac Minor', is of interest to potential users who require greater capacity than is available in Shorts' 18-amplifier general purpose computer, but who do not require the capacity or higher accuracy of the recently announced large Simlac computer.

The instrument has a capacity of 32 amplifiers and 100 passive elements. Like the larger version, Simlac Minor is of standard rack construction, the entire computer being housed in two 19in Post Office type racks. The left-hand rack contains the computing elements and the right-hand rack houses the instrumentation, removable patch panels, power units and three special non-linear units.

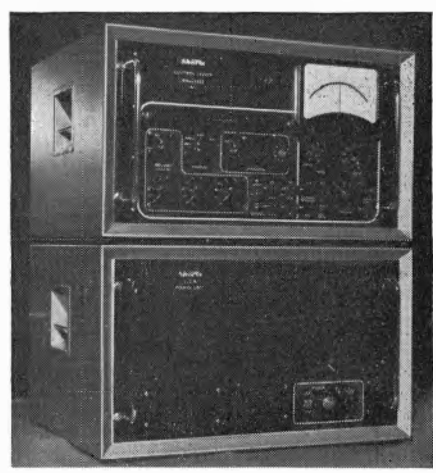
To obtain maximum utilization of the space available oven-stabilization is not employed. Nevertheless, by careful design and choice of components the overall accuracy of the computing elements is 0·1 per cent. The amplifiers and passive elements are standard Simlac equipment except that the 'hold' facility is an optional extra.

EE 21 777 *for further details*

L.F. Oscillator and Control System Analyser

(Illustrated below)

This low frequency oscillator is designed for low frequency testing where



extremely low d.c. content and accurately maintained amplitudes are of utmost importance. The amplifiers are continuously corrected for drift so that the d.c. content of the output is normally well below 100μV.

The control system analyser is unique in its ability to operate at unusually low frequencies operating successfully at a frequency as low as one cycle per 100sec. Its range extends up to 1kc/s and the instrument is capable of investigating a broad range of systems.

The analyser incorporates the low frequency oscillator and presents steady readings of in-phase and quadrature components even at the lowest frequency of 0·01c/s. The d.c. level is always less than 0·005 per cent and, by careful balance of the drift-corrected amplifiers, it can be maintained at 0·0005 per cent.

EE 21 778 *for further details*

(Pages 321 to 334 French and German Section)

Le Salon de L'AUTOMATION DE L'ELECTRONIQUE ET DES INSTRUMENTS

Une description basée sur des renseignements fournis par les fabricants de nouveaux organes, accessoires et instruments d'essai

Traduction des pages 315 à 320

A.P.T. ELECTRONIC INDUSTRIES LTD

50-52 King Street, Southall, Middlesex

CONTRÔLEUR DE LAMPES

(Illustration à la page 315)

Cette appareil de contrôle comporte des alimentations en courant stabilisé, un dispositif de mesure multi-gammes pour l'examen de tous les paramètres de lampes et un pont C.A. pour la mesure de g_m. Ces trois sections sont incorporées dans des blocs à châssis de 48,2 cm, ayant chacun un panneau de 26,6 cm, et elles sont livrées en boîtiers pour instruments.

Les lampes peuvent être examinées conformément aux spécifications "CV", ou d'autres spécifications, jusqu'aux limites d'alimentation et une conductibilité mutuelle de 100 mA/V. On peut aussi mesurer le rendement de diodes à faibles signaux (simples et doubles), les redresseurs de puissance (demi-onde et onde entière), les redresseurs secs H.T. et les stabilisateurs à atmosphère gazeuse.

Les prises de courant mesurées peuvent être employées séparément lorsque l'appareil de contrôle n'est pas en usage, et des prises de jack permettent de brancher des appareils de mesure extérieurs pour des travaux d'étalonnage nécessitant une très grande précision.

Les prises de courant suivantes sont prévues:

Alimentation anodique:
- 0-300 V (en 3 gammes), stabilisés à 0,1%.
- 0-250 mA (mesurés: 1/10/30/100/300 mA déviation totale d'echelle).

Alimentation d'écran:
- 0-300 V (en 3 gammes), stabilisés à 0,1%.

Alimentation de grille:
- 0-100 V nég. et 0-25V pos., stabilisés à 0,1% (mesurés: 100-0-100 μA, déviation totale d'échelle).

Alimentation de chauffage:
- 0-55 V C.A. (en 4 gammes, chacune de 25 W max.) (mesurés: 10/30/100 V, déviation totale d'échelle et 0,5 A, déviation totale d'échelle).

EE 21 751 *pour plus amples renseignements*

ADVANCE COMPONENTS LIMITED

Roebuck Road, Hainault, Ilford, Essex

BLOCS D'ALIMENTATION TRANSISTORISÉS

(Illustration à la page 315)

Le bloc d'alimentation type PP.5 est un petit instrument transistorisé fort compact, fournissant une alimentation en tension continue stabilisée à un courant maximum de 0,5 A, la tension de sortie étant à variation continue dans une gramme étalonnée directement de 0 à 15 V.

La disjonction à maxima, y compris les court-circuits, s'effectue électroniquement et elle peut être réenclanchée commodément à partir du panneau avant. De plus, le point de fonctionnement du circuit de disjoncteur à maxima est réglable de manière à protéger le circuit d'utilisation (c'est à dire, le protéger contre les fuites thermiques dans l'amplificateur, etc.). La protection est également assurée contre les conditions ambiantes excessives à la charge maxima.

La mesure se fait par appareil de mesure interconnecté à double gamme, mesurant tant le courant de sortie que la tension.

Le bloc PP.5 fonctionne à partir d'une entrée secteur de 200 à 250 V C.A. L'impédance de sortie est inférieure à 0,01 ohm au courant continu, et elle est inférieure à 0,2 ohm jusqu'à 100 kHz.

EE 21 752 *pour plus amples renseignements*

AMPHENOL (Great Britain) LTD

Victoria Road, Burgess Hill, Sussex

CONNECTEURS HAUTE FRÉQUENCE

(Illustration à la page 315)

Les connecteurs haute fréquence américains, type C et N, sont actuellement produits en versions anglaises, avec un mécanisme perfectionné de serrage de câble, conforme aux exigences militaires britanniques.

Ces serre-fils sont munis de douilles alternatives pour câbles Uni-Radio 102 et 73, 67 ou 81, et ils sont tous pourvus de contacts fixes. Les deux types sont fabriqués sous forme de fiches droites et rectangulaires, de jacks libres et de jacks de panneau et, à l'exception du mécanisme de serrage, ils sont équivalents aux types C Nos. UG-573/U, UG-710A/U, UG-570/U, UG-572/U et UG-571/U, ainsi qu'aux types Nos. UG-21B/U, UG-22B/U et UG-23B/U. Ils ont tous une impédance nominale de 50 ohms. Des numéros de modèles militaires sont attendus incessamment pour ces pièces.

Une clef anglaise spéciale est nécessaire au montage de ces connecteurs.

EE 21 753 *pour plus amples renseignements*

ELECTRONIC ENGINEERING occupera le Stand No L502 au Salon des Instruments, de l'Électronique et de l'Automation, et les visiteurs y seront les bienvenus.

ASSOCIATED ELECTRICAL INDUSTRIES LTD

Rugby, Warwickshire

CALCULATEUR ANALOGIQUE

(Illustration à la page 316)

Le calculateur analogique AEI 955 résout des équations différentielles linéaires et non-linéaires, et il peut être appliqué à l'étude du comportement dynamique de systèmes physiques.

Il permet la mise en régime de fonctions linéaires (addition, intégration, etc.) et non-linéaires (angle mort/limite, fonctions arbitraires et multiplication, etc.) par l'utilisation d'éléments de calcul de dimensions courantes. On peut, ainsi, adapter les dimensions et la disposition du calculateur aux besoins de n'importe quel problème, la mise en régime s'effectuant par cordons à travers l'avant du calculateur.

Les impédances de calcul peuvent être d'une précision de 1,0 ou de 0,1%, selon le choix. L'amplificateur de calcul, à glissement corrigé continûment, se caractérise par son grand gain, une largeur de bande étendue, un glissement réduit, un courant de grille bas et un faible niveau de bruit.

Le mode de fonctionnement du calculateur, qui peut être soit continu soit à répétition, se choisit par commande faisant corps avec un oscilloscope à double affichage de haute précision (±0,5%). Un élément d'auto-fixage permet de mettre le calculateur à l'état "fixe", après n'importe quelle période de temps pré-spécifié, ou à n'importe quelle tension de problème prédéterminé. Un enregistrement permanent de résultats peut être obtenu grâce à l'enregistreur à plume à deux canaux, du type à écriture linéaire et utilisant du papier sensible à la chaleur comme moyen d'enregistrement.

EE 21 754 *pour plus amples renseignements*

B. & R. RELAYS LTD

Temple Fields, Harlow, Essex

RELAIS DE PUISSANCE TRANSISTORISÉS

(Illustration à la page 316)

Ce relais est prévu pour le branchement de puissance lorsqu'il n'existe qu'un seul petit signal de fonctionnement. Il consiste en un dispositif à transistors au silicium à trois étages, exigeant une alimentation de 30 mA à 12 V C.C. Il fonctionne sur courant de 15 μA, appliqué au circuit de commande, et peut résister à une surcharge 100 fois supérieure. La température de service ambiante maxima est de 60°C.

Les dimensions hors-tout de ce relais transistorisé à circuit imprimé sont de 50,8 mm × 38,1 mm × 31,7 mm) et le relais lui-même peut être branché sur l'amplificateur. Les contacts et les connecteurs sont entièrement dorés. Le circuit imprimé s'adapte à une douille de 18 voies polarisées de type courant.

EE 21 755 *pour plus amples renseignements*

BELLING & LEE LTD

Great Cambridge Road, Enfield, Middlesex

TECHNIQUE DE MONTAGE ET DE FIXATION

Le procédé Prestincert constitue une méthode entièrement nouvelle de montage de pièces de toutes sortes sur panneaux et châssis en n'importe quelle matière, sur coulages en coquille, etc., ou de rivetage de tôles, ne nécessitant absolument aucun travail préparatoire de poinçonnage ou de perçage.

Le procédé produit des joints d'une très grande efficacité mécanique et, le cas échéant, électrique. Il peut donc être utilisé avec un égal succès dans les industries électriques, électroniques et mécaniques, c'est à dire dans n'importe quel cas où des rivets sont normalement appliqués. Dans l'industrie électronique, par exemple, la méthode se prête au montage et au câblage de pièces sur plaquettes de circuits imprimés, en une simple opération, rendant ainsi le soudage facultatif. Elle peut s'effectuer soit par une suite d'operations à point unique, soit sur base de multi-étages simultanés, selon les exigences du cas.

Pour la production sur une grande échelle, les Prestincerts et les composantes peuvent être livrés par trémie aux emplacements de fixage, le montage et le câblage devenant ainsi entièrement automatiques. Pour les petits parcours, on emploiera des presses actionnées à la main, tandis que pour les éléments isolés, on peut utiliser des outils manuels portatifs. Le procédé est des plus souples, ne nécessitant pas un matériel capital coûteux. Dans les industries aéronautiques, automobiles et de constructions navales, ses applications sont quasi infinies, avec une efficacité de production et une robustesse mécanique que n'égale aucun autre procédé de rivetage.

EE 21 756 *pour plus amples renseignements*

BLACKBURN ELECTRONICS LTD

Brough, Yorkshire

MULTIPLICATEURS À EFFET DE HALL

(Illustration à la page 316)

Ces multiplicateurs (types BIE 293 et BIE 294) ont été conçus pour fonctionner en relation avec les calculateurs analogiques et les systèmes d'instrumentation.

Ils répondent à l'équation:

$XY/10^3 = Z$ (pour le type BIE 293)

$XY, 10^3 = Z$ (pour le type BIE 294)

Ils utilisent l'effet de Hall dans les semi-conducteurs et ils sont transistorisés. Leur précision est de ±0,5% de l'échelle totale et leur réponse de fréquence va du courant continu à 100 Hz.

EE 21 757 *pour plus amples renseignements*

HORLOGE BINAIRE TRANSISTORISÉE

Cet appareil produit des impulsions de minutage de 1 sec. avec une précision de 0,05%. Chaque impulsion de 1 sec. est identifiée par un groupe de codes connexes à 16 chiffres binaires pour une quelconque seconde dans une période de 9999 secondes. Les impulsions conviennent à l'enregistrement sur bande magnétique ou d'autres systèmes.

Lorsque les signaux de temps enregistrés sont réenregistrés sur le décodeur (BIE 417, non exposé), on peut identifier n'importe quel signal de 1 sec. quelle que soit la vitesse ou la direction de l'avance du ruban.

L'encombrement réduit et le faible poids de cette horloge la rendent propre au fonctionnement de bord, conjointement avec un système d'enregistrement de bord.

EE 21 758 *pour plus amples renseignements*

"DIAMOND H" SWITCHES LTD

Gunnersbury Avenue, London W.4

RELAIS

(Illustration à la page 316)

Les relais miniature et hermétiquement fermés de la série R, ont été étudiés à l'usage des avions et leur gamme de températures de service va de −65°C à 200°C. Ne pesant que 113 grammes ou même moins, ils présentent une résistance anti-vibratoire allant jusqu'à 2 kHz à 20 grammes, et une résistance aux chocs de fonctionnement dépassant 50 grammes.

Les relais de la série BW (figurant sur notre gravure) sont des éléments universels, à bas prix, mesurant 38,1 mm × 38,1 mm × 47,6 mm. Leur puissance nominale est de 25 A de charge résistive à 250 V C.A. et de 1 CV à 115 V C.A. ou 2 CV à 250 V C.A. Ils sont livrés pour montage de côté, sur panneau ou interchangeable, avec divers bobinages de courant alternatif ou direct. Le dispositif de contact classique est bipolaire à deux directions, mais d'autres combinaisons comportant le cadencement peuvent être fournies.

EE 21 759 *pour plus amples renseignements*

EKCO ELECTRONICS LTD

Southend-on-Sea, Essex

JAUGE D'HUMIDITÉ

La nouvelle jauge d'humidité EKCO est un instrument transistorisé prévu essentiellement pour déterminer la teneur en eau du papier. Elle est normalement mise en oeuvre à l'extrémité sèche d'une machine à fabriquer le papier, utilisant un pont de capacités auto-équilibreur pour mesurer la capacité effective du papier ainsi que sa teneur en eau. Aucun contact n'a lieu avec le papier et l'instrument est autocompensé pour les variations d'espacement des composantes de mesure. On peut mesurer des teneurs en eau de 2 à 15% et la répétabilité est de ±14%. Le contrôle se fait au moyen d'un signal de sortie de 0 à 15 mA et le pourcentage de teneur en eau peut être lu directement. Lorsqu'elle est accouplée à une jauge β Ekco de modèle courant, la jauge d'humidité assure une compensation continue de manière à garantir une lecture réelle de "poids sec" sur la jauge β.

EE 21 760 *pour plus amples renseignements*

ENGLISH ELECTRIC VALVE CO. LTD

Chelmsford, Essex

TUBES ÉMETTEURS

Trois nouvelles triodes, les types BR 1131, BR 1138 et BW 194, viennent d'enrichir la gamme des tubes émetteurs de grande puissance de la English Electric Valve Co.

BR 1131 est une triode à refroidissement forcé par air dont les caractéristiques sont semblables à celles de la BR 152B (CV 28), mais à filament de tungstène thorié d'une puissance de 8,5 V-22 A. Sa dissipation anodique maxima est de 3,5 kW.

BR 1138 est une triode à refroidissement forcé par air, identique à la BR 152B (CV 28) sauf pour le filament qui est en tungstène thorié d'une puissance nominale de 8,5 V-22 A. La BR 1138 peut être utilisée pour remplacer directement la CV 28 dans les équipements en service, la différence de tension de filament étant facilement ajustée en reliant les filaments de paires de BR 1138 en série.

BW 194 est une version refroidie

par circulation d'eau, du modèle réputé BR 194, avec une dissipation anodique de 50 kW.

Les principaux détails de ces tubes sont donnés ci-après:

TYPE	V_f (V)	I_f (A)	I_k (crête) (A)	μ	g_m (mA/V)	V_a (kV)	P_a (kW)	f (MHz)
BR1131	8,5	22	4	40	3,1	10	3,5	15/80
BR1138	8,5	22	2	40	3,1	10	1,0* 1,25†	15/80
BW194	13,0	240	100	34	38	15	50	5/30

* = *à refroidissement naturel*
† = *à refroidissement forcé par air.*

EE 21 761 *pour plus amples renseignements*

EVERSHED & VIGNOLES LTD

Acton Lane, London W.4

Enregistreur-contrôleur

Le nouvel enregistreur-contrôleur à trois termes Evershed est contenu dans deux petits boîtiers et ne nécessite qu'un panneau de 144 mm × 144 mm. Malgré cet encombrement réduit, un cadran de 101,6 mm a pu être aménagé grâce à une courbure appropriée de la partie de l'instrument montée à l'avant du panneau. Un indicateur de "valeur requise" et des boutons de réglage pour diverses fonctions de commande sont également compris dans cette section. Des circuits transistorisés sont employés et tous les dispositifs de puissance (y compris une source d'alimentation stabilisée transistorisée) sont renfermés dans ces deux éléments.

EE 21 762 *pour plus amples renseignements*

GOODMANS INDUSTRIES LTD

Axiom Works, Wembley, Middlesex

Générateur de vibrations

Le nouveau générateur de vibrations VG109 MkII peut déclencher des poussées allant jusqu'à 8167,6 kg et il est à refroidissement par huile au lieu du refroidissement par air comme les premières machines. Le diamètre de la table de ce générateur est de 5842 mm et la masse formée par le cadre mobile et l'assemblage de la table est d'environ 45,3 kg. La course maxima est de ±25,4 mm et le facteur de force statique est d'environ 12,7 kg/A. La poussée pour une dissipation de 10 kW est de 3402 à 4536 kg, et pour 30 kW elle est de 5897 à 8395 kg. Le haut de la table annulaire est scellé pour les opérations en chambre climatérique.

EE 21 763 *pour plus amples renseignements*

Amplificateur de commande

(*Illustration à la page 317*)

Conçu pour compléter le système de vibration Goodmans, l'amplificateur de commande type E.501 contrôle l'amplitude vibratoire, l'accélération ou la vitesse, à un niveau pouvant être préréglé ou établi selon un programme à partir d'une source extérieure. L'appareil consiste en un étage d'entrée spécial, un amplificateur de tête, un circuit de mesure, un étage détecteur d'erreurs, un amplificateur compresseur et un palpeur à charge cathodique. Une erreur de 25dB peut être comprimée jusqu'à environ 4dB, dans des conditions favorables. La sortie amplifiée et redressée d'un capteur de vibrations approprié est comparée à une tension directe prédéterminée, afin d'obtenir un signal indiquant l'erreur du niveau vibratoire par rapport à la valeur requise. Le signal d'erreur commande l'amplificateur compresseur, auquel un oscillateur extérieur à ondes sinusoïdales fournit une tension d'entrée alternative, donnant ainsi un signal de fréquence à ondes sinusoïdales, déterminé par le réglage de l'oscillateur, mais d'une amplitude telle qu'il produit le niveau correct de vibration (après avoir passé par un amplificateur de puissance, si nécessaire). La gamme de fréquences de service s'étend de 20 Hz à 10 kHz, mais on peut obtenir une réponse au-delà de ces limites.

Des douilles d'entrée sont prévues pour cinq sources de signaux dont la commutation peut être choisie à l'entrée du palpeur cathodique. Pour les accéléromètres piézoélectriques d'une sensibilité de 10 à 20 mV/*g*, le réglage de la sensibilité de mesure peut être effectué pour des valeurs d'échelle totale de 10 *g* et 50 *g* respectivement, dans les deux gammes disponibles.

La construction particulière du palpeur cathodique d'entrée permet d'employer les accéléromètres piézoélectriques avec des longueurs de câble allant jusqu'à 30,48 m, sans résultats défavorables. Une douille coaxiale assure une sortie de tension du palpeur cathodique et elle peut être reliée directement à n'importe quel oscilloscope cathodique de type conventionnel sans entraîner pour autant une conversion d'impédance supplémentaire.

Un signal d'entrée d'oscillateur de 150 mV efficaces est parfaitemente approprié et l'amplificateur peut fournir une sortie maxima de 4 V efficaces. A un niveau de sortie de 1 V, la distorsion nonlinéaire totale est inférieure à 2%.

EE 21 764 *pour plus amples renseignements*

INTERNATIONAL RECTIFIER CO. (Great Britain) LTD

Hurst Green, Oxted, Surrey

Cellules photoélectriques de lecture au silicium

(*Illustration à la page 317*)

Ces matrices de lecture photovoltaïques au silicium ont une durée de réponse de l'ordre de 10 µsec et elles ont été spécifiquement étudiées à l'intention des calculateurs et ordinateurs électroniques, pour lesquels il est nécessaire de pouvoir détecter rapidement la lumière passant par les cartes perforées ou les bandes magnétiques.

Chaque matrice est formée d'un assemblage multiple de cellules au silicium en segments individuels. L'énergie lumineuse frappant l'un des segments provoque un flux de puissance n'émanant que de cette seule cellule. Une donnée de courant caractéristique serait 300 µA pour 0,01 pouces² de surface de cellule active, avec un éclairage de 100 pieds-bougies.

L'adaptabilité de leur conception, permet de fournir ces éléments, au choix, en assemblages à cellule unique ou à cellules multiples. Elle se distinguent, en outre, par le fait qu'elles ne sont que peu affectées par la température, par leur longue durée de fonctionnement, leur réponse uniforme de cellule à cellule, dans une même matrice, et. enfin, leur construction solide.

EE 21 765 *pour plus amples renseignements*

Redresseurs haute tension

(*Illustration à la page 317*)

Cette série de redresseurs à cartouches au silicium couvre la gamme de 600 V à 10 kV de tension inversée de crête. Les redresseurs sont sous enveloppe de céramique et peuvent être utilisés à des altitudes allant jusqu'à 27432. Ils sont prévus pour des sorties de courant continu de 75 à 250 mA (à 25°C) et leur gamme de températures de service va de −55°C à +150°C. Le courant inverse maximum à une tension inversée de crête spécifiée et à 25°C il est de 10 µA.

EE 21 766 *pour plus amples renseignements*

KELVIN & HUGHES (Industrial) LTD

Kelvin House, Wembley Park Drive, Wembley, Middlesex

Système de dépouillement de données

On verra pour la première fois à ce Salon un système de dépouillement de données conçu en vue d'assurer la plus grande souplesse d'application et permettre l'analyse et l'enregistrement d'un nombre illimité de variables de traitement, à un degré de précision de 0,1% de la lecture et à des vitesses dépassant toutes celles obtenues auparavant avec ce genre d'ordinateur.

L'entrée étalonnée est un courant direct ou une tension directe, de sorte que le potentiel d'applications de l'équipement n'est limité que par la disponibilité de transducteurs adéquats. Ces transducteurs s'obtiennent facilement pour des applications se rapportant à la pression, au débit, à l'effort, aux sorties par impulsions ou à courant alternatif, tandis que les thermocouples servent à des applications de température. Des éléments compensateurs de jonction à froid sont prévus pour chaque type d'entrée de thermocouple.

L'information est entièrement présentée sous forme numérique. Des dispositifs de contrôle de hautes et basses limites, avec indicateurs-avertisseurs, ainsi que pour le changement de gamme, de suppression de zéro et de linéarisation incorporée, tels que requis pour les thermocouples, le débit, etc. peuvent être intégrés dans la séquence d'analyse. L'enregistrement se fait par machine à écrire électrique sur feuilles vierges ou préimprimées, les points non normaux étant imprimés en rouge.

Un voltmètre numérique est utilisé pour mesurer les variations de traitement et présenter le résultat sous forme numérique. Il comporte un élément de cadencement, des relais un amplificateur à deux étages et des blocs d'alimentation. Ces derniers comprennent une alimentation stabilisée, à 0,05% près, pour les besoins de références de mesures. Les tensions sont commutées en séquence et comparées avec la tension d'entrée jusqu'à ce qu'une valeur hors d'équilibre finale, à 0,1% près, soit atteinte.

Le système, y compris les transducteurs et les blocs d'alimentation, est logé dans un panneau du type cabine blindée, les divers éléments étant montés sur crémaillère afin d'en faciliter la sortie et la réparation.

EE 21 767 *pour plus amples renseignements*

GEORGE KENT LTD

Luton, Bedfordshire

SUPPRESSEUR DE QUADRATURE

(*Illustration à la page 318*)

Le suppresseur de quadrature type 1 a été étudié pour être utilisé avec la gamme d'instruments électroniques George Kent, lorsqu'il est appliqué à la mesure de signaux de courant alternatif.

Il consiste en un petit assemblage électronique qui élimine la possibilité de saturation d'amplification et d'erreurs de mesure résultant de la présence d'une composante en quadrature dans le signal d'entrée. Un degré élevé de sécurité de fonctionnement et de robustesse a été réalisé par l'emploi exclusif de transistors, de thermistors et d'autres pièces solides. La partie principale du Suppresseur est un potentiomètre à thermistors breveté Kent, qui comprend un circuit imprimé à prises, portant les pièces essentielles. Ces dernières sont collectivement comparables à l'amplificateur, au moteur d'équilibrage et au fil à contact glissant d'un servo-système classique.

Le nouvel appareil supprime efficacement les effets de quadrature qui se produisent fréquemment dans les servo-systèmes à courant alternatif, tels que ceux pour mesurer la conductivité électrolytique ou le rythme de débit de liquides. Ces faux signaux peuvent avoir des effets fort nuisibles car toute composante en quadrature se trouvant dans l'entrée d'un servo-système à courant alternatif ne peut être réduite à zéro par la réaction alternative principale, et elle donne lieu à un signal erroné irréductible. Ce dernier peut, à son tour, provoquer une erreur dans la sortie du système ou causer une saturation partielle ou même totale de l'amplificateur, résultant en une perte de sensibilité.

Le suppresseur de quadrature apporte une solution au problème en appliquant une action d'annulation automatique à n'importe quelles composantes en quadrature captées, quelles soient de grandeur stable ou pas, tout en n'influant pas le moins du monde sur le signal d'entrée principal. Le signal d'erreur de quadrature résiduelle, provenant du servo-amplificateur, est amplifié et démodulé par un circuit contenant quatre transistors au germanium de faible puissance, utilisés pour commander différentiellement la puissance fournie aux réchauffeurs de deux thermistors chauffés indirectement. Les deux chenilles de thermistors forment un diviseur de tension de gain variable qui commande un signal de phase de quadrature précise (obtenu par un décalage de 90° de la référence de courant alternatif) qui est ramené à l'entrée du servo-amplificateur et supprime le signal résiduel. Le circuit est à couplage direct de bout en bout, et un troisième thermistor fournit une réaction globale de courant continu négatif de la sortie à l'entrée. Ceci a pour effet de stabiliser les points de fonctionnement de tous les quatre transistors sans usage quelconque de condensateurs, et de maintenir les caractéristiques de sortie du potentiomètre à thermistor sensiblement constantes dans des températures ambiantes entre 0° et 60°C.

EE 21 768 *pour plus amples renseignements*

APPAREIL DE RÉFÉRENCE DE TENSION

(*Illustration à la page 318*)

L'appareil de référence type 3 est une source de référence solide de courant continu, conçue et produite par George Kent Ltd., pour être employée avec la gamme d'enregistreurs auto-équilibreurs, d'indicateurs et de contrôleurs électroniques de cette société. C'est un assemblage à prises compact, fonctionnant directement à partir du secteur alternatif normal, et donnant une sortie extrêmement stable et exempte d'ondulation de 5,0 mA à 5V pour le curseur du circuit de mesure d'instrument.

Deux étages en cascade à diodes Zéner réduisent les effets des variations normales d'alimentation par un facteur d'au moins 200, tandis que la compensation de température individuelle de chacun de ces étages maintient la sortie substantiellement indépendante dans des températures ambiantes de 10° à 70°C. Un avantage inhérent à l'appareil est qu'il continue à fonctionner avec une précision légèrement inférieure si la tension d'alimentation est fortement réduite, et ce même jusqu'à 30% de sa valeur nominale.

La sortie de courant continu est isolée de la masse et elle est, en outre, efficacement isolée du secteur par un transformateur d'alimentation équilibrée à blindage spécial. Cette particularité est de grande valeur dans certains circuits de mesure ayant une impédance limitée vers la masse.

EE 21 769 *pour plus amples renseignements*

KYNMORE ENGINEERING CO. LTD

19 Buckingham Street, London W.C.2

RELAIS À HAUTE SENSIBILITÉ

(*Illustration à la page 318*)

Ces relais, qui sont fabriqués par la société Siemens & Halske, peuvent être utilisés lorsqu'il est nécessaire de contrôler des signaux dans la zone pico-watt ou lorsque ces signaux doivent être en mesure d'amorcer ou d'arrêter un processus commandé, ainsi que pour une commande à contact positif dans des limites rigoureusement préréglées, pouvant être répétée de manière sûre.

Le principe de fonctionnement est le suivant:

Lorsque la valeur préréglée a été atteinte, le contact est actionné positivement, moyennant une dissipation de puissance des plus réduites. Ceci résulte de ce que le fonctionnement de l'induit à cadre mobile suspendu sur bande à aimant-noyau est allié à celui d'un amplificateur à pression de contact magnétique. Même lorsque des signaux à pente extrêmement douce forment l'entrée, le contact se ferme par séparation instantée. L'aiguille de l'induit suspendu sur bande contient un petit morceau de fil de fer. Dans la largeur de bande de service préréglée, un petit aimant permanent sert d'arrêt à la limite réglable de déviation. Lorsque le signal d'entrée augmente, l'aiguille se déplace vers l'amplificateur à pression de contact magnétique jusqu'à ce que le couple à suspension par bande soit équilibré par le flux de l'aimant. A ce moment là, la moindre augmentation de signal force l'aiguille à sauter vivement vers l'arrêt magnétique où une pression de contact d'environ 2g sera maintenue. Vu que les surfaces de contact de l'aiguille et de l'arrêt magnétique sont doublées d'or, le contact pourra être rétabli en toute sûreté, même après une longue période d'emmagasinement ou d'inutilisation. Le contact-aiguille excite un relais asservi incorporé dont le bobinage est alimenté par une source de 24 V C.C. Les contacts du relais asservi sont prévus pour le service dans les contacteurs ou avertisseurs industriels, ce qui permet de commander divers circuits indépendants. Le relais asservi peut être fourni avec un nombre de différentes combinaisons de contact. Un de ces contacts sert de contact de "fixage", en parallèle avec le contact à aiguille-relais. Ceci a pour effet de maintenir le contact jusqu'a ce qu'un commutateur à bouton-poussoir manuel soit actionné pour réenclencher le relais. Le bouton de réenclenchement manuel remplit une fonction à double effet en ce qu'il excite d'abord un bobinage chutant dont l'induit détache le contact-aiguille, qui est "mort" puisqu'il est parallèle au contact de fixage du relais asservi. Ce n'est qu'ensuite que le circuit de bobinage du relais asservi sera interrompu. Ceci empêche tout amorçage d'arc au contact d'aiguille. Si le signal original persiste lorsqu'on actionne le réenclenchement, le relais excitera à nouveau immédiatement le relais asservi.

Ces relais sont munis d'une commande variable externe pour régler le point de fonctionnement et une échelle divisée linéairement. Ils sont fournis pour diverses gammes de courant et de tension, la sensibilité maxima pouvant être obtenue est de 10^{10} W.

EE 21 770 *pour plus amples renseignements*

J. LANGHAM THOMPSON LTD

176 High Road, Bushey Heath, Hertfordshire

TRANSDUCTEUR DE PRESSION

Conçu pour la mesure de pressions stables et à fluctuation rapide, la réponse de fréquence de ce transducteur le rend particulièrement propre à l'étude des phénomènes transitoires dans les systèmes hydrauliques et pneumatiques.

Les modèles offerts sont prévus pour les gammes de 0 à 141kg/cm² et de 0 à 1 400kg/cm² de pression de jauge avec des pressions de surcharge élevées.

Le palpeur de pression consiste en un cylindre de cuivre au glucinium, partiellement alésé afin de former un tube à extrémité solide, ayant deux enroulements d'extensomètre résistifs-actifs au dessus de la section alésée, ainsi que deux enroulements compensateurs de température au dessus de la section solide. Ces derniers sont reliés sous forme de réseau en pont équilibré à quatre bras.

La pression s'exerçant à l'intérieur du cylindre cause une dilatation de la section alésée se répercutant en changements résistifs des enroulements actifs, de sorte qu'un signal proportionnel à la pression résulte de l'application d'une tension alternative ou continue au pont. La sortie de la gamme totale est de 1,5 mV/V appliqués, la tension de fonctionnement maximum étant de 30 V. La précision globale est supérieure à 0,5%, y compris la linéarité et l'hystérésis.

EE 21 771 *pour plus amples renseignements*

LAURENCE, SCOTT & ELECTROMOTORS LTD

Mansfield House, 376 Strand, London W.C.2

APPAREIL DE VITESSE CONSTANTE

Cet appareil est capable de maintenir une rotation d'arbre à une vitesse constante, à un degré supérieur à une partie dans trois mille, sur toute la gamme nominale du moteur. Il existe deux modèles, l'un pour l'emploi avec un petit moteur à induction, en tant que générateur de force motrice, et l'autre avec un moteur à aimant permanent. L'appareil fournit un signal de référence de vitesse, obtenu d'une bobine exploratrice dont les pôles sont montés près de la périphérie d'une roue dentée, couplée directement à l'arbre du moteur d'entraînement. A la vitesse constante déterminée, le signal sélectif est de 50 kHz. Ce signal est amplifié et transmis à un circuit à filtre piézoélectrique, résonant à un peu plus de 50 kHz. La sortie de ce filtre est redressée et elle commande finalement un étage de sortie à charge cathodique. Dans le cas de l'entraînement par courant continu, le moteur est relié directement dans le circuit cathodique, mais dans le cas de la version par courant alternatif, la charge cathodique est constituée par l'enroulement de commande d'un amplificateur magnétique qui commande à son tour la vitesse du moteur.

EE 21 772 *pour plus amples renseignements*

MARCONI INSTRUMENTS LTD

St. Albans, Hertfordshire

COMPTEUR 10 MHz

(Illustration à la page 319)

Le compteur électronique 10 MHz, type TF 1345, est un instrument de comptage/minutage à action rapide, avec étalon de fréquence incorporé de grande précision. La lecture se fait par indicateurs au néon sur écran numérique à 8 décades. L'instrument peut compter jusqu'à 10 millions par seconde, mesurer la fréquence de 10 Hz à 10 MHz, et des périodes de formes d'ondes allant jusqu'à 100 kHz. Un choix d'accessoires à prises étend la gamme de fréquences jusqu'à 220 MHz, permet la mesure de temps jusqu'à 1 μsec et augmente la sensibilité jusqu'à 10 mV. Le contrôle de la durée d'affichage est soit manuel, soit à variation continue de 0,1 à 10 sec. avec réenclanchement automatique et à répétition. L'appareil est livrable pour montage sur banc d'essai ou sur bâti.

EE 21 773 *pour plus amples renseignements*

OSCILLOSCOPE À DOUBLE AFFICHAGE

(Illustration à la page 319)

Le TF 1331 est un oscilloscope à double trace, à couplage de courant continu, avec temps de lecture directe et étalonnage de tension. Grâce au commutateur électronique de faisceau, deux signaux indépendants peuvent être affichés soit par balayage alternatif soit à un taux de commutation de 100 kHz. L'une ou l'autre entrée peut aussi être affichée séparément, sans commutation. La réponse de fréquence s'étend du courant continu à 15 MHz, et la sensibilité est variable en cinq pas de 50 mV/cm à 50 V/cm. La vitesse de balayage de la base de temps varie de 1 sec/cm à 0,1 μsec/cm. par étalement X. Le déclenchement peut s'effectuer intérieurement ou extérieurement, ou bien encore, à partir d'une source de fréquence d'alimentation intérieure. Une sonde d'entrée de 10 MΩ est fournie comme accessoire facultatif. Il existe également une version à faisceau unique: le type TF 1330.

EE 21 774 *pour plus amples renseignements*

METRIX INSTRUMENTS LTD

54 Victoria Road, Surbiton, Surrey

CONTRÔLEUR DE PROCESSUS

(Illustration à la page 320)

Le contrôleur transistorisé Metrawatt constitue une méthode précise quoique peu coûteuse de contrôle de température, de pH, de conductivité, de puissance et d'autres variables dont la fonction peut être traduite sous forme de tension électrique.

Le système de contrôle emploie un ou plusieurs circuits oscillants haute fréquence, dont les bobines de couplage sont montées sur guides ajustables, pouvant être réglés à n'importe quelles valeurs de contrôle voulues sur l'échelle de l'instrument. Un petit volet sur le bras de l'appareil de mesure à cadre mobile pénètre dans l'entrefer des bobines oscillatrices lorsque l'aiguille atteint les valeurs des positions de contrôle préréglées. Les circuits oscillants actionnent à leur tour des contacteurs au mercure ou de grande puissance—au choix—afin d'effectuer la commande finale du système. Etant donné que l'appareil de mesure est sans contacts mécaniques, le mouvement peut donc être d'une sensibilité extrême, donnant une commande de grande précision (0,2%) avec une réponse rapide.

Un système simple à réaction thermique, fourni sur demande, transforme l'instrument en un régulateur proportionnel. L'effet d'inertie thermique dans le système de contrôle est alors éliminé presque complètement.

Cet appareil, qui est entièrement transistorisé, englobe un circuit de protection qui veille à ce que l'alimentation de commande soit débranchée en cas de panne dans le thermocouple ou le transducteur ou, encore, dans l'appareil lui-même.

L'effet des fluctuations de tension secteur n'est que de 0,1% pour une variation de secteur de 15%.

EE 21 775 *pour plus amples renseignements*

MUIRHEAD & CO. LTD

Beckenham, Kent

VOLTMÈTRE DE DÉCADES EFF. DE PRÉCISION

(Illustration à la page 320)

Le voltmètre Muirhead-Wigan D930A est un instrument très précis à gammes étendues de tensions (1mV à 300V) et de fréquences (5Hz à 100 kHz). Etant un véritable voltmètre de valeurs efficaces, il peut mesurer des tensions à formes d'ondes sinusoïdales et non-sinusoïdales.

Les lectures de tension s'obtiennent directement par quatre commandes marquées "unités" (1-10), "dizaines" (0-10), "centaines" (0-10) et "milliers" (0-10), et par un commutateur de ×0,001 à ×100.

Une précision de lecture du cadran allant jusqu'à 0,1% peut donc toujours être atteinte et des lectures peuvent être interpolées jusqu'à 0,025%.

La résistance d'entrée constante de 100kΩ permet une lecture de basse tension avec une faible consommation de courant. Même à 300V, la dissipation de puissance d'entrée est inférieure à 1W.

Les caractéristiques expérimentales de cet instrument sont comme suit:

Précision de la mesure de tension:

30 Hz-5 kHz	60 mV - 10 V:	± 0,05% (sauf à la fréquence d'alimentation)
	1 mV-300 V:	± 0.1% ou 30 μV (selon la valeur la plus élevée)
20 Hz-20 kHz	1 mV-300 V:	± 0,2% ou 30 μV (selon la valeur la plus élevée)
5 Hz-100 kHz	10 mV-300 V:	± 0,25% jusqu'à 50 kHz décroissant à ±1% à 100 kHz (à l'aide d'un oscillateur extérieur stable tel que le D-890-A/2 pour la normalisation à la fréquence d'essai)

Précision de la mesure de rapport de tension dans n'importe quelle gamme:

Gammes: 100 V-300 V 10 V-100 V etc. jusqu'à 10 mV-100 mV	De ± 0,03% à 0,15% (selon le rapport et la tension) jusqu'à 50 kHz (non inférieur à ±0,2% à la fréquence d'alimentation)
Gamme 1mV-10mV	De ± 0,05% à 0,15% (selon le rapport) jusqu'à 10 kHz (non inférieur à ±1% à la fréquence d'alimentation)
Impédance d'entrée	100 ohms shuntés par 200 pF (max.)
Rapport maximum crête/ valeur efficace de la forme d'onde d'entrée:	3:1

EE 21 776 *pour plus amples renseignements*

SHORT BROTHERS & HARLAND LTD

Seaplane Works, Queens Island, Belfast, Irlande

CALCULATEUR ANALOGIQUE

(Illustration à la page 320)

Ce nouveau calculateur universel, dénommé "Simlac Minor", est d'un intérêt particulier pour les utilisateurs virtuels ayant besoin d'une plus grande capacité que celle du calculateur universel à 18 amplificateurs, mais qui n'auraient pas besoin, toutefois, de la capacité ou du degré de précision du grand calculateur "Simlac", récemment réalisé.

Le "Simlac Minor" a une capacité de 32 amplificateurs et 100 éléments passifs. A l'instar du grand modèle, il est monté sur crémaillère standard, l'appareil entier étant logé dans deux râteliers de 48,2 cm du type employé par les services postaux. Le râtelier de gauche contient les éléments de calcul et celui de droite loge l'instrumentation, les panneaux amovibles, les blocs d'alimentation et trois postes non-linéaires spéciaux.

Afin d'assurer une utilisation maxima de l'espace disponible, la stabilisation dite "de tour" n'est pas employée. Néanmoins, grâce à la construction et au choix judicieux des pièces, la précision globale des éléments de calcul est de 0,1%. Les amplificateurs et les éléments passifs sont des accessoires courants Simlac, en dehors du dispositif de fixage, qui est un supplément facultatif.

EE 21 777 *pour plus amples renseignements*

OSCILLATEUR B.F. ET ANALYSEUR DE SYSTÈME DE CONTRÔLE

Cet oscillateur basse fréquence a été étudié en vue d'essais de basses fréquences, pour lesquels une teneur en courant continu extrêmement basse et des amplitudes maintenues avec précision sont de la plus haute importance. Toute dérive des amplificateurs est continûment corrigée afin que la teneur en courant continu de la sortie soit normalement bien au dessous de 100 μV.

L'analyseur de système de contrôle est sans égal quant à sa faculté de fonctionner à des fréquences exceptionnellement basses, pouvant même, en effet, fonctionner avec succès à une fréquence de 1 cycle par 100 secondes. Sa gamme s'étend jusqu'à 1 kHz et il est capable d'analyser une multiple variété de systèmes.

Il incorpore un oscillateur B.F. et présente des lectures stables de composantes en quadrature et en phase, même à la fréquence la plus basse, soit 0,01 Hz. Le niveau de courant continu est toujours inférieur à 0,005% et, en équilibrant attentivement les amplificateurs à glissement corrigé, on peut maintenir ce niveau à 0,0005%.

EE 21 778 *pour plus amples renseignements*

Résumés des Principaux Articles

Un Analyseur de spectres pour vidéo-fréquences par A. B. Whitwell et N. Williams

Résumé de l'article aux pages 263 à 274

Un analyseur de spectres universel pour la gamme de fréquences de 500 kHz est décrit dans cet article. Les principes de réalisation des appareils de ce genre ne sont pas nouveaux, mais l'intérêt manifesté pour l'utilisation des analyseurs de spectre comme auxiliaires universels de laboratoire et de contrôle de production s'accroît rapidement. On peut, en effet, les employer pour une série étendue de mesures et les étudier pour une haute sensibilité, en raison de la réduction du coefficient de bruit effectif de l'étage l'entrée, par suite de l'action de filtrage ultérieure. Ainsi, l'appareil dont il est question ici a une sensibilité de seuil d'environ 0,2 μ/V dans une bande de 70 Hz et sa gamme dynamique est de 50 dB.

Analyse graphique et mode de construction de l'oscillateur d'arrêt par D. Chambers

Résumé de l'article aux pages 275 à 279

L'action de l'oscillateur d'arrêt est décrite dans cet article qui, de même, examine en détail chaque pièce du circuit. Les formes d'ondes de courant et de tension sont construites selon une méthode progressive, et le tube, le condensateur d'arrêt, le noyau de transformateur et le rapport de tours sont déterminés pour une application particulière.

Mesure directe pour un calculateur analogique à transformateur par J. F. Young

Résumé de l'article aux pages 280 à 288

Les tensions de calculateur sont déphasées et additionnées par des amplificateurs et les résultantes sont fournies à des amplificateurs de gain réglables. Ces derniers alimentent des amplificateurs de puissance à réaction d'intensité qui commandent des ampèremètres et des wattmètres. Des réseaux de déphasage à résistance constante sont utilisés avec un wattmètre pour mesurer les volts-ampères réactifs. Un déphaseur étalonné est utilisé pour la mesure de phase.

Un réseau à résistance constante ayant les propriétés d'un shunt universel dans lequel le rapport de shunt est commandé par échelles de décades par E. R. Wigan

Résumé de l'article aux pages 289 à 293

Par une simple modification, le réseau en T à résistance constante peut être transformé en shunt universel à résistance constante, réglable sur une base décimale. L'auteur décrit un shunt de 1000 ohms à trois décades, ayant un rapport de shunt maximum de 11,111/1,0 et réglable par plots de 0,001. Entre ces derniers, une interpolation continue peut être effectuée, dont la précision est analysée par l'auteur.

Un générateur de successions d'impulsions à éclairs stroboscopiques pour l'étude de l'écoulement turbulent des fluides par R. E. King

Résumé de l'article aux pages 294 à 295

Un tube sélectif à cathodes froides est employé pour obtenir un train d'impulsions d'une série déterminée dans une gamme étendue de fréquences de répétition d'impulsions. La sortie du générateur déclenche un éclair stroboscopique qui fournit une indication photographique des directions et vitesses locales de l'écoulement turbulent dans un fluide.

Un oxymètre enregistreur perfectionné par A. W. Melville, D. H. Smith et J. B. Cornwall

Résumé de l'article aux pages 296 à 300

Les auteurs analysent la théorie de l'oxymétrie autitive et décrivent un appareil qui résout continûment l'équation théorique et fournit un enregistrement de la saturation absolue d'oxygène artériel.

Les methodes employées jusqu'ici pour obtenir un signal proportionnel au logarithme de l'intensité de la lumière tombant sur les cellules photoélectriques de l'appareil auriculaire ont été trouvées peu sûres et ont dû être remplacées.

Le nouveau circuit présente d'autres avantages, assurant un degré élevé de stabilité et de répétabilité.

L'instrument est destiné en particulier à l'usage des anesthésistes, durant les opérations thoraciques prolongées, mais il peut être employé en n'importe quel cas ou l'enregistrement continu de la saturation d'oxygène s'avère nécessaire.

Un stimulateur physiologique à transistors à jonction par W. T. Catton et L. Molyneux

Résumé de l'article aux pages 301 à 302

Après une expérience de deux ans dans les classes de physiologie avec un stimulateur d'un premier modèle utilisant un seul transistor à pointe, il fut décidé, en raison du caractère suranné de ce dispositif, de réaliser un nouveau circuit à transistors à jonction. Le nouveau circuit, décrit dans cet article, ne comporte pas de modification des commandes du panneau avant, donne une meilleure forme d'onde et une impédance de sortie réduite et offre, en outre, l'avantage d'une tension de pile d'alimentation plus faible. Le nouveau stimulateur assure les mêmes services que le modèle précédent, c'est à dire; tension de sortie variable de 0 à 25V, fréquence variable de 2 à 100/sec. et largeur d'impulsion constante de 1 msec environ.

Un modulateur par impulsions de largeur variable par D. C. Brown et J. E. Baughen

Résumé de l'article aux pages 302 à 303

Un modulateur par impulsions de largeur variable, utilisant l'effet de stockage à ouverture dans un transistor à jonction, vient d'être réalisé. Par l'emploi d'un signal échantillonné à 10 kHz, on a pu atteindre une sensibilité de 20 μsec/V.

Un générateur d'impulsions à taux variable modulées en tension par E. J. C. Fowell et A. Cowley

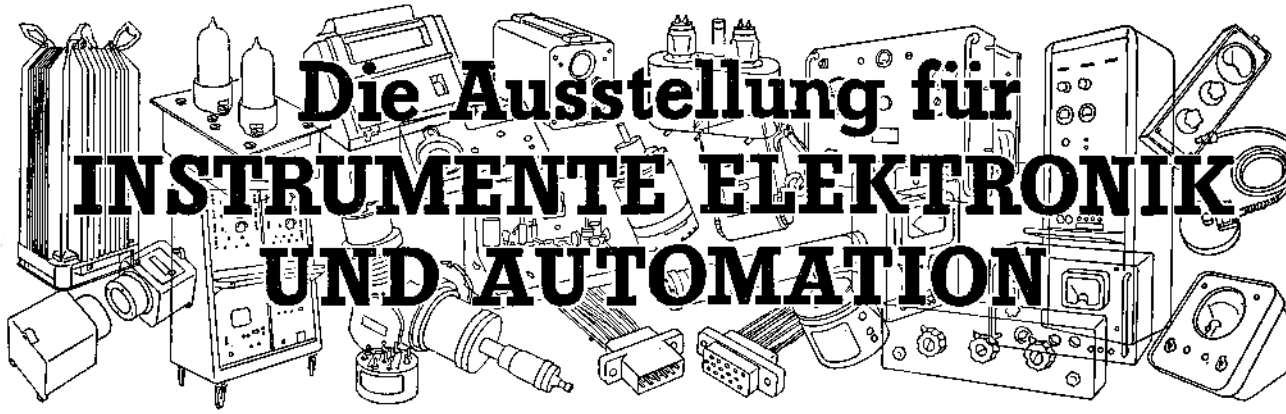
Résumé de l'article aux pages 304 à 306

Cet article décrit un circuit de générateur d'impulsions, dont la fréquence de sortie est une fonction linéaire de la tension d'entrée.

Le circuit utilise un système de circulation d'impulsions à double transitron intégrateur à couplage cathodique, avec les modifications appropriées pour améliorer la précision de transfert aux basses fréquences de sortie.

Quatre gammes de fréquences de sortie sont prévues entre 5 et 2500 impulsions par seconde.

Un schéma de circuit, ainsi que des résultats de performances, sont également donnés.



Die Ausstellung für INSTRUMENTE ELEKTRONIK UND AUTOMATION

Beschreibung neuer Bauelemente, Zubehörteile und Prüfgeräte auf Grund der von Herstellern gemachten Angaben.

Übersetzung der Seiten 315 bis 320

A.P.T. ELECTRONICS INDUSTRIES LTD

50-52 King Street, Southall, Middlesex

RÖHREN-PRÜFGERÄT

(*Abbildung Seite 315*)

Das Prüfgerät umfasst stabilisierte Netzgeräte, Mehrbereich-Instrumentierung zur Untersuchung aller Röhren-Parameter und eine Wechselstrom-Brücke für S-Messungen. Diese drei Geräteteile sind als 19 Zoll (482,6 mm) Gestelleinheiten mit 267 mm Frontplatten konstruiert und werden in einem Instrumentgehäuse geliefert.

Röhren können nach „CV" oder anderen Pflichtblättern im Rahmen der Netzteilkapazität und bis zu S = 100 mA/V geprüft werden. Die Arbeitskennwerte von Einzel- und Doppeldioden geringer Leistung, Einweg- und Doppelweggleichrichtern, Hochspannungs-Trockengleichrichtern und gasgefüllten Stabilisatoren können auch gemessen werden.

Wenn das Prüfgerät nicht in Betrieb ist, können die Netzgeräte mit ihren Messgeräten für andere Zwecke benutzt werden. Aussenliegende Messgeräte können über Klinken zum Eichen eingeschaltet werden, wenn äusserste Genauigkeit erforderlich ist.

Folgende Netzgeräte sind vorgesehen:

Anode

0...300 V in drei Bereichen, konstant innerhalb 1%

0...250 mA (Messbereiche 1; 10; 30; 100; 300 mA Vollausschlag)

Schirmgitter

0...300 V in drei Bereichen, konstant innerhalb 1%

0...50 mA (Messbereiche 1; 10; 30; 100 mA Vollausschlag)

Gitter

0...100 V und 0...+25 V, konstant innerhalb 1% (Messanzeige 100...0...100 μA Vollausschlag)

Heizer

0...55 V Wechselstrom in vier Bereichen für je 25 kW max (Messbereiche 10; 30; 100 V Vollausschlag und 0,5 A Vollausschlag)

EE 21 751 *für weitere Einzelheiten*

ADVANCE COMPONENTS LTD

Roebuck Road, Hainault, Ilford, Essex

TRANSISTORISIERTE NETZGERÄTE

(*Abbildung Seite 315*)

Das kleine, kompakte, transistorisierte Netzgerät Type PP5 liefert einen stabilisierten Gleichstrom von max 0,5 A. Die Ausgangsspannung ist über einen direkt geeichten Bereich von 0...15 V kontinuierlich variabel.

Elektronische Schutzvorrichtungen gegen Stromüberlastung sichern auch gegen Kurzschlüsse. Wiedereinstellung erfolgt auf herkömmliche Weise von der Frontplatte. Der Arbeitspunkt des Überstromschutzes ist ausserdem einstellbar, um die in Gebrauch befindlichen Verbraucherkreise gegen thermisches Durchgehen der Verstärker usw. zu sichern. Der Schutz erstreckt sich auf übertriebene Umgebungsbedingungen bei Höchstbelastung.

Das Messgerät kann durch Umschaltung zur Anzeige von Ausgangsstrom als auch -spannung auf einer Doppelskala benutzt werden.

Das Gerät arbeitet an Wechselstrom-Netzspannungen von 200...250 V. Die Ausgangsimpedanz liegt unter 0,01Ω bei Gleichstrom und unter 0,2Ω bis zu 100 kHz.

EE 21 752 *für weitere Einzelheiten*

AMPHENOL (Great Britain) LTD

Victoria Road, Burgess Hill, Sussex

H.F. KUPPLUNGEN

(*Abbildung Seite 315*)

Die amerikanischen HF-Kupplungen Muster C und N werden jetzt auch in englischer Ausführung mit verbesserter Kabelhaltung entsprechend den britischen militärischen Anforderungen hergestellt.

Sie sind mit auswechselbaren Buchsen für Uni-Radio-Kabel 102, 73, 67 oder 81 lieferbar. Alle Kontakte werden „festgehalten". Sowohl gerade als rechtwinklige Kupplungen für freie Verbindungen oder Frontplattenmontage werden hergestellt und entsprechen mit Ausnahme der Kabelhaltung den C-Mustern UG-573/U, UG-710A/U, UG-570/U, UG-572/U und UG-571/U und den N-Mustern UG-21B/U, UG-22B/U und UG-23B/U. Alle haben eine Nennimpedanz von 50Ω. Die Typennummern für den britischen Militärbedarf für diese Bauelemente werden in Kürze bekannt gegeben.

Zum Zusammenbau dieser Kupplungen ist ein besonderer Schlüssel erforderlich.

EE 21 753 *für weitere Einzelheiten*

ASSOCIATED ELECTRICAL INDUSTRIES

Rugby, Warwickshire

ANALOGRECHNER

(*Abbildung Seite 316*)

Der Analogrechner AEI 955 löst lineare und nichtlineare Differentialgleichungen und kann zur Untersuchung der dynamischen Leistung physikalischer Systeme benutzt werden.

Der Rechner kann für lineare (Addieren und Integrieren usw.) und nichtlineare Funktionen (Tote Zone und/oder Grenzen, willkürliche Funktionen, Multiplizieren usw.) mit Rechnerbausteinen genormter Grösse eingestellt werden. Dadurch können Grösse und Anordnung des Rechners dem zu untersuchenden Problem angepasst werden. Einstellung erfolgt durch Verbindungsschnüre auf der Vorderseite des Rechners.

Die Genauigkeit der Rechnerimpedanzen kann wahlweise 1% oder 0,1% sein. Der Rechenverstärker hat kontinuierliche Driftberichtigung, hohe Verstärkung, grosse Bandbreite, geringe Drift, geringen Gitterstrom und ist geräuscharm.

Kontinuierliche oder wiederholende Betriebsweise wird durch Steuerelemente eingestellt, die mit einem Doppelspur-Oszillografen hoher Genauigkeit (±0,5%) zusammengebaut sind. Eine automatische Schaltung ermöglicht es, den Rechner nach einer vorgewählten Zeit oder vorherbestimmten Problemspannung „festzuhalten". Die Ergebnisse können durch ein 2-Kanal-Schreibgerät der geradlinigen Ausschreibtype aufgezeichnet werden, das dafür wärmeempfindliches Papier benutzt.

EE 21 754 *für weitere Einzelheiten*

B & R RELAYS LTD

Temple Fields, Harlow, Essex

TRANSISTORISIERTES STEUERRELAIS

(Abbildung Seite 316)

Dieses transistorisierte Relais wurde für Starkstromschalten entwickelt, für das nur kleine Betriebssignale zur Verfügung stehen. Es ist ein dreistufiges Silizium-Transistor-Gerät, das 30 mA bei 12 V Gleichstrom verbraucht. Das Relais spricht auf einen Steuerstrom von 15 μA im Steuerkreis an und widersteht hundertfacher Überlastung. Höchstzulässige Umgebungstemperatur ist 60°C.

Die Gesamtabmessungen dieses transistorisierten Relais mit gedruckter Schaltung und in Einsteckausführung sind 50,8 × 38,1 × 31,75 mm. Alle Kontakte und Sockelstifte sind vergoldet. Die gedruckte Schaltung passt in einen 18 fachen, unverwechselbaren Sockel.

EE 21 755 *für weitere Einzelheiten*

BELLING & LEE LTD

Great Cambridge Road, Enfield, Middlesex

BEFESTIGUNGS- UND MONTAGETECHNIK

Das „Prestincerts" Verfahren ist eine vollständig neue Montagemethode für Bauelemente aller Art auf Tafeln, Chassis aus allen Materialien, Spritzgusskörper usw. oder zum Zusammennieten von Lamellen ohne vorhergehendes Stanzen oder Bohren von Löchern.

Dieses neue Verfahren ergibt äusserst gute mechanische und, wo erforderlich, elektrische Verbindungen. Es kann daher mit gleichem Erfolg überall dort Verwendung finden, wo im elektrischen, elektronischen und allen technischen Sektoren normalerweise genietet wird. In der Elektronik ist die Methode z.B. für Aufbau und Verdrahtung von Bauelementen in gedruckten Schaltungen mittels eines einfachen Arbeitsvorganges geeignet, wobei Löten freisteht. Die Arbeit kann je nach Bedarf entweder reihenmässig, einzeln, oder gleichzeitig an mehreren Punkten ausgeführt werden.

Bei Massenproduktion können die „Prestincerts" und Bauelemente durch Kastenspeiser zugeführt werden, wodurch Montage und Verdrahtung vollautomatisch werden. Für kleine Partien werden Pressen mit Handaufgabe und für Einzelanfertigung tragbare Handwerkzeuge benutzt; die Methode ist sehr anpassungsfähig und benötigt kein hohes Anlagekapital. Die mechanischen Anwendungsmöglichkeiten im Flugzeug-, Schiff- und Kraftfahrzeugbau sind unbegrenzt; der Produktionsleistung und mechanischen Festigkeit kommt keine andere Nietmethode gleich.

EE 21 756 *für weitere Einzelheiten*

BLACKBURN ELECTRONICS LTD

Brough, Yorkshire

HALL-EFFEKT-MULTIPLIKATOREN

(Abbildung Seite 316)

Diese Multiplikatoren (Typen BIE 293 und BIE 294) wurden für Analogrechner und Instrumentierungssysteme entwickelt.

Type BIE 293 erfüllt die Gleichung $(XY/10^3) = Z$ und Type BIE 294 $XY \cdot 10^3 = Z$.

Die Bausteine benutzen den Hall-Effekt in Halbleitern und sind transistorisiert.

Die Multiplikatoren haben eine Genauigkeit von ±0,5% des Gesamtbereiches und einen Frequenzgang von Gleichstrom bis 100 Hz.

EE 21 757 *für weitere Einzelheiten*

TRANSISTORISIERTE BINÄRE UHR

Dieses Gerät gibt 1-Sekunden-Taktimpulse mit 0,05% Genauigkeit. Innerhalb eines Zeitabschnittes von 9 999 s ist jeder 1s-Impuls in einzigartiger Weise durch eine 16Bit-Binärcodegruppe gekennzeichnet. Die Impulse können auf Magnettonbändern oder anderen Systemen aufgezeichnet werden.

Wiedergabe der aufgezeichneten Zeitsignale durch den nicht ausgestellten Entschlüssler BIE 417 ermöglicht Identifizierung jedes 1s-Intervalls unabhängig von Geschwindigkeit oder Laufrichtung des Tonbandes.

Infolge kleiner Abmessungen geringen Gewichtes ist diese Uhr besonders zusammen mit Registriergeräten für Bordbetrieb in Flugzeugen geeignet.

EE 21 758 *für weitere Einzelheiten*

"DIAMOND H" SWITCHES LTD

Gunnersbury Avenue, London W.4

RELAIS

(Abbildung Seite 316)

Luftdicht verschlossene Miniaturrelais der Serie C sind für Verwendung in Flugzeugen gedacht und haben einen Betriebstemperaturbereich von −65° . . . +200°C. Sie wiegen bis zu 113 Gramm, haben Vibrationswiderstand bis zu 2 kHz bei 20 g und Betriebsstosswiderstand bis zu 50 g.

Die abgebildete BW-Serie besteht aus preisgünstigen Mehrzweckrelais mit Abmessungen 38,1 × 38,1 × 47,6 mm. Die Wechselstrom-Nennschaltleistung ist 25 A (induktionsfrei) bei 250 V und 1 PS bei 115 V oder 2 PS bei 250 V. Diese Relais können für Seiten- oder Frontplattenmontage sowie mit Einstecksockel und verschiedenen Gleich- oder Wechselstromspulen geliefert werden. Die Standard-Kontaktausrüstung besteht aus einem zweipoligen Umschalter, jedoch können andere Anordnungen und auch Folgekontakte vorgesehen werden.

EE 21 759 *für weitere Einzelheiten*

EKCO ELECTRONICS LTD

Southend-on-Sea, Essex

FEUCHTIGKEITSMESSER

Der neue Ekco Feuchtigkeitsmesser ist ein transistorisiertes Gerät, das hauptsächlich der Bestimmung des Feuchtigkeitsgehalts von Papier dient. Üblicherweise wird am trockenen Ende einer Papiermaschine gearbeitet und die effektive Kapazität des Papiers plus Feuchtigkeit mit einer selbstabgleichenden Kapazitätsbrücke gemessen. Das Gerät berührt das Papier nicht und kompensiert Abstandsschwankungen der Messelemente selbst. Messungen können für Feuchtigkeitsgehalte von 2 . . . 15% vorgenommen und innerhalb ±¼% wiederholt werden. Für Regelungszwecke steht ein Ausgangssignal von 0 . . . 15 mA zur Verfügung, und der Feuchtigkeitsgehalt kann direkt in Prozent abgelesen werden. In Verbindung mit einem Standard-Betadickenmesser kann der Feuchtigkeitsmesser laufend einen Ausgleich vornehmen, so dass der Betadickenmesser echtes „Trockengewicht" anzeigt.

EE 21 760 *für weitere Einzelheiten*

ENGLISH ELECTRIC VALVE CO. LTD

Chelmsford, Essex

SENDERÖHREN

Das E.E.V. Co. Programm für Hochleistungs-Senderöhren wurde durch drei neue Trioden BR 1131, BR 1138 und BW 194 erweitert.

- BR 1131 ist eine druckluftgekühlte Triode, deren Kennlinien denen der BR 152B (CV 28) ähnlich sind, die aber einen thorierten Wolframheizfaden für 8,5 V/22 A benutzt und eine Anoden-Höchstverlustleistung von 3,5 kW hat.
- BR 1138 ist eine druckluftgekühlte Triode, die mit Ausnahme des thorierten Wolframheizfadens für 8,5 V/22 A mit der BR 152B (C28) identisch ist. In vorhandenen Geräten kann die BR 1138 als Ersatzröhre Verwendung finden. Der Heizspannungsunterschied wird durch Serienschaltung von zwei BR 1138 ausgeglichen.
- BW 194 ist eine wassergekühlte Ausführung der bekannten BR 194 mit einer Anodenverlustleistung von 50 kW.

TYPE	U_f (V)	I_f (A)	I_k (Spitzen) (A)	μ	S (mA/V)	U_a (V)	N_a (kW)	f (MHz)
BR 1131	8,5	22	4	40	3,1	10	3,5	15/80
BR 1138	8,5	22	2	40	3,1	10	1,0* 1,25+	15/80
BW 194	13,0	240	100	34	38	15	50	5/30

* *Kühlung ohne Druckluft* + *Druckluftkühlung*

EE 21 761 *für weitere Einzelheiten*

EVERSHED & VIGNOLES LTD

Acton Lane, London W.4

SCHREIBER-REGLER

Der neue Schreiber-Regler für drei Glieder besteht aus zwei kleinen Bausteinen und erfordert nur einen Standard-Plattenausschnitt von 114 x 114 mm. Trotz dieser kleinen Abmessungen wird durch geeignete Krümmung des auf der Platte montierten Geräteteils eine Aufzeichnungsbreite von 101,6 mm erzielt. Eine Anzeige für den „gewünschten Wert" und Einstellknöpfe für die verschiedenen Reglerfunktionen befinden sich auch in diesem Geräteteil. Transistor-Schaltungen werden benutzt, und die gesamte Stromversorgung einschliesslich einer transistorisierten, stabilisierten Stromquelle ist in den beiden Bausteinen untergebracht.

EE 21 762 *für weitere Einzelheiten*

GOODMANS INDUSTRIES LTD

Axiom Works, Wembley, Middlesex

VIBRATIONSGEBER

Der neue Vibrationsgeber VG 109 Bauserie II kann Schubleistungen bis zu 8165 kg abgeben und benutzt Ölkühlung anstelle der Luftkühlung der ersten Serie. Der Gebertisch hat einen Durchmesser von 584 mm und zusammen mit dem Schwingspulenanbau eine Masse von ungefähr 12,7 kg/A. Für 10 kW Verlustleistung stehen 3 400 . . . 4 535 kg Schub zur Verfügung und für 30 kW 5 900 . . . 8 400kg. Die ringförmige Tischoberfläche ist für Betrieb in einer Klimakammer abgedichtet.

EE 21 763 *für weitere Einzelheiten*

STEUERVERSTÄRKER

(Abbildung Seite 317)

Der Steuerverstärker E.501 wurde zur Vervollständigung des Goodman-Vibrationssystems entwickelt und dient der Steuerung der Vibrationsamplitude, Beschleunigung oder Geschwindigkeit, deren Werte vorgewählt oder von aussen programmiert werden können. Das Gerät besteht aus einer besonderen Eingangsstufe, Hauptverstärker, Messkreis, Fehlerdetektorstufe, Kompressionsverstärker und Kathodenverstärker. Unter geeigneten Umständen kann ein Fehler von 25 dB auf ungefähr 4 dB zusammengepresst werden. Die verstärkte und gleichgerichtete Ausgangsspannung eines geeigneten Vibrationswandlers wird mit einer vorherbestimmten Gleichspannung verglichen, um ein Signal zu erhalten, das die Abweichung der Vibration vom gewünschten Wert kennzeichnet. Das Fehlersignal steuert den Kompressionsverstärker, dem ein Sinuswellengeber eine Eingangswechselspannung zuführt und dadurch ein sinusförmiges Signal hervorruft, dessen Frequenz durch die Oszillatoreinstellung bestimmt wird, und dessen Amplitude (eventuell über einen Endverstärker) die richtige Vibrationshohe einregelt. Der Arbeitsbereich liegt zwischen 20 Hz und 10 kHz, aber das Gerät spricht auch ausserhalb dieser Grenzen an.

Eingangssockel sind für fünf Signalquellen vorgesehen, die im Eingangskreis des besonderen Kathodenverstärkers wahlweise angeschaltet werden können; für Quarz-Beschleunigungsmessköpfe mit einer Empfindlichkeit von 10 . . . 20 mV/g kann die Messgerätempfindlichkeit auf Vollausschlag für 10 g bzw. 50 g in beiden vorgesehenen Bereichen eingeregelt werden.

Die besondere Konstruktion des Eingangskathodenverstärkers erlaubt Verwendung von Quarz-Beschleunigungsmessköpfen mit bis zu 30 m Kabel ohne ungünstige Einwirkung. Der Ausgang der Kathodenverstärkerstufe kann über einen konzentrischen Sockel ohne zusätzliche Impedanzwandlung jedem üblichen Elektronenstrahl-Oszillografen zugeleitet werden.

Ein Oszillator-Eingangssignal von 150 mV_{eff} ist für den Verstärker geeignet, der eine maximale Ausgangsspannung von 4 V_{eff} abgeben kann. Für einen Ausgangspegel von 1 V ist der Klirrfaktor unter 2%.

EE 21 764 *für weitere Einzelheiten*

INTERNATIONAL RECTIFIER CO. (Great Britain) LTD

Hurst Green, Oxted, Surrey

SILIZIUM-FOTOZELLEN-LESER

(Abbildung Seite 317)

Diese fotoelektrischen Silizium-Lesematrizen haben eine Ansprechzeit, die bei 10 μs liegt und sind besonders für Rechner und Datenverarbeitungsgeräte bestimmt, für die schneller Nachweis des durch Lochkarten oder -streifen fallenden Lichtes erforderlich ist.

Jede Matrize besteht aus einer Mehrfachanordnung von einzeln unterteilten Siliziumzellen. Auf einen besonderen Abschnitt fallendes Licht wird nur in diesem Abschnitt einen Strom hervorrufen. Eine mit 1 000 Kerzen beleuchtete aktive Zellenfläche von 6,46 mm^2 erzeugt z.B. einen Strom von 300 μA.

Die Konstruktion ist so anpassungsfähig, dass auf Wunsch entweder Einzelzellen oder Mehrzellenanordnungen geliefert werden können. Sie sind durch unwesentliche Temperaturabhängigkeit, gute Betriebshaltbarkeit, gleichförmige Kennlinien der Zellen in einer Matrize und stossfeste Ausführung gekennzeichnet.

EE 21 765 *für weitere Einzelheiten*

HOCHSPANNUNGS-GLEICHRICHTER

(Abbildung Seite 317)

Diese Silizium-Gleichrichterpatronen sind für Spitzensperrspannungen von 600 V . . . 10 kV lieferbar. Die Gleichrichter sind keramisch eingekapselt und für Verwendung bis zu 27 500 m Höhe geeignet. Die Patronen sind für Gleichstromabgaben von 75 . . . 250 mA (bei +25°C) erhältlich und haben einen Temperaturarbeitsbereich von −55°C . . . +150°C. Der maximale Sperrstrom bei Spitzensperrnennspannung und 25°C ist 10 μA.

EE 21 766 *für weitere Einzelheiten*

KELVIN & HUGHES (INDUSTRIAL) LTD

Kelvin House, Wembley Park Drive, Wembley, Middlesex

DATENVERARBEITUNGSSYSTEM

Das zum ersten Mal ausgestellt System wurde für grösste Anpassungsfähigkeit konstruiert. Es kann eine unbegrenzte Zahl von Verfahrensvariablen mit 0,1% Genauigkeit und höheren Geschwindigkeiten ablesen und aufzeichnen als früher mit derartigen Anlagen möglich war.

Dem Eingang werden Gleichstrom oder -spannung zugeführt, so dass die Anwendungsmöglichkeiten der Anlage nur vom Vorhandensein geeigneter Umwandler abhängt. Solche Umwandler stehen für Druck-, Durchströmungs- und Dehnungsmessungen, Impuls- und Wechselstromabgabe zur Verfügung, während Thermoelemente der Temperaturmessung dienen. Kompensationseinheiten für Kaltverbindungen werden für jede Thermoelement-Eingangstype vorgesehen.

Die Information wird durchweg in Digitalform dargestellt. Folgende Vorrichtungen stehen für Integration in die Abtastfolge zur Verfügung: Überwachung der oberen und unteren Grenzwerte mit Alarmanzeige, Bereichänderungen, Entnullen und eingebaute Linearisierung, die für Thermoelement, Durchströmung usw. benötigt wird.

Ausdrucken erfolgt mittels elektrischer Schreibmaschinen auf leeren oder vorgedruckten Protokollblättern, wobei Punkte ausserhalb der Toleranzen rot gedruckt werden.

Ein Voltmeter misst die Verfahrensvariablen und zeigt die Ergebnisse in Digitalform an. Es enthält ein Reihenschaltergerät, Relais, einen zweistufigen Verstärker und Netzteil. Letzterer hat auch eine innerhalb 0,05% stabilisierte Spannung für Vergleichsmesszwecke. Spannungen werden der Reihe nach eingeschaltet und mit der Eingangsspannung verglichen, bis ein endgültiger unausgeglichener Wert innerhalb 0,1% erreicht wird.

Das System einschliesslich Umformer und Netzgeräten ist in einem Schrank untergebracht, in dem die Bausteine zwecks leichten Ausbaus und zur vereinfachten Wartung in Gestelle eingebaut sind.

EE 21 767 *für weitere Einzelheiten*

GEORGE KENT LTD

Luton, Bedfordshire

Quadratur-Unterdrücker

(Abbildung Seite 318)

Der Quadratur-Unterdrücker Type I wurde zur Verwendung mit elektronischen Geräten der Firma George Kent entwickelt, wo dieselben zum Messen von Wechselstromsignalen eingesetzt werden.

Der Unterdrücker ist ein kleines elektronisches Einsteckgerät, das durch Anwesenheit von Quadratur-Komponenten im Eingangssignal hervorgerufene Verstärkersättigung und Messfehler verhindert. Hohe Betriebssicherheit und Robustheit wird durch ausschliessliche Verwendung von Transistoren, Thermistoren und anderen Festkörperelementen erreicht. Der Hauptbaustein des Unterdrückers ist das patentierte KENT-Thermistor-Potentiometer, das aus einer gedruckten Schaltung auf einer Einsteckplatte besteht, auf die die hauptsächlichen Bauelemente montiert sind. Zusammen sind diese Elemente mit dem Verstärker, Ausgleichmotor und Messdraht eines herkömmlichen Servosystems vergleichbar.

Die oft in Wechselstromservosystemen wie z.B. beim Messen elektrolytischer Leitfähigkeit oder der Durchströmungsgeschwindigkeit von Flüssigkeiten auftretenden Quadraturkomponenten werden durch das Gerät wirksam unterdrückt. Diese Störsignale haben einen sehr schädlichen Einfluss, da im Eingang eines Wechselstromservosystems auftretende Quadraturkomponenten nicht durch Wechselstrom-Hauptgegenkopplung genullt werden können und ein nichtreduzierbares Fehlersignal hervorrufen. Dadurch können wieder Fehler im Signalausgang auftreten, oder der Verstärker kann teilweise oder sogar vollständig gesättigt werden und an Empfindlichkeit einbüssen.

Dieses Problem wird durch den Quadratur-Unterdrücker gelöst, der automatisch alle Quadraturkomponenten, beständiger Grösse oder nicht, im Eingangssignal annuliert, ohne in irgendwelcher Weise das Haupteingangssignal zu beeinträchtigen. Das vom Servoverstärker abgegebene Quadraturfehlersignal wird in einer Schaltung mit vier Ge-Transistoren kleiner Leistung verstärkt und demoduliert. Es wird dann dazu benutzt, die indirekte Heizung von zwei Thermistoren unterschiedlich zu steuern. Die beiden Heissleiterperlen bilden einen Spannungsteiler variabler Verstärkung, der ein genaues Quadratursignal steuert (durch 90° Phasenverschiebung von der Wechselstrom-Bezugsquelle erhalten), das zum Eingang des Servoverstärkers zurückgespeist wird und das Restsignal unterdrückt. Die Schaltung ist durchweg galvanisch gekoppelt, und ein dritter Thermistor gibt negative Gleichspannungs-Gegenkopplung vom Ausgang zum Eingang. Dadurch werden die Arbeitspunkte der vier Transistoren ohne Kondensatoren stabilisiert und die Ausgangskennlinien des Thermistor-Spannungsteilers im wesentlichen über einen Umgebungstemperaturbereich von 0°... 60°C konstant gehalten.

EE 21 768 *für weitere Einzelheiten*

Bezugsspannungsgerät

(Abbildung Seite 318)

Die Firma George Kent Ltd. hat eine Festkörper - Gleichspannungs - Bezugsquelle Type 3 entwickelt, die für Verwendung mit den von der Firma hergestellten elektronischen Kompensations-Schreibern, Anzeige und Steuergeräten bestimmt ist. Der kompakte Einsteckbaustein arbeitet direkt vom Wechselstromnetz und gibt eine äusserst stabile, restwellenfreie Ausgangsleistung von 5,0 mA bei 5 V für den Messdraht des Messgerätkreises.

Zwei Zener-Dioden-Stufen in Kaskade setzen den Einfluss der üblichen Netzschwankungen mindestens um einen Faktor von 200 herab. Temperaturkompensation für jede einzelne dieser Stufen ergibt eine Ausgangsspannung, die von Schwankungen in der Umgebungstemperatur zwischen 10°... 70°C weitgehend unabhängig ist. Ein dem Gerät eigener Vorteil ist, dass es mit nur wenig reduzierter Genauigkeit arbeitet, selbst wenn die Netzspannung drastisch bis zu 30% der Nennspannung herabgesetzt wird.

Der Gleichstrom-Ausgang ist von Erde isoliert, und ein besonders abgeschirmter und abgeglichener Netzübertrager isoliert ihn vom Netz. In gewissen Messkreisen mit endlicher Impedanz sind diese Eigenschaften sehr wertvoll.

EE 21 769 *für weitere Einzelheiten*

KYNMORE ENGINEERING CO LTD

19 Buckingham Street, London W.C.2

Hochempfindliches Relais

(Abbildung Seite 318)

Diese von Siemens und Halske hergestellten Relais können überall dort benutzt werden, wo Signale der pW-Grössenordnung zu überwachen sind, wo sie geregelte Verfahren einzuleiten oder anzuhalten haben und wo eine zuverlässige, zwangsschlüssige Kontaktarbeitsweise innerhalb enger eingestellter Grenzen erforderlich ist.

Die Arbeitsweise ist wie folgt: Wenn der eingestellte Wert erreicht ist, wird der Kontakt zwangsschlüssig mit nur äusserst geringem Kraftbedarf betätigt. Das beruht auf der Zusammenwirkung des Blattfeder-Tauchspulankers des Kernmagneten mit einem magnetischen Kontaktdruckverstärker. Selbst wenn die Eingangssignale sehr langsam ansteigen, arbeiten die Kontakte mit Schnappaktion. Der Zeiger des Blattfederankers ist mit einem kurzen Eisendrahtstück versehen. Im vorgewählten Arbeitsbereich bildet ein kleiner Dauermagnet einen Anschlag an der einstellbaren Ablenkungsbegrenzung. Wenn das Eingangssignal ansteigt, bewegt sich der Zeiger zum magnetischen Kontaktdruckverstärker, bis die Blattfederspannung durch den Magnetkernfluss balanziert wird. An diesem Punkt verursacht der geringste Signalanstieg das Herüberspringen des Zeigers zum magnetischen Anschlag, wo ein Kontaktdruck von ungefähr 2 g aufrechterhalten wird. Da die Kontaktoberflächen des Zeigers und des magnetischen Anschlags vergoldet sind, wird selbst nach langer Lagerung und Betriebsruhe guter Kontakt gemacht. Der Zeigerkontakt erregt ein Hilfsrelais, dessen Spule mit 24 V Gleichspannung gespeist wird. Die Kontakte des Hilfsrelais sind für Betätigung industrieller Kontaktgeber oder Alarmvorrichtungen berechnet. Auf diese Weise können verschiedene getrennte Kontaktgeber gesteuert werden. Das Hilfsrelais kann mit verschiedenen Federsatzkombinationen ausgerüstet werden. Ein Kontakt wird als Haltekontakt benutzt und liegt parallel zum Relaiszeigerkontakt. Auf diese Weise wird der Kontakt aufrechterhalten, bis das Relais durch Bedienung eines Handdruckknopfes ausgelöst wird. Der Rückholknopf erfüllt eine Doppelfunktion, da er zuerst eine Abfallspule erregt, deren Anker den Zeigerkontakt unterbricht, der stromlos ist, da er parallel zum Haltekontakt des Hilfsrelais liegt. Erst dann wird der Spulenkreis des Hilfsrelais unterbrochen. Daher kann an den Zeigerkontakten keine Funkenbildung stattfinden. Wenn bei Betätigung des Rückholknopfes das Signal noch anhält, erregt das Relais sofort wieder das Hilfsrelais.

Diese Relais haben eine aussenliegende Einstellvorrichtung für den Arbeitspunkt und eine linear unterteilte Skala. Sie sind für eine Reihe von Strom- und Spannungsbemessungen sowie eine Höchstempfindlichkeit von 10^{-10}W lieferbar.

EE 21 770 *für weitere Einzelheiten*

J. LANGHAM THOMPSON LTD

176 High Road, Bushey Heath, Hertfordshire

Druckgeber

(Abbildung Seite 319)

Durch seinen Frequenzgang ist dieser Geber, der zur Messung sowohl gleichbleibender als auch schnell schwankender Drucke entwickelt wurde, besonders zur Untersuchung von Einschwingvorgängen in hydraulischen und pneumatischen Systemen geeignet.

Modelle sind für Bereiche 0 ... 141 atü und 0 ... 1 400 atü mit hohem Bruchdruck bei Überlastung lieferbar.

Das druckempfindliche Element besteht aus einem Berylliumkupferzylinder, der z.T. ausgebohrt ist und eine Röhre mit geschlossenem Ende bildet.

Auf dem ausgebohrten Teil sind zwei aktive induktionsfreie Dehnungsstreifen-Wicklungen und auf dem geschlossenen Teil zwei Temperaturausgleichwicklungen aufgebracht. Diese werden als vier Arme einer ausgeglichenen Brückenschaltung verbunden.

Druck innerhalb des ausgebohrten Zylinderteils ruft eine Ausdehnung des Zylinders und damit Widerstandsänderungen in den aktiven Wicklungen hervor; wird eine Gleich- oder Wechselspannung an die Brücke gelegt, so wird dadurch ein dem Druck proportionales Signal hervorgerufen. Die volle Ausgangsspannung ist 1,5 mV für jedes angelegte Volt. Die Betriebshöchstspannung ist 30 V. Die Gesamtgenauigkeit einschliesslich Linearität und Hysteresis ist besser als 0,5%.

EE 21 771 *für weitere Einzelheiten*

LAURENCE, SCOTT & ELECTROMOTORS LTD

Mansfield House, 376 Strand, London W.C.2

Regler für konstante Drehzahl

Der Regler kann die Drehzahl eines rotierenden Schaftes mit einer Genauigkeit konstant halten, die für den vollen Nennbereich des Antriebsmotors besser als ein Teil in dreitausend ist. Eine der zwei lieferbaren Typen hat einen kleinen Induktionsmotor als Antriebsmaschine, die andere einen Gleichstrom-Dauermagnetmotor. Der Regler gibt ein von der Geschwindigkeit abhängiges Signal ab, das von einer Abtastspule hervorgerufen wird, deren Polschuhe nahe dem Umkreis eines direkt mit dem Antriebsschaft gekuppelten Zahnkranzes liegen. Für die berechnete konstante Geschwindigkeit ist die Frequenz des abgegebenen Signals 50 kHz. Dieses Signal wird verstärkt und durch ein Quarzfilter gespeist, dessen Resonanz gerade über 50 kHz liegt. Der Ausgang dieses Filters wird gleichgerichtet und steuert schliesslich eine Kathodenverstärkerstufe. Bei Gleichstromantrieb liegt der Motor direkt im Kathodenkreis, aber in der Wechselstromausführung wird der Abschluss des Kathodenverstärkers durch die Steuerwicklung eines magnetischen Verstärkers gebildet, der seinerseits die Motorengeschwindigkeit regelt.

EE 21 772 *für weitere Einzelheiten*

MARCONI INSTRUMENTS LTD

St. Albans, Hertfordshire

10MHz-Zähler

(Abbildung Seite 319)

Der 10MHz elektronische Zähler Type TF 1345 ist ein Schnellzähler/Zeitmesser mit eingebautem Präzisions-Frequenznormal. Die Anzeige erfolgt mittels Glimmröhren in 8-Dekaden-Digitalform. Das Gerät zählt bis zu 10 Millionen pro Sekunde, misst Frequenzen von 10 Hz ... 10 MHz und Wellenformenperioden bis zu 100 kHz. Einsteckzubehör erweitert den Frequenzbereich bis zu 220 MHz, erlaubt Zeitmessungen bis zu 1 μs herunter und erhöht die Empfindlichkeit zu 10 mV. Die Zeitsteuerung der Anzeige geschieht entweder manuell oder kontinuierlich variabel von 0,1 ... 10 s mit automatischer und sich wiederholender Nullstellung. Das Gerät ist für Prüfbank oder Gestelleinbau lieferbar.

EE 21 773 *für weitere Einzelheiten*

Doppelspur-Oszillograf

(Abbildung Seite 319)

Modell TF 1331 ist ein gleichstromgekoppelter Doppelspur-Oszillograf mit direkt ablesbarer Zeit- und Spannungseichung. Elektronisches Schalten des Strahls ermöglicht Anzeige zweier unabhängiger Signale entweder mit abwechselnder Abtastung oder 100 kHz-Schaltgeschwindigkeit. Jeder der Eingänge kann auch getrennt ohne Schalten dargestellt werden. Der Frequenzgang liegt zwischen Gleichstrom und 15 MHz, und die Empfindlichkeit ist in sieben Stufen zwischen 50 mV/cm... 50 V/cm einstellbar. Die Kippgeschwindigkeit ist in 15 Bereichen von 1 s/cm ... 0,1 μs/cm regelbar und kann durch Benutzung der X-Ausdehung bis auf 0,02 μs/cm erhöht werden. Triggerung kann von innen oder aussen oder auch von einer internen Netzfrequenzquelle erfolgen. Ein 10MΩ-Eingangsmesskopf ist als wahlweises Zubehör lieferbar. Eine Einstrahlausführung TF 1330 wird auch hergestellt.

EE 21 774 *für weitere Einzelheiten*

METRIX INSTRUMENTS LTD

54 Victoria Road, Surbiton, Surrey

Verfahrensregler

(Abbildung Seite 320)

Der transistorisierte Metrawatt-Regler bietet eine preisgünstige und doch genaue Methode zur Regelung von Temperatur, Wasserstoffionenkonzentration, Leitfähigkeit, Leistung und anderen Variablen, deren Funktion in elektrische Spannung umgeformt werden kann.

Das Reglersystem benutzt ein oder mehrere HF-Oszillatorkreise, deren Koppelspulen auf einstellbaren Führungsschlitten sitzen, die für jeden gewünschten Regelwert der Instrumentskala eingestellt werden können. Ein kleiner Flügel am Arm des Drehspulmessgerätes wird in den Luftspalt zwischen den Oszillatorspulen eingeführt, sowie der Zeiger den Wert der vorbestimmten Regelstellung erreicht. Die Oszillatorschaltung betätigt Quecksilber- oder Hochleistungs-Kontaktgeber, die für die endgültige Regelung erforderlich sein mögen. Da das Messgerät keine mechanischen Kontakte hat, kann es äusserst sensitiv sein, um hohe Reglergenauigkeit (0,2%) zu geben und schnell anzusprechen.

Ein einfaches Wärmegegenkopplungssystem steht auf Wunsch zur Verfügung, um das Gerät in einen proportionalen Regler umzuwandeln; der Einfluss der Wärmeträgheit auf das Reglersystem ist dadurch fast vollständig ausgeschaltet.

Das völlig transistorisierte Gerät hat eine Schutzschaltung, die dafür sorgt, dass die Stromversorgung abgeschaltet wird, sowie ein Fehler im Thermoelement oder Geber oder im Gerät selbst auftritt.

Der Einfluss von Netzschwankungen ist nur 0,1% für Spannungsänderungen von 15%.

EE 21 775 *für weitere Einzelheiten*

MUIRHEAD & CO. LTD

Beckenham, Kent

Präzisions-Dekaden-Effektivvoltmeter

(Abbildung Seite 320)

Das Muirhead-Wigan D 930A ist ein sehr genaues Voltmeter mit grossem Spannungsbereich (1 mV...300 V) und grossem Frequenzbereich (5 Hz ... 100 kHz). Das Voltmeter misst die echte Effektivspannung sinusförmiger und nichtsinusförmiger Wellenformen.

Die Spannung wird direkt von vier Dekadenskalen abgelesen, die mit „Einer" (1...10), „Zehner" (0...10), „Hunderter" (0...10) und „Tausender" (0...10) bezeichnet sind. Ein sechsstufiger Bereichschalter gibt die Erweiterungsfaktoren von 0,001...100.

Es ist daher immer möglich, mit 0,1% Genauigkeit abzulesen und bis zu 0,025% zu integrieren.

Der konstante Eingangswiderstand von 100 kΩ ermöglicht das Messen niedriger Spannungen mit geringem Leistungsbedarf. Selbst bei 300 V ist die Eingangsverlustleistung unter 1 W.

Das folgende Richtlinien-Pflichtblatt wird für das Gerät angegeben:

Spannungs-Messgenauigkeit:

30 Hz...5 kHz	60 mV...10 V ±0,05% (ausser für Netzspannung); 1 mV ... 300 V ±0,1% oder 30 μV (jeweils der grössere Wert).
20 Hz...20 kHz	1 mV ... 300 V ± 0,2% oder 30 μV (jeweils der grössere Wert).
5 Hz...100 kHz	10 mV...300 V ±0,25% bis zu 50 kHz abfallend zu ±1% für 100 kHz (mit Hilfe eines stabilen Fremdoszillators wie z.B. des D-890-A/2 zur Eichung der Testfrequenz).

Genauigkeit der Spannungsverhältnis-Messungen für jeden der Bereiche:

Bereiche 100 V...300 V	±0,03%...0,15%
10 ...100 V	bis zu 50 kHz
10 V...100 V	(entsprechend
usw bis	Verhältnissen und
10 mV...100 mV	Spannungen, jedoch nicht schlechter als ±0,2% bei Netzspannung).
1 mV...10 mV	±0,05%...0,15% bis zu 10 kHz (gemäss Verhältnissen, jedoch nicht schlechter als ±1% bei Netzfrequenz).

Eingangsimpedanz

100 parallel mit 200 pF (max)

Höchstes Spitzen/effektiv Verhältnis der Eingangswellenform

3 : 1

EE 21 776 *für weitere Einzelheiten*

SHORT BROTHERS & HARLAND LTD

Seaplane Works, Queens Island, Belfast, Ireland

ANALOGRECHNER

Der neue "Simlac Minor" Universalrechner wird alle diejenigen interessieren, für die Short's 18-Verstärker-Universalrechner zu klein ist und die für die Kapazität oder Genauigkeit des vor kurzem herausgebrachten grossen Simlac-Rechners keinen Bedarf haben.

Die Maschine hat bis zu 32 Verstärker und bis zu loo passive Elemente. Der Simlac Minor ist ebenso wie das grössere Modell für Gestellmontage konstruiert und in zwei genormten 19 Zoll (48,26 cm) Gestellen untergebracht. Im linken befinden sich die Rechenelemente und im rechten die Instrumentierung, auswechselbare Stecker-Schalttafeln, Netzgeräte und drei nichtlineare Sonderbausteine.

Um Höchstausnutzung des Raumes zu erreichen, wird auf Ofenstabilisierung verzichtet. Trotzdem ist die Gesamtgenauigkeit der Rechenelemente infolge durchdachter Konstruktion und sorgfältiger Wahl der Einzelteile 0,1%. Die Verstärker und passiven Elemente sind Simlac-Standardausrüstung mit Ausnahme der "Halten"-Schaltung, die ein wahlweiser Zusatz-Baustein ist.

EE 21 777 *für weitere Einzelheiten*

NF-OSZILLATOR UND REGLERSYSTEM-ANALYSATOR

(*Abbildung Seite 320*)

Dieser NF-Oszillator wurde für niederfrequentes Prüfen entwickelt, für das äusserst geringer Gleichstromgehalt und genau eingehaltene Amplituden sehr wichtig sind. Die Verstärker haben kontinuierliche Driftkorrektion, so dass der Gleichstromgehalt der Ausgangsspannung im allgemeinen beträchtlich unter looμ liegt.

Der Reglersystemanalysator ist einzigartig, da er auf sehr niedrigen Frequenzen, z.B. noch bei einer Periode in loo s, erfolgreich arbeitet. Der Bereich erstreckt sich bis zu 1 kHz und das Gerät ist für die Untersuchung sehr vieler Systeme geeignet.

Der NF-Oszillator ist in den Analysator eingebaut und gibt selbst bei der niedrigsten Frequenz von 0,01 Hz unveränderliche Anzeigen der phasenmässigen und Quadraturkomponenten. Der Gleichspannungspegel ist immer geringer als, 0,005% und kann durch sorgfältiges Abgleichen des driftkorrigierten Verstärkers bei 0,0005% gehalten werden.

EE 21 778 *für weitere Einzelheiten*

Zusammenfassung der wichtigsten Beiträge

Ein Spektralanalysator für Videofrequenzen

von A. L. Whitwell und N. Williams

Zusammenfassung des Beitrages auf Seite 268-274

In diesem Beitrag wird ein Universal-Spektralanalysator für Frequenzen von 500 Hz . . . 95 kHz beschrieben. Die Konstruktionsgrundsätze für Geräte dieser Art sind nicht neu, rufen jedoch wachsendes Interesse für Einsatz als Universal-Laborgerät oder in der Betriebskontrolle hervor. Sie können für eine grosse Anzahl von Messungen Verwendung finden und für hohe Empfindlichkeit entwickelt werden, da durch nachfolgende Filterwirkung die effektive Rauschzahl der Eingangsstufe reduziert wird. Das beschriebene Gerät hat z.B. eine Empfindlichkeitsschwelle von ungefähr 0,2 μV in einem 70 Hz-Band und einen dynamischen Umfang von 50 dB.

Graphische Analyse und Konstruktion eines Sperrschwingers

von D. Chambers

Zusammenfassung des Beitrages auf Seite 275-279

In diesem Beitrag wird die Arbeitsweise des Sperrschwingers beschrieben und die Wahl aller Schaltungselemente eingehend untersucht. Spannungs- und Stromwellenformen werden nach der Schritt-für-Schritt-Methode aufgebaut und die für bestimmte Verwendungszwecke geeignetste Röhre, Sperrkondensator, Transformatorkern und Übersetzung bestimmt.

Direktes Messen für Transformator-Analogrechner

von J. F. Young

Zusammenfassung des Beitrages auf Seite 280-288

Die Phasen von Rechnerspannungen werden verschoben und in Verstärkern zusammenaddiert; die Ergebnisse werden Regelverstärkern zugeführt. Diese speisen Endverstärker mit Stromgegenkopplung, die Ampere- und Wattmeter steuern. Phasenschiebernetzwerke konstanten Widerstands werden zusammen mit einem Wattmeter benutzt, um induktionsfreie Voltampere zu messen. Als Phasenmesser wird ein geeichter Phasenschieber verwandt.

Ein Netzwerk konstanten Widerstands, das die Eigenschaften eines Universal-Nebenschlusses hat, in dem das Nebenschlussverhältnis durch Dekadenskalen bestimmt wird von E. R. Wigan

Zusammenfassung des Beitrages auf Seite 289-293

Durch eine einfache Abwandlung kann das überbrückte T-Netzwerk mit konstantem Widerstand in einen Universal-Nebenschluss mit konstantem Widerstand verwandelt werden, der nach dem Dezimalsystem einstellbar ist; ein 3-Dekaden-1000 Ω-Nebenschluss wird beschrieben, der ein maximales Nebenschlussverhältnis von 11,111/1,0 hat und in Stufen von 0,001 einstellbar ist. Zwischen diesen Stufen kann eine kontinuierliche Interpolation vorgenommen werden, deren Genauigkeit diskutiert wird.

Stroboskop-Impulsfolgegeber zur Untersuchung der Wirbelströmung in Flüssigkeiten von R. E. King

Zusammenfassung des Beitrages auf Seite 294-295

Eine Kathodenselektorröhre wird zur Erzeugung einer Impulsreihe in vorbestimmter Folge über einen breiten Taktbereich benutzt. Der Folgegeberausgang triggert ein Stroboskop, das fotografischen Aufschluss über Richtung und örtliche Geschwindigkeit der Flüssigkeitsverwirbelung gibt.

Ein verbessertes Aufzeichnungsgerät für Blutsauerstoff von A. W. Melville, D. H. Smith und J. B. Cornwall

Zusammenfassung des Beitrages auf Seite 296-300

Die Theorie der Blutsauerstoffmessung im Ohr wird eingehend untersucht. Ein Gerät wird beschrieben, das kontinuierlich die theoretischen Gleichungen löst und die absolute Sauerstoffsättigung der Pulsader aufzeichnet.

Die bisher benutzten Methoden arbeiteten mit einem Signal, das dem Logarithmus der auf die Ohrmuschel-Fotozellen auffallenden Lichtstärke proportional war. Sie stellten sich als unzuverlässig heraus und wurden ersetzt.

Die neue Schaltung hat auch andere Vorteile, die zu hoher Stabilität und Reproduzierbarkeit führen.

Das Gerät wurde besonders für anästhetisierendes Personal zum Einsatz bei langdauernder Thorakotomie entwickelt, kann jedoch überall Verwendung finden, wo kontinuierliche Aufzeichnung der Sauerstoffsättigung vorteilhaft ist.

Ein mit Flächentransistoren bestücktes physiologisches Reizstromgerät von W. T. Catton und L. Molyneux

Zusammenfassung des Beitrages auf Seite 301-302

Nach zweijähriger Erfahrung mit einem älteren Reizstromgerät mit einem Spitzentransistor im Physiologieunterricht wurde unter Berücksichtigung des Veraltens des Gerätes die Entwicklung einer neuen Schaltung mit Flächentransistoren in Angriff genommen. Die Anordnung der Bedienungselemente auf der Frontplatte wird durch die neue Schaltung nicht geändert, die jedoch eine bessere Impulsform abgibt und eine niedrigere Ausgangsimpedanz hat. Ausserdem ist die erforderliche Betriebsspannung niedriger. Die Anwendungsmöglichkeiten bleiben dieselben wie beim älteren Modell; die Ausgangsspannung ist von 0 . . . 25 V und die Frequenz von 2 . . . 100 Hz variabel; die konstante Impulsdauer ist ungefähr 1 ms.

Ein Impulslängenmodulator von D. C. Brown und J. E. Baughen

Zusammenfassung des Beitrages auf Seite 302-303

In dem beschriebenen Impulslängenmodulator findet der Löcherspeichereffekt eines Flächentransistors Anwendung. Bei Benutzung eines mit 10 kHz abgetasteten Signals wurde eine Empfindlichkeit von 20 μs/V erzielt.

Ein spannungsmodulierter Impulsgeber mit variabler Frequenz von E. J. C. Fowell und A. Cowley

Zusammenfassung des Beitrages auf Seite 304-306

Der Beitrag beschreibt eine Impulsgeberschaltung, deren Ausgangsfrequenz eine geradlinige Funktion der Eingangsspannung ist.

Die Schaltung benutzt schirmgittergekoppeltes doppeltes Phantastronsystem mit entsprechenden Anderungen zur Verbesserung der Übertragungsgenauigkeit für niedrige Ausgangsfrequenzen.

Die Ausgangsfrequenz ist in vier Bereiche zwischen 5 . . . 2500 Hz unterteilt.

Die vollständige Schaltung und erzielte Leistung werden angegeben.